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The Marketing Guy Said No.
The Test People Said No.
The Applications Engineer Said No.
The Designers Said Maybe.

But Shep Said Yes.

It doesn't always make sense to give customers more than they think they need. After all, if testing for PCS and wireless happens right now at 0.8 MHz to 1.9 GHz, it makes sense to do an amp that tests there. No unwanted extras. No stress on anyone's budget. At least that's what the experts say.

But here's wisdom: things are changing so fast in the wireless world, there's little logic in manufacturing or buying anything that meets only immediate needs. Only a brain donor would bank on an amp with a really narrow band. A more righteous approach gives you room at the top— extra band you're gonna be glad you have in the test room, where performance counts. And tomorrow, when the rules are rewritten.

Shep knows this. He knew it all along. That's why he insisted AR "S" Series amplifiers each offer bandwidth from 0.8 to 4.2 GHz and power from 1 to 100 watts, in increments just right for your application. With low harmonics. And other extras, like IEEE 488 and RS 232 interfaces for remote operation.

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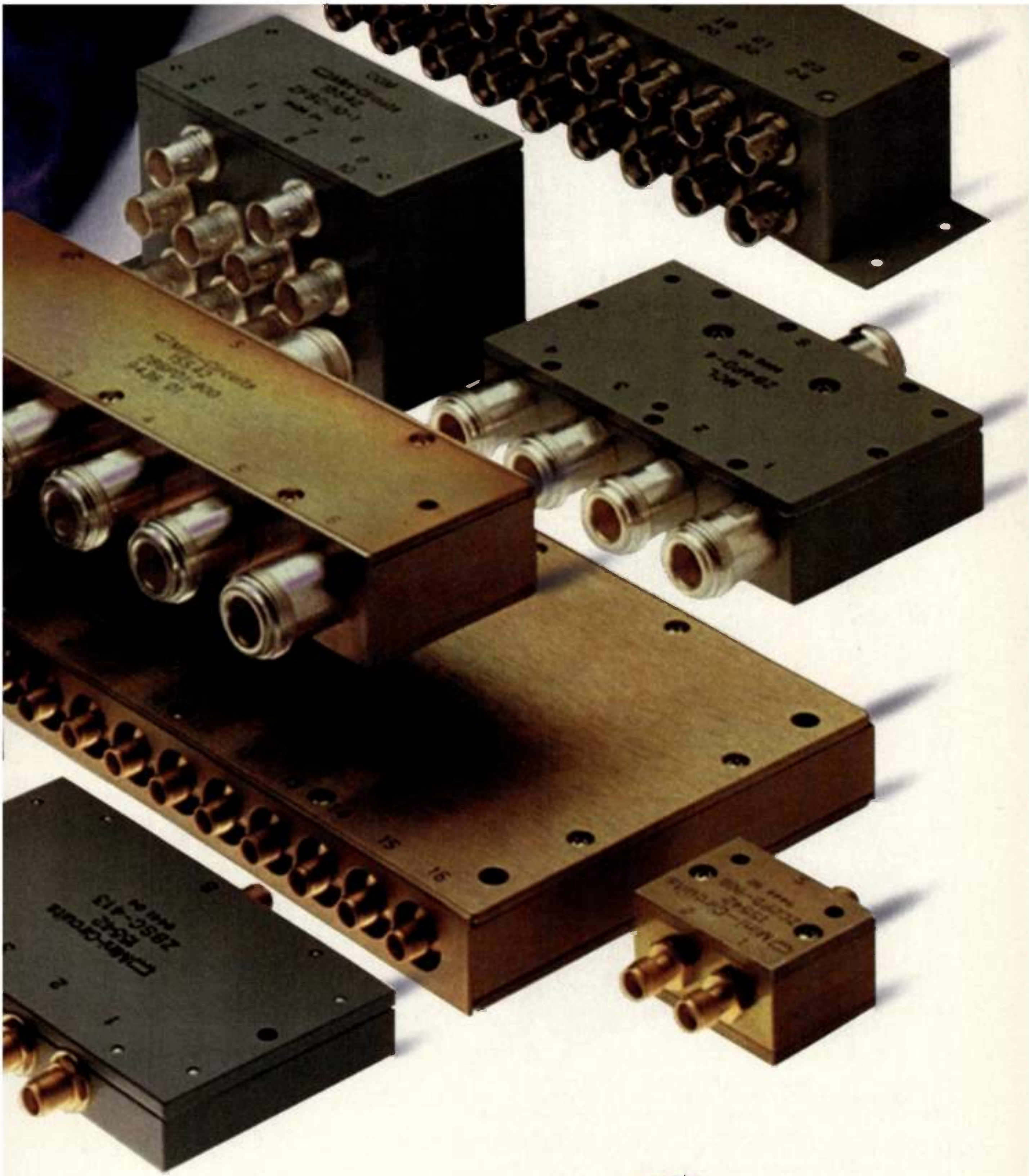

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Flatness, dB	±0.4	±0.4	±0.4	±0.4
Power Output, dBm (@ 1dB comp.)	30	30	30	30
Dynamic Range				
IP2, dBm	88	88	88	88
IP3, dBm	47	48	47	48
NF, dB	3.5	3.5	3.5	3.5

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in this issue

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RF editorial

What's old is new again

By Don Bishop
Editorial Director

Ask someone to define electronics, and you probably will be told that electronics is the use of materials and techniques to direct electron flow for various purposes, such as control, calculation, measurement and communication. Mechanical methods preceeded many of today's electronic applications. Hybrid, electromechanical methods combine the two, usually when certain mechanical methods remain less expensive than counterpart electronic methods.

Videocassette recorders (VCRs), as electromechanical devices, remain a ubiquitous example in consumer electronics. So do audio cassette recorders. Time is on the side of electronic replacements as digital converters and computer memory become less expensive.

Decades ago, the Columbia Broadcasting System (CBS) proposed an electromechanical method for color television—maybe not entirely expecting it to gain government approval, although it did for a while. It wasn't long before it was replaced by all-electronic television. Almost everyone wants the "new" electronics to supplant "old" mechanical methods wherever possible and as soon as it is affordable.

In this context, it's fascinating to see the emergence of micro-electromechanical systems (MEMS). Combining computer chips with mechanical systems that are so small that "microscopic" fails to describe their size, MEMS can oscillate at high frequencies, as can electrons, supporting functions at a level of miniaturization that purely electronic circuits may not. How much smaller a wireless telephone needs to be might be debatable, but adding multiple power-efficient features within a tiny wireless telephone package stimulates the competitive instincts. It's been predicted, by the way, that MEMS will reduce wireless telephones to wrist-watch size. Exactly what will be the configuration of a really tiny wireless telephone? My guess is a combination throat microphone and earpiece similar to a hearing aid.

Going back to the turn of the century an electromechanical method—the high-frequency alternator—provided RF for long-distance radio communications. The motors, nozzles and valves that are part of some MEMS someday may be used to generate RF. If and when it happens, the transition will include some devices that cling to the "old" electronics methods until MEMS prices fall!

Virtual networks

On another subject, have you noticed the proliferation of cellular and PCS base station antennas? I've seen towers that look as though they have antennas for four carriers. Some cities have six carriers, but I'm not sure I've seen that many system antennas on one tower yet. Makes me wonder how long it will be before we see "virtual networks."

System operators increasingly focus on "selling minutes." They say they are growing less interested in any part of the business besides the promotion and sale of airtime. That means outsourcing site management, network management, billing, fraud control, customer service and other functions. Maybe the time is coming for one of two mega carriers to consolidate everyone else's systems and to offer "virtual network" airtime services to what would become "virtual carriers."

With virtually (sorry) no difference between the operational peformance of each virtual carrier, the key to each one's success will be its management's selection of outsource providers. As wireless systems age, specialization increases. *RF*

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PV 810C VCO	810-850	-103 dBc/Hz	5V, <25mA
PSF 2510 Synthesizer fixed Freq	2510	-105 dBc/Hz	5V, <40mA
PSB 1880 Synthesizer	1885-1945	-101 dBc/Hz	5V, <25mA

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INFO/CARD 49

Editorial Forum

By Gregg V. Miller, Technical Editor

The honeymoon is over

Ever since I have been involved with the cellular industry, it has been the view of the cellular people that the cellular phone is not designed to replace the landline phone, but to complement it. Cellular can provide features that landline cannot and visa versa. And, this stance has been working well for both the cellular and landline industries.

However, recently there has been an advertisement from AT&T Wireless that says they want your cellular phone to be your only phone.

Excuse me?

This changes the whole ball game for everyone involved with cellular. For instance, in *Wireless Review*, a sister publication of *RF Design*, the editors and publisher tried to spend a week using their cellular phone as their only phone. While there were the normal advantages with using a cellular phone (anywhere connectivity, making a phone call and a fax at the same time, having people reach you after business hours), they were not happy with the performance to say the least.

Using the phone constantly brought out the disadvantages of cellular. The dropped calls, no service areas and limited battery life that cellular users usually can live with now became serious problems. Instead of looking at the cellular phone as an add-on benefit, they saw it as an inconvenience when compared to landline.

So once again, the marketing departments of the world have put their foot in their mouths. In an obvious effort to increase market penetration by touting the advantages of cellular, they are actually bringing the pitfalls of cellular to the forefront. And when customer service representatives start to hear more and more about these problems, who do you think will get the blame?

Congratulations, RF engineers. RF

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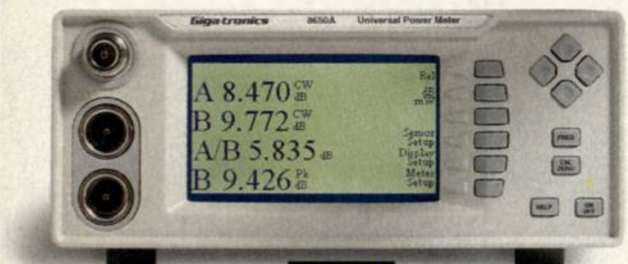
Choose The Best Power Meter For

When it comes to choosing a power meter for fast, accurate measurements of complex, modulated communications signals, your choice is simple.

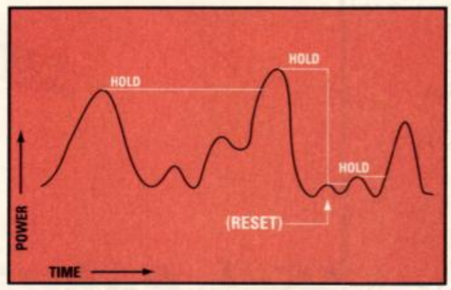


NEW

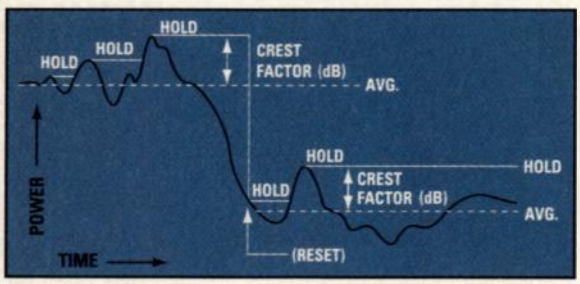
If you need to measure the CW, peak and average power of GSM, TDMA and CDMA signals with features such as time gating and crest factor measurement, choose the Giga-tronics 8540C Series Universal Power Meter. Since 1993, it has been the standard for communications testing.

If you need these capabilities and features plus even greater speed, the bandwidth to measure third-generation CDMA signals over a wide range, and statistical power measurement analysis to evaluate communications system efficiency, choose the new Giga-tronics 8650A Series Universal Power Meter.

Both the Giga-tronics 8540C and new 8650A Series Universal Power Meters feature blazing speed and wide dynamic range for CW power measurements. But that's just the start.

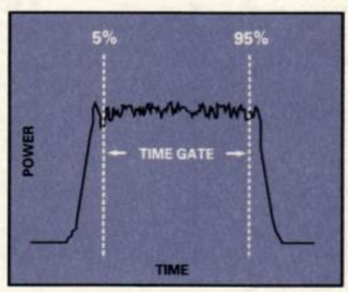


Both meters let you measure the peak power directly, and you can even measure the maximum peak power level using our peak hold feature.



Both meters automatically measure average power during the 'on' period of amplitude modulated TDMA signals. And you can directly measure the crest factor of the burst signal.

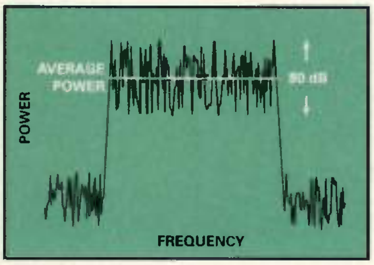
Both meters have a time gating feature that lets you set a measurement start and duration time within the burst portion of a signal. For example, you can use this feature to accurately measure average power during 5% to 95% of the burst duration — the 'useful' portion of the burst defined in the GSM specification.



Both meters have the wide single sensor dynamic range required for CDMA signal open-loop tests, and the speed you need to quickly measure the 1 dB steps over a 48 dB range during closed-loop tests. And both meters feature random, as well as uniform, sampling to minimize the aliasing effects of modulated signals for faster average power measurement speeds.

For More Information, Just Click www.gigatronics.com

Your Communications Test Needs



The new 8650A Universal Power Meter has a 10 MHz bandwidth to test third-generation CDMA signals over a 80 dB range with a single sensor. A 20 MHz sampling rate lets you capture up to 26,000 readings per second. And statistical power measurement analysis lets you evaluate communications system efficiency.

Giga-tronics power meters achieve this unmatched level of performance because meter architecture provides for a broad choice of sensors. Just by changing a sensor, you can measure the CW, peak, or average power of TDMA, GSM and CDMA signals faster, more accurately, and over a wider range.

NEW

Check the chart to see how the 8540C and 8650A compare with meters from other leading manufacturers. Then contact Giga-tronics for more information on the best power meter for your communications test needs.

Features and Specifications	NEW Giga-tronics 8650A	Giga-tronics 8540C	Anritsu ML2430A	HP EPM-440A
GPIB CW Measurement Speed (rdgs/s)				
Normal Mode	>300	>30	50–150	200
Swift Mode	>1,750	>175	150–250 with display off	NA
Buffered Mode	>26,000	>2,600	100–500	NA
GPIB Modulated Measurement Speed (rdgs/s)				
Normal Mode	>150	15	10–15	NA
Swift Mode	>300	30	<50 with display off	NA
Buffered Mode	>800	NA	NA	NA
Uniform Sample Rate	20 MHz	10 kHz	35 kHz	<10 kHz
Random Sample Rate	2.5–5 MHz	7–10 kHz	NA	NA
Maximum Diode Sensor Video Bandwidth	20 MHz	1.5 MHz	250 kHz	≈220 Hz
Maximum CW Single Sensor Dynamic Range	90 dB	90 dB	90 dB	90 dB
Maximum Single Sensor Dynamic Range				
TDMA/GSM	60 dB	87 dB	90 dB	50 dB
CDMA (IS-95)	80 dB	75 dB	50 dB	50 dB
CDMA (10 MHz bandwidth)	80 dB	50 dB	50 dB	50 dB
Maximum Peak Power Sensor Rise Time	100 ns	100 ns	10 μs	NA
Automatic Time Gate Setting	Yes	Yes	No	No
Direct Crest Factor Measurement	Yes	Yes	No	No
Statistical Power Measurement Analysis	Yes	No	No	No

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RF calendar

February 23–25 **Wireless Symposium**—*San Jose, CA.* Information: Debbie Cameron, Wireless Symposium Exhibition, 611 Route 46W, Hasbrouck Height, NJ 07604. Tel. 1-888-wir-ereg; e-mail *wirelessreg@penton.com.*

March 19–21 **Embedded Systems Conference**—*Chicago.* Information: Douglas St. John, Miller Freeman. Tel. 415-538-3848 or 888-239-5563; e-mail *esc@mfi.com.*

23–24 **10th Annual Digital Engineering Conference—The Consumer Electronics Future**—*Hasbrouck Heights, NJ..* Information: Consumer Electronics Manufacturers Association, 2500 Wilson Blvd., Arlington, VA, 2201-3834. Tel. 703-907-7660; e-mail *engcema@eia.org.*

April 19–21 **1999 International Conference on Gallium Arsenide Manufacturing Technology**—*Vancouver.* Information: Network Device, 1230 Bordeaux Drive, Sunnyvale, CA. Tel 408-734-9888; Fax 408-734-9889; e-mail *dday@network-device.com*; Web site **www.GaAsManTech.org**.

26–28 **DSP World Spring Design Conference**—*Santa Clara, CA.* Information: Liz Austin Miller Freeman. Tel. 415-538-3848 or 888 239-5563; e-mail *dspworld@mfi.com.*

27–30 **RF Design Seminar Series**—*Las Vegas.* Information: Intertec Trade Shows and Conferences, 6300 S. Syracuse Way, Suite 650, Englewood, CO, 80111. Tel. 800-288-8606 or 303-220-0600; Fax 303-770-0253.

27–30 **Base Station Workshops**—*Las Vegas.* Information: Intertec Trade Shows and Conferences, 6300 S. Syracuse Way, Suite 650, Englewood, CO, 80111. Tel. 800-288-8606 or 303-220-0600; Fax 303-770-0253.

28–30 **IWCE**—*Las Vegas.* Information: Intertec Trade Shows and Conferences, 6300 S. Syracuse Way, Suite 650, Englewood, CO, 80111. Tel. 800-288-8606 or 303-220-0600; Fax 303-770-0253.

May 24–26 **Markets and Applications for High Frequency Magnetic Materials 99**—*Santa Clara, CA.* Information: Karen Zacharias, Conference Coordinator, Gorham/Intertech Conferences, 411 US Route One, Portland, ME, 04105. Tel 207-781-9800; Fax 207-781-2150; e-mail *info@intertechusa.com.*

RF courses

UCLA Extension— *Project Management and Principles*— Feb 16–19; *Microwaves and Wireless Simplified*—Feb 17–19; *HBT IC Technology for Communications Applications*—Feb 18–20, Los Angeles; Information: UCLA Extension, Department of Engineering, Information Systems and Technical Management, Short Courses, 10995 Le Conte Ave., Suite 542, Los Angeles, CA, 90024-2883. Tel. 310-825-3858; Fax 310-206-2815; e-mail *mhennessi@unex.ucle.edu*.

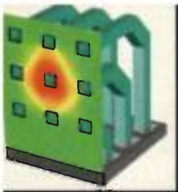
University of Missouri-Rolla— *Grounding and Shielding Electronic Systems–How to Diagnose and Solve Electrical Noise Problems*—Mar 8–10, San Diego; Apr 28–30, San Jose, CA; Jun 1–3, Toronto; Jun 8–10, Ottawa; *Circuit Board Layout to Reduce Noise Emission and Susceptibility*—Mar 8–10, San Diego; Apr 28–30, San Jose, CA; Jun 1–3, Toronto; Jun 8–10, Ottawa. Information: UMR Continuing Education, Tel 573-341-4132/4200; Fax 573-341-4992.

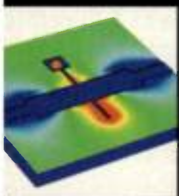
Johns Hopkins University— *GSM Systems Operations and Technology*—Mar 15–17; *Wireless Digital Communications Systems: Specification, Test, Components and Evaluation*—Mar 23–24, Research Triangle Park, NC; *Wireless and Personal Communications Systems*—Mar 29–31, Research Triangle Park, NC. Information: Anita Hellstrom. Tel. 800-683-7267; Fax 301-871-9608; e-mail *info.oei@spl.jhu.edu*.

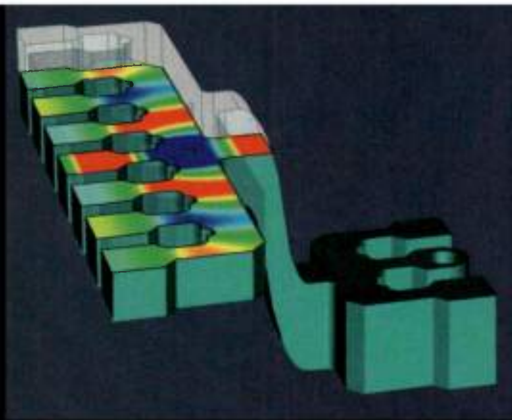
Besser Associates— *Signal Integrity of High Speed Digital Design*— Feb 18–19; *RFIC Design*—Feb 22–26; *RF Circui Design Using EM Field Simulators*—Mar 1–2; *Wireless Handsets: Architecture and Frequency Planning*—Mar 3–5; *RF and Wireless Made Simple*—Mar 8–9; *DSP Made Simple*—Mar 10–12; *Behavioral Modeling*—Mar 15–17, Mountain View, CA. Information: Besser Associates, 4800 El Camino Real, Suite 210, Los Altos, CA 94022. Tel. 650-949-3300; Fax 650-949-4400; e-mail *info@bessercourse.com;* Web site **www.bessercourse.com**.

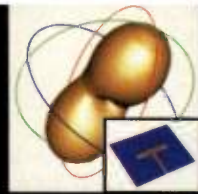
Georgia Institute of Technology—*Electrical Engineering Refresher*—Mar 6; *RF / Wireless Principles and Practice*—Apr 19–23, Atlanta; *CMOS Analog Integrated Circuits*—Jun 14–18, Santa Clara, CA.. Information: Distance Learning, Continuing Education and Outreach, Georgia Institute of Technology, Atlanta, GA, 30332-0385. Tel 404-894-2547.

Tustin Technical Institute—*Calibration and Measuremen Systems*—Mar 4–5; *Test Procedures for EMI / EMC / ESD*—Mar 8–9; *Thermal Analysis and Heat Transfer*— Apr 14–16; *Grounding and Shielding for EMI / EMC / ESD*—Apr 19–21, Santa Barbara, CA.. Information:Tustin Technical Institute, 22 East Los Olivos Street, Santa Barbara, CA 93105. Tel. 805-682-7171; Fax 805-687-6949; e-mail *Training@TTIedu.com;* Web site **www.ttiedu.com**.









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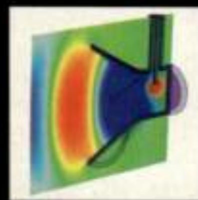

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CEMA establishes new mobile standards group

The Consumer Electronics Manufacturers Association (CEMA), Arlington, VA, has created a new committee to establish new engineering standards for mobile electronics. The R-6 Mobile Electronics Committee will develop standards in conjunction with other organizations or standard bodies such as the Society of Automotive Engineers (SAE).

The R-6 committee will be co-chaired by James Tranchina, vice president of engineering of mobile electronics for Audiovox and Richard Coe, director of engineering and applications for Clarion Sales.

One of the first actions of the committee is to participate in the development of the Intelligent Transportation Systems (ITS) data bus (IDB) standards currently being developed by the SAE. Other areas the committee will pursue include the development of two working groups to consider high speed FM subcarrier standards.

The first meeting of the committee is scheduled for this month, February 1999, during the CEMA Engineering Forum in Point Clear, AL. Anyone interested in participating in the committee should contact Tom Mock, director of engineering for CEMA at 703-907-7649 or via e-mail at *tomock@eia.com*.

ITS system market growth to be in the billions

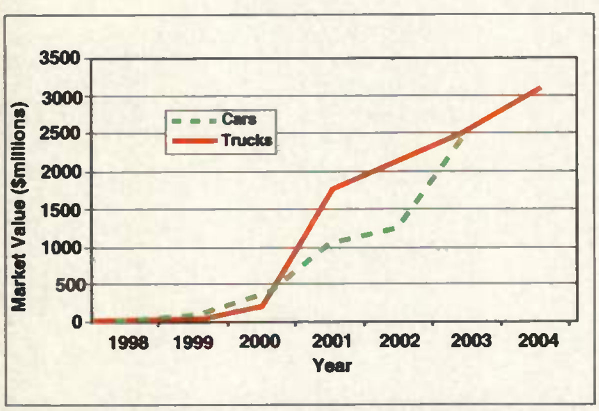
The global market for in-vehicle intelligent transportation systems (ITS) is projected to grow from its current $1 billion to $18 billion per year in the next five-years. In a new report from Allied Business Intelligence (ABI), Oyster Bay, NY, *Intelligent Transportation Systems: Wireless In-Vehicle Navigation and Communication Technologies, Global Markets & Forecasts,* the ITS market will be the next multbillion dollar market.

Major ITS systems include: in-vehicle communication systems (IVCS); in-vehicle navigation systems (IVNS); electronic toll collection (ETC) using smart cards and transponders; automatic vehicle identification (AVI); automatic vehicle location (AVL); and collision avoidance systems (CAS)

"CAS will be the big market winner," says Michael Kujawa, senior transportation analyst with ABI. "The CAS market will surpass $10 billion per year within five-years."

ABI notes that more than 14,000 vehicles are now in the General Motors OnStar program. Drivers get voice-activated cellular communications, mayday support, navigation and point of interest directions from a central service. Other car manufacturers are watching the program very closely.

For more information concerning this report, go to ABI's Web site at **www.alliedworld.com**.



One of the leading technologies expected to have significant market growth is in truck and car collision avoidance systems.

IVI gains new members in push toward interoperablity

The Interchangeable Virtual Instruments (IVI) foundation has gained 10 new members including Advantest, Anritsu, Ascor, LeCroy, Racal Instruments, Rohde & Schwarz, Tektronix, Teradyne, TYK and Wavetek.

The IVI Foundation is an open consortium that builds on VXI plug and play driver standards so that users can change instruments in test systems without making software modifications. More information concerning the foundation is available on the foundation's Web site at **www.ivifoundation.org**.

Contracts

M-tron awarded $1.8 million contract—M-tron Industries, Yankton, SD, has received a contract from a major manufacturer of network switching equipment valued at $1.8 million. The contract is for frequency control products.

Celeritek wins $4.7 million award—Celeritek, Santa Clara, CA has received a $4.7 million contract from Innova, Seattle. Celeritek will provide Innova with transceiver products for Innova's XP2 single T1/E1 digital radio.

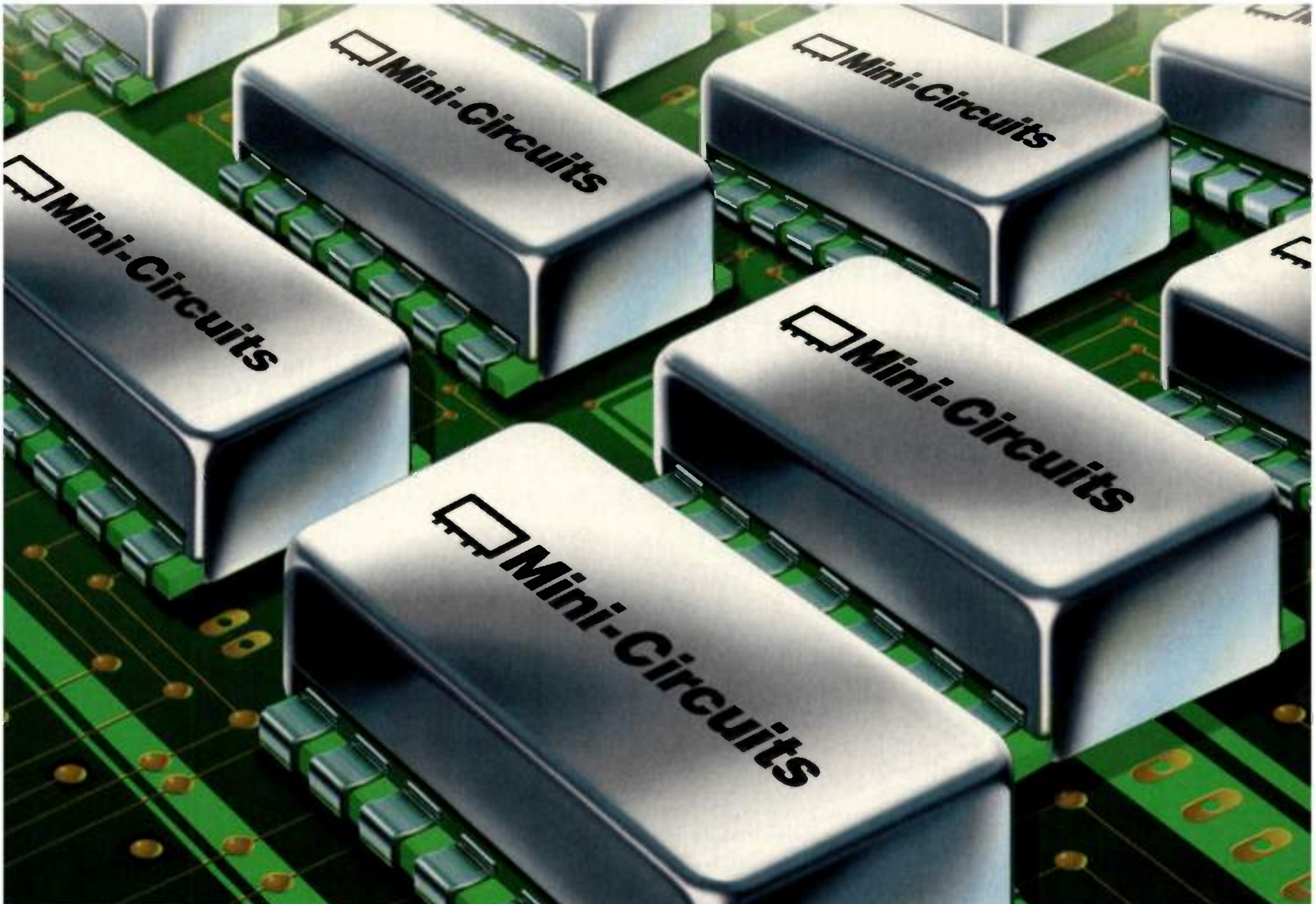
Business Briefs

Intarsia teams with Philips—Intarsia, Silicon Valley, and Philips Components, The Netherlands, are teaming to develop standards for integrated passive devices in chip scale packages. The effort will focus on developing standard peripheral and grip array packaging outlines for use in future portable and handheld electronic devices.

TAS acquires NoiseCom test lines— Telecom Analysis Systems (TAS), Eastontown, NJ, has signed an agreement to acquire the wireless and satellite communications testing product lines of NoiseCom, Paramus, NJ, for $19 million. The engineering, technical support and administrative staff from NoiseCom will join TAS to support existing NoiseCom products.

Seattle Silicon and Microchip Technology join forces—Seattle Silicon, Bellevue, WA and Microchip Technology, Chandler, AZ, have signed a shared technology licensing agreement to produce system-on-a-chip 8-bit microcontroller integrated circuits (ICs). Seattle Silicon will license Microchip's PIC16C5X 8-bit processor core as the standard platform for a new custom application-specific IC (ASIC) design.

Sawgrass acquires Andersen Laboratories—Sawgrass Electronics, Worchester, MA, has acquired Andersen Laboratories, Bloomfield, CT. The acquisition also includes Creative Electric, Auburn, NY. Anderson manufacturers high-performance acoustic signal processing devices.



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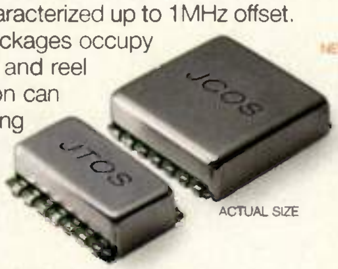

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	JTOS-50	25-47	-108	-19	15V	20	13.95
	JTOS-75	37.5-75	-110	-27	16V	20	13.95
	JTOS-100	50-100	-108	-35	16V	18	13.95
	JTOS-150	75-150	-106	-23	16V	20	13.95
	JTOS-200	100-200	-105	-25	16V	20	13.95
	JTOS-300	150-280	-102	-28	16V	20	15.95
	JTOS-400	200-380	-102	-25	16V	20	15.95
	JTOS-535	300-525	-97	-28	16V	20	15.95
	JTOS-765	485-765	-98	-30	16V	20	16.95
NEW	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-1650	1200-1650	-95	-20	13V	30	19.95
	JTOS-1910	1625-1910	-92	-13	12V	20	19.95
	JTOS-2000	1370-2000	-95	-11	22V	30 (@8V)	19.95
	JTOS-3000	2300-3000	-90	-22	***	25 (@5V)	20.95
	JCOS-820WLN	780-860	-112	-13	***	25 (@9V)	49.95
	JCOS-820BLN	807-832	-112	-24	14V	25 (@10V)	49.95
	JCOS-1100LN	1079-1114	-110	-15	***	25 (@8V)	49.95

Notes: *Prices for JCOS models are for 1 to 9 quantity. **Required to cover frequency range. ***Tuning Voltage for JTOS-3000 is 0.5 to 12V, JTOS-820WLN and JCOS-1100LN is 0 to 20V. For additional spec information, and details about 5V tuning models available, consult RF/IF Designer's Guide or call Mini-Circuits.

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F 234 Rev D

Designing an LNA for a CDMA front end

LNA design is critical in modern communication systems. Understanding necessary additional design considerations can save both time and money.

By Jarek Lucek
and Robbin Damen

The smallest signal that can be received by a receiver defines the receiver's sensitivity. The largest signal that can be received by a receiver establishes an upper power level limit of what can be handled by the system while preserving voice or data quality. The dynamic range of the receiver, the difference between the largest possible received signal and the smallest possible received signal, defines the quality of the receiver chain. The low noise amplifier (LNA) function plays an important role in the receiver design. Its main function is to amplify extremely low signals without adding noise, thus preserving the required signal-to-noise ratio (SNR) of the system at extremely low power levels. Additionally, for large signal levels, the LNA amplifies the received signal without introducing any distortions, which eliminates channel interference. Proper LNA design is crucial in today's communication technology. Because of the complexity of the signals in today's digital communications, additional design considerations need to be addressed during an LNA design procedure.

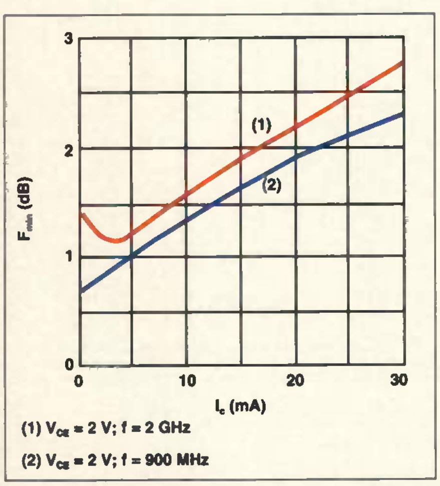

Figure 1. BFG425W minimum noise figure as a function of the collector current.

Typical trade offs in LNA design

An LNA design presents a considerable challenge because of its simultaneous requirement for high gain, low noise figure, good input and output matching and unconditional stability at the lowest possible current draw from the amplifier. Code-division, multiple access (CDMA) systems add to the challenge because of their high linearity or high third-order intercept point (IP3) requirement. Although gain, noise figure, stability, linearity and input and output match are all equally important, they are interdependent and do not always work in each other's favor. Typically, the CDMA LNA requires:

- Low supply voltage (V_{ce} = 2 V).
- Low current consumption ($I_c \leq 10$ mA).
- High gain (≥ 15 dB).
- High input IP3 (≥ 5 dBm).
- Low noise figure (≤ 2 dB).
- Unconditionally stable.
- Input return loss (≥ 10 dB).
- High isolation.
- Small dimension/low part count.
- Low cost.

Most of these conditions can be met by carefully selecting a transistor and understanding parameter trade-offs. Low noise figure and good input match is rarely simultaneously obtained without using novel feedbac[k] arrangments [1]. Unconditional sta[bility] will always require a certai[n] gain reduction because of eithe[r] shunt or series resistive loading o[f] the collector. High IP3 require[s] higher current draw, although th[e] lowest possible noise figure is usuall[y] achieved at lower current levels[.] Envelope termination technique ca[n] be used to improve IP3 performanc[e] while operating LNA at low curren[t] levels. Additional improvement of IP[3] can also be achieved by proper powe[r] output matching (P1dB match). Th[e] P1dB match, being different fron[m] conjugate gain match, reduces th[e] gain although improving IP3 perfor[-]mance.

Transistor selection

Transistor selection is the first an[d] most important step in an LNA de[-]sign. The designer should carefull[y] review the transistor selection[,] keeping the most important LNA de[-]sign trade-offs in mind. The tran[-]sistor should exhibit high gain, hav[e] a low noise figure, offer high IP3 per[-]formance at the lowest possible cur[-]rent consumption, while preservin[g] relatively easy matching at frequenc[y] of operation.

Examination of a datasheet is [a] good starting point in a transisto[r] evaluation for LNA design. The tran[-]sistor's S-parameters should be pub[-]lished at different collector/emitte[r] voltages and different current level[s] for frequencies ranging from low t[o] high values. The data sheet shoul[d] also contain noise parameters, whic[h] are essential for low noise design[.] Spice models for the transistor an[d] its package are also useful for IP[3] and P1dB simulations.

The designer should first look a[t]

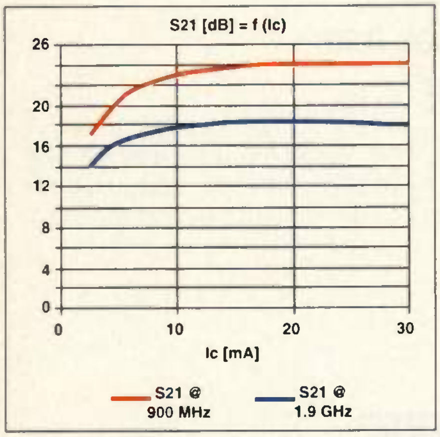

Figure 2. Forward transducer power gain.

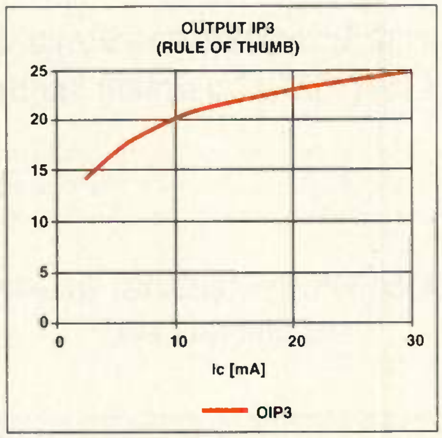

Figure 3. OIP3 vs. collector current.

three main design parameters: noise, gain and IP3, and decide what V_{ce} and I_c levels will produce optimal performance. A closer examination of NF vs. collector current, shown in Figure 1, indicates that the minimum noise figure can be achieved at around 4 mA at both 900 MHz and 1.9 GHz.

Gain available from the transistor vs. collector current is shown in Figure 2 and reveals another important aspect in LNA design: the forward transducer power gain of 18 dB remains constant at 1.9 GHz for current levels above 10 mA (24 dB for 900 MHz). Small gain degradation is expected at low current operation, below 10 mA.

The forward transducer power gair represents the gain from the tran sistor itself with its input and outpu presented with 50 Ω impedance. The S_{21} values are provided by the manu facturer of the transistor at multiple frequencies and different V_{ce} and cur rent levels. Additional gain can be ob tained from source and load matching circuits [2,3,4]. Maximum stable gair (MSG) and maximum power gair (G_{max}) are good indicators of addi tional obtainable gain from the LNA circuit.

LNA linearity is another importan CDMA LNA parameter. A figure o merit for linearity is the IP3. A two tone test is used for derivation of IP3 [5]. As a rule of thumb for bipola junction transistors (BJT), the output-IP3 can be estimated from the following formula:

$$OIP3 = 10\ log(V_{ce} \cdot I_c \cdot 5)\quad [\mathrm{dBm}]$$

where V_{ce} is in V and I_c is in mA.

The graph of OIP3 vs. collector cur rent can be derived. Figure 3 show

Neighborhood watch, Arctic style.

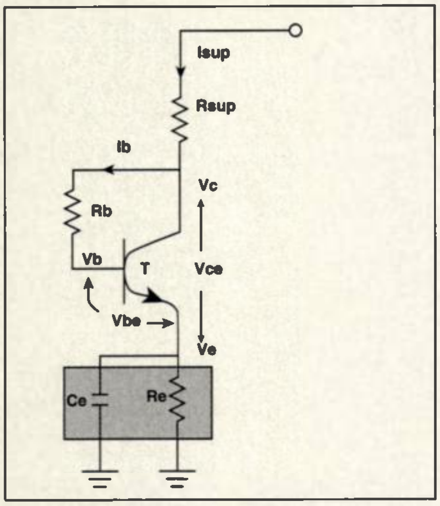

Figure 4. Typical LNA biasing circuit.

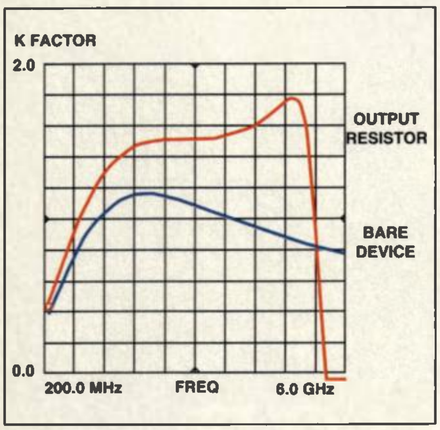

Figure 5. Stability factor over frequency.

the result. The relation between IIP3 and OIP3 is defined as:

$$IIP3 = OIP3 - Gain \quad [\mathrm{dBm}]$$

Using 15 dB for target gain and by examining the graph of Figure 3, one can determine that the transistor will need to be operated at at least 10 mA to produce a 5 dBm of IIP3 without any margins. Additional IIP3 enhancement techniques will be needed to produce IIP3 of at least 5 dBm at 10 mA of collector current.

V_{ce} = 2 V and I_c of 10 mA is the point where the transistor will produce an acceptable gain of at least 15 dB with a noise figure below 2 dB at both 900 MHz and 1.9 GHz. IIP3 will also be above 5 dBm with a collector current level of 10 mA.

LNA design

1. *DC biasing* represents the first step in LNA design. The chosen DC bias circuit should exhibit stable thermal performance and reduce the influence of h_{FE} spread. It also should be a cost-effective and simple solution, one that does not increase the complexity of the design and preserves smallest possible size for the overall LNA. The resistive feedback arrangement shown in Figure 4 is the simplest form of DC biasing that fulfills all the major requirements.

Two bias feedback arrangements are possible: one with a combination of R_{sup} and R_b and a second one with a simple R_e and C_e combination. The operation of the R_{sup} and R_b is as follows: R_{sup} and R_b will establish a biasing point. Because the operation of the LNA is going to be class A (constant current draw for dynamic range of power levels), a stable biasing point over different temperatures and for different lot codes of transistors is needed, where a small variation in h_{fe} can be expected. V_c in terms of V_{sup} and I_{sup} can be expressed as follows:

$$V_c = V_{sup} - I_{sup} \cdot R_{sup}$$

As I_{sup} decreases, which could be the case with a part with lower h_{fe}, V_c will increase at the same time. With an increase of V_c, higher I_b will result. With higher I_b, increase in I_c ($\sim I_{sup}$) will take place as high as a stable level set by R_{sup} and R_b. The same circuit handles thermal variations well. With a temperature increase, I_{sup} will increase, which will lower V_c. Lower V_c will result in lower I_b and lower I_b will lower I_c ($\sim I_{sup}$). This circuit is inexpensive, simple and takes little real estate, while its performance is well behaved and understood. For R_b to have little influence on source matching, which is crucial for noise performance, the feedback network should be decoupled with an inductor (making biasing invisible at RF band of operation).

Another possible bias feedback can be realized with emitter resistor and capacitor, shown in shaded color in Figure 4. With I_{sup} ($\sim I_e$) decreasing, V_e will decrease. V_{be} will increase with a decrease in V_e. With increase in V_{be}, I_{sup} will increase, although keeping a stable biasing point. C_e should be selected carefully, because R_e will also have a direct effect on RF gain of LNA. C_e should present a short at frequency of operation to limit its influence on gain and noise performance of the circuit.

Other biasing methods are suitabl for class A networks. These are usually closed feedback arrangement with dynamic bias control provide by active components [6]. Althoug suitable for LNA application, thes active feedback bias networks increase complexity of the LNA network, introduce additional components and increase the real-estat area of the solution.

2. *Stability design* analysis shoul be the next step in LNA design Unconditional stability of the circui is the goal of the LNA designer Unconditional stability means tha with any load presented to the inpu or output of the device, the circui will not become unstable—will not oscillate. Instabilities are primaril caused by three phenomena: interna feedback of the transistor, externa feedback around the transisto caused by external circuit, or exces of gain at frequencies outside of th band of operation. S-parameters provided by the manufacturer of th transistor will aid in stabilit analysis of the LNA circuit. Two mai methods exist in S-parameter stability analysis: numerical and graphical. Numerical analysis consists o calculating a term called Rollet Stability Factor K [2,3,4]. An intermitted quantity called delta (Δ should be calculated first to simplif the final equation for the K-factor.

$$\Delta = S_{11} \cdot S_{12} - S_{21} \cdot S_{12}$$

then

$$K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{12}|^2}{2 \cdot |S_{11}| \cdot |S_{12}|}$$

When the K factor is greater tha unity, the circuit will be unconditionally stable for any combination o source and load impedance. When K is less than unity, the circuit is potentially unstable and oscillation ma occur with a certain combination o source and/or load impedance presented to the transistor. The K facto represents a quick check for stabilit at given frequency and given bia condition. A sweep of the K-facto over frequency for a given biasin point should be performed to ensur unconditional stability outside of th band of operation. Figure 5 shows tw stability factor curves: for the transistor itself and for the complete LN circuit. The designer's goal is to design an LNA circuit that is uncondi

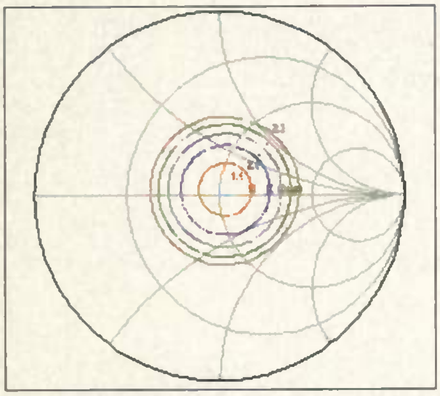

Figure 6. BFG425W 1.9 GHz, 2 V, 10 mA noise circles.

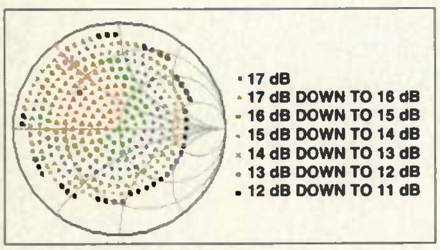

Figure 7. Gain contours for 1.9 GHz LNA.

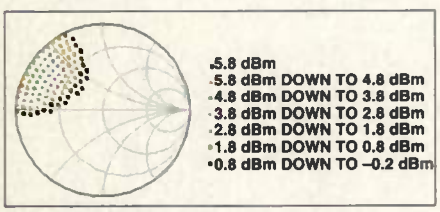

Figure 8. IIP3 contours for 1.9 GHz LNA.

tionally stable for the complete range of frequencies where the device has a substantial gain.

An LNA designer can use at least five methods for circuit stabilization. The first one consists of resistive loading of the input. This method, although capable of improving the stability of the circuit, also degrades the noise of the LNA and is almost never used. Output resistive loading is a preferred method of circuit stabilization. This method should be carefully used because its effects are lower gain and lower P1dB point (thus lower IP3 point). The third method uses collector to base resistor-inductor-capacitor (RLC) feedback to lower the gain at the lower frequencies and hence improve the stability of the circuit. The fourth method consists of filter matching, usually used at the output of the transistor, to decrease the gain at a specific narrow bandwidth frequency. This method is frequently used for eliminating gain at high frequencies, much above the band of operation. Short circuit quarterwave lines designed for problematic frequencies, or simple capacitors with the same resonant frequency as the frequency of oscillation (or excessive gain) can be used to stabilize the circuit. The final stabilization method can be realized with a simple emitter feedback inductor. A small emitter inductor can make the circuit more stable at higher frequencies.

3. *Noise matching*—The next step in LNA design consists of noise and input return loss (IRL defines how well the circuit is matched to 50 Ω) matching of the source. A typical approach in LNA design is to design an input matching circuit that terminates the transistor with a conjugate of Γ_{opt}, which represents the terminating impedance of the transistor for the best noise match. In many cases, this means that the input return loss of the LNA will be sacrificed. The optimal IRL can be achieved only when the input matching network terminates the device with a conjugate of S_{11}, which in many cases is different from the conjugate of Γ_{opt}. An emitter inductor feedback can rotate S_{11} closer to Γ_{opt}, which can help with obtaining close to minimum noise figure and respectable IRL simultaneously. This additional inductance at the emitter of the transistor will also reduce the overall available gain of the network and can be used in balancing trade-offs between the gain, IIP3 and stability in LNA design. A typical method used in designing input matching network is to display noise circles and gain/loss circles of the input network on the same Smith chart. This provides a visual tool in establishing an input matching network for the best IRL and noise trade off. This method is widely used and is also well published [7].

A slightly different design approach will be followed in the CDMA LNA example because of a special case described below. Figure 6 shows noise figure circles for a transistor at 2 V, 10 mA and 1.9 GHz. The input match is exclusively used for obtaining optimal noise performance of the LNA although preserving good IRL. A closer examination of Figure 6 reveals that Γ_{opt} coincides with the 50 Ω point. This means that almost no matching is required with the input network of the transistor (simple 50 Ω line along with the self resonatin at frequency of operation coupling ca pacitor will be sufficient) to obtai minimum specified noise figure at th given frequency of operation an given operating point. For the 90 MHz circuit, a small emitter induc tance will be used to bring S_{11} poin and Γ_{opt} point closer together, thu preserving respectable IRL. This in ductance will be achieved with smal strip lines connected directly to th emitters of the transistor.

4. *Loadpull matching*—The las step in LNA design involves outpu matching of the transistor Traditionally, this step used to be rela tively simple. An additional resisto either in series or parallel, has bee placed on the collector of the transisto for circuit stabilization. Conjugat matching has been exclusively use for narrowband LNA design to maxi mize the gain out of the circuit. Wit additional IP3 requirement forced o the LNA, the trade-off between IP and gain must be considered Linearity matching is widely know by high-power amplifier designers, es pecially those who deal with linea systems, but is relatively unknown fo a small signal designer. The so-calle load pulling is used to establish IP and gain impedance contours. Th load pulling can be realized by usin nonlinear Spice model of the transisto with simulation software. Harmoni balance can be used for establishin two tone environment. The load pulling method sweeps impedance c the whole Smith chart and plots co tours of constant gain and IP3 num bers. Figure 7 shows gain contours a 1.9 GHz and Figure 8 shows IIP3 co tours. The optimal gain impedanc point does not match the optimal IIP point, which means that the desig will have to be realized by means of trade off. Typically, the designe should design the LNA circuit at th point where the gain does not degrad as much, and the IP3 is still re spectable. If one were to draw a lin between the optimal gain and IIP3 im pedance points, every point on tha straight line will represent a good are of trade-off, with the ends repre senting the two optimal points.

The rule of thumb for P1dB an IP3 is:

$$IP3 = P_{1dB} + 10 \quad \text{in dBm}$$

This means that by knowing th

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MSWA-2-20 (Absorptive)	DC-2.0	1.3	+27	40	3.45
MSWT-4-20 (Transfer)	DC-2.0	1.8 TX** 2.0 RX***	+28 TX** +20 RX***	30	3.95

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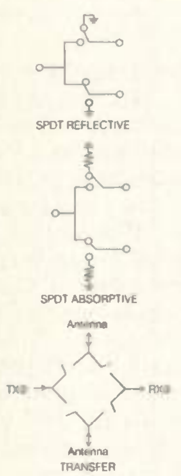

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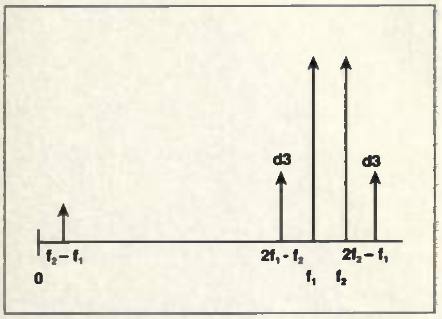

Figure 9. Two tones with in and out of band distortions.

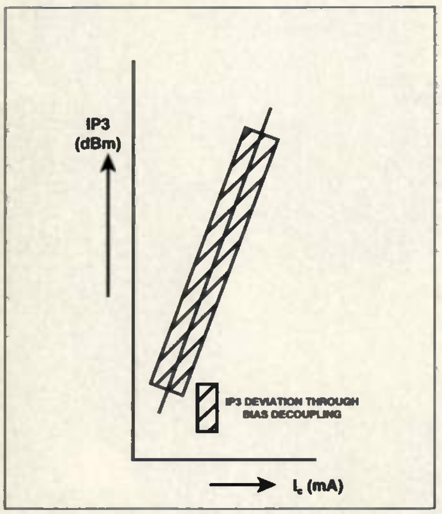

Figure 10. IP3 deviation through by-pass enhancement.

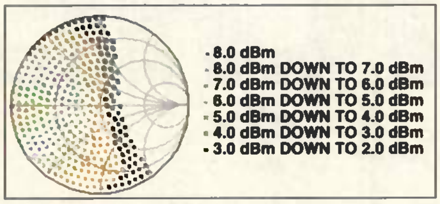

Figure 12. IIP3 contours for decoupling corrected LNA circuit, 1.9 GHz LNA.

P1dB point, one can estimate the IP3 levels. The 10 dB rule can further be improved with appropriate bypassing of the base and collector [8]. As previously indicated, the IIP3 is established by injecting two equal-in-magnitude signals with small frequency offset into an active circuit. As the active circuit approaches nonlinear region, close to P1dB, the two carries will generate distortion products, both in and out of band. (See Figure 9.)

The low frequency products, $f_2 - f_1$, can modulate the base emitter and collector emitter LNA supply voltages. For improved linearity, the fluctuation of the base and collector voltages should be eliminated by means of proper by-passing, hence presenting the base and the collector with low impedance at so-called video frequencies (between DC and usually as high as 40 MHz, depending on the bandwidth of the signal that is being presented to the LNA). In the case of CDMA system, the video bandwidth should extend well beyond 1.25 MHz or at least 5 MHz. The designer should exhibit caution during by-passing design. A poor selection of the by-pass capacitors could also degrade IP3 performance as shown in Figure 10.

Figure 11 shows 1.9 GHz LNA with the transistor. Capacitor C2 and C5 will resonate at frequency of operation. C3 and C4 combination will work at video frequencies, thus making sure that both collector and base bias are not modulated with the distortion signals. As a rule of thumb, the impedance of by-passing circuit should be lower than 25% of the input impedance of the transistor at particular frequency spacing. In that case, the following is valid:

The impedance of the transistor is:

$$Zin\ (5\ MHz) \approx \frac{h_{fe}}{g_m} = \frac{h_{fe}}{\frac{Ic}{Vt}} = \frac{70}{\frac{10}{25}} = 175\ \Omega$$

Cd should be 25% less than 175 Ω:

$$Cd < 0.25 \cdot 175\ \Omega \approx 44\ \Omega$$

At 5 MHz spacing, the Cd should be at least:

$$Cd \geq \frac{1}{2\pi f \cdot 44} \geq \frac{1}{2 \cdot (3.14) \cdot 5E6 \cdot 44} \geq 1\ nF$$

Although preserving the gain performance of the LNA, the by-passing method (also known as an envelope termination technique) can improve LNA's IIP3 performance without increasing current consumption. Figure 12 shows IIP3 contours after implementation of video frequency decoupling. Comparison of Figure 12 and Figure 8 reveals substantial improvement in IIP3 trade off. Because the gain contours for IIP3 improved circuit will remain the same, the main improvement in IIP3 performance is achieved by extending the available IIP3 impedance points closer to the optimal gain impedance levels.

LNA circuit realization

Figure 11 shows 1.9 GHz high IP3 CDMA LNA circuit with the transistor, and Figure 13 demonstrates a typical 900 MHz LNA. Both circuits were realized with design methods described in this article.

Table 1 summarizes the measured performance of 1.9 GHz LNA circuit. Table 2 summarizes the performance

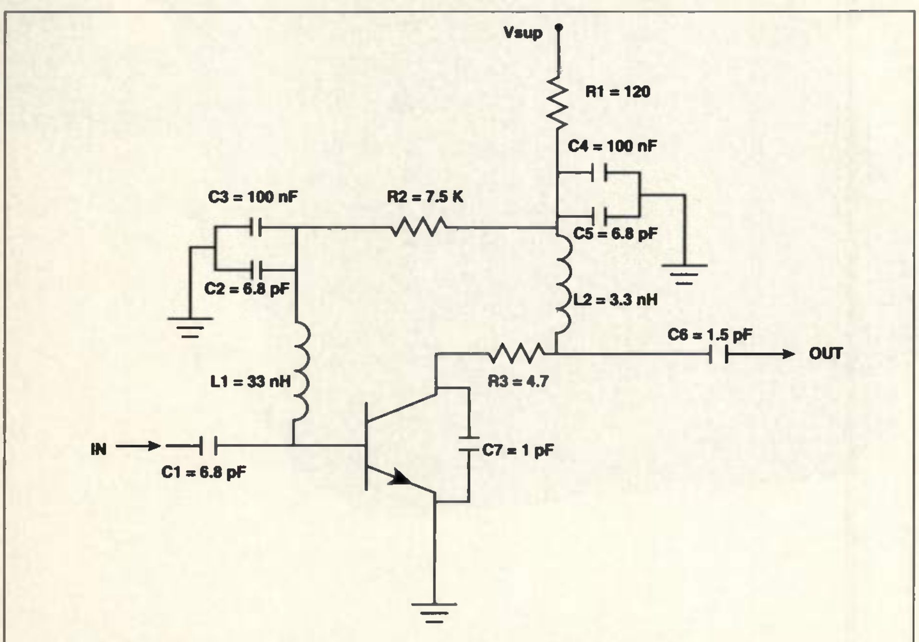

Figure 11. 1.9 GHz LNA.

Parameters	Units	Measured performance with IIP3 by-pass improvement	Measured Performance without IIP3 by-pass improvement
Vsuply	Volts	3.3	3.3
Vce	Volts	2	2
Ic	mA	10.3	10.3
Gain	dB	16.8	16.8
NF	dB	1.9	1.9
IIP3	dBm	+5 (at 1.25 MHz spacing)	–2.5 (at 1.25 MHz spacing)
IRL	dB	13	14
ORL	dB	11	10
Isolation	dB	27	27

Table 1. 1.9 GHz LNA performance.

Parameters	Units	Measured performance with IIP3 by-pass improvement	Measured Performance without IIP3 by-pass improvement
Vsuply	Volts	3.3	3.3
Vce	Volts	2	2
Ic	mA	10	10
Gain	dB	16.9	17
NF	dB	1.8	1.85
IIP3	dBm	+5 (at 1.25 MHz spacing)	–4 (at 1.25 MHz spacing)
IRL	dB	7	7
ORL	dB	11	11
Isolation	dB	28.5	28.5

Table 2. 900 MHz LNA performance.

of 900 MHz version of LNA. *RF*

Acknowledgment

Special thanks to Tom Buss, Korne Vennema and Norbert van der Bos from Philips Semiconductors. Our many fruitful discussions on the topics of LNA design greatly contributed to realization of this article.

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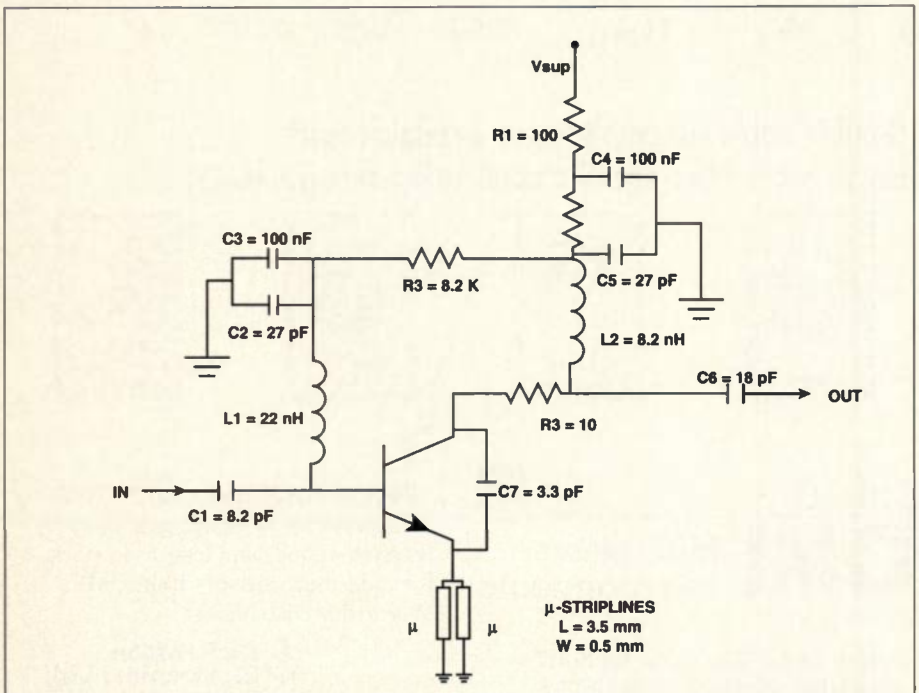

Figure 13. 900 MHz LNA circuit.

About the author

Jarek Lucek is currently working for Philips Semiconductors as a market application engineer. He is responsible for developing applications for LNAs, PA drivers, PAs, mixers and VCOs for subscriber applications. His previous experience includes designing high-power PA stages for 1, 1.5 and 2 GHz feedforward, highly linear infrastructure PAs for Motorola. He also has experience from Decibel Products in designing high-power PA repeaters. He holds a B.S.E.E. from the University of Illinois. He can be reached at 508-337-7927 or by e-mail at *jarek.lucek@sv.sc.philips.com*.

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The transistor described in this article is the BFG425W from Philips Semiconductors.

Computer programs that calculate field strength

Here are two computer programs that can help design antennas.

By James Eagleson

When designing RF systems for any application, the idea is to get a signal from at least one point to another point without using wires, cables or a waveguide. For wireless communications, the transducer used to convert radio energy into an electromagnetic wave is, of course, an antenna. The function of an antenna is esentially the same whether the application is data transfer across a wireless local-area network (WLAN); voice communication over a two-way radio; cordless, cellular or personal communications service (PCS) phones; radar systems; or any of a variety of low power devices such as radio frequency identification (RFID), radio control, remote access, wireless alarms or remote keyless entry (RKE) systems.

Two programs are helpful in determining antenna design performance. The first is AFCALC. This program allows simple calculation of antenna factor (AF) when antenna gain and frequency are known, or it calculates antenna gain if frequency and AF are known. These are useful when making Federal Communications Commission (FCC), Industry Canada, or European Community/European Telecommunications Standards Institute (EC/ETSI) measurements regarding field strength or when trying to establish actual effective isotropic radiated power (EIRP) or path loss.

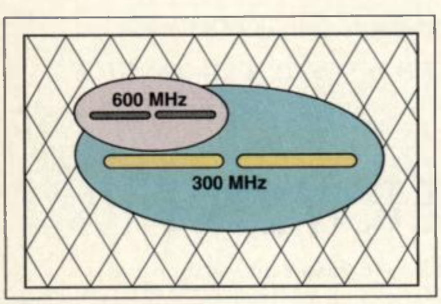

Figure 1. Antenna capture area.

The second program, FSCALC, is really a compilation of several short routines that allow calculation of "path loss," field strength, required power to obtain a given field strength, EIRP and other variations of these calculations. It was developed to quickly evaluate antenna and field strength relationships while designing various low- and medium-power radio communication devices and systems.

AFCALC

Essentially the power received by an antenna depends on the power density of the signal at the antenna's location, the antenna's orientation relative to the incoming wavefront and the ability of the antenna to capture the power that is available.

Calculating power density at a given distance is basically the same as finding the area of a sphere. At a given radius, "R," a sphere has a given surface area. When using an antenna radiating equally well in all directions (a point source or isotropic radiator), power density at a given distance will be

$$S = \frac{P_o}{4\pi R^2} \qquad (1)$$

where

S = Power Density (usually in watts/meter^2)
P_o = Power Output (usually in W)
π = 3.1416
R = Radius (r = radius or distance in meters)

Thus, the power density at a given distance is totally independent of frequency because frequency is not an element in this formula.

But antenna size is not independent of frequency. It is related to frequency by the relationship

$$L = \frac{F}{300{,}000{,}000} \qquad (2)$$

where

L = λ or wavelength (in meters)
F = Frequency (in Hz)

The formula is more convenientl[y] used in MHz, not Hz, and becomes

$$L = \frac{F}{300} \qquad (3$$

In one-millionth of a second (on[e] cycle or wavelength at 1 MHz) a wav[e] will travel 300 meters in air (or space) Thus, at 300 MHz, wavelength calculates to one meter and at 600 MH[z] wavelength is only a half meter.

Because a resonant dipole is a hal[f] wavelength long, it will be 0.50 mete[r] long at 300 MHz and 0.25 meter lon[g] at 600 MHz. (The effect of antenn[a] construction on its length will be ignored for simplicity.) The electricall[y] active area around the antenna (o[r] "capture area") will be shaped somewhat like Figure 1 so that the are[a] (length × width) of the 300 MHz antenna will be four times (2^2) that of th[e] 600 MHz antenna. AF takes this effec[t] into account.

AF can be calculated by

$$AF\,(dB) = 20\log(F) - G - 29.78 \qquad (4$$

where

F = frequency (in MHz)
K = 29.78 (a correction factor)

AFCALC uses this formula to eithe[r] calculate AF from antenna gain an[d] frequency or to calculate the antenn[a] gain from a given AF and frequenc[y] The first thing it does is to ask whic[h] method is to be used.

```
>>ANTENNA FACTOR CALCULATOR v1.00<<

CALCULATE:

1 = Antenna Factor
2 = Antenna Gain

q = quit

Enter selection . . .
```

If working in the 300–320 MH[z] spectrum (often used for Part 15 devices of various kinds), the antenn[a]

factor for a resonant dipole would be found by entering its gain and frequency.

Antenna Gain (dBi)	= 2.15 <Enter
Antenna Gain Ratio	= 1.64
Frequency (MHz)	= 315 <Enter
Antenna Factor	= 18.04 dB

In this example, the theoretical 2.15 dBi gain of a dipole over an isotropic radiator is used. This assumes a perfect match and 100% efficiency for the antenna even though this often is not the case in the real world. Most of the time, the error is small for a test dipole.

So why know the AF?

When using a spectrum analyzer to determine field strength, path loss or antenna gain, the AF needs to be measured, known or calculated for each antenna in the test setup. It is fairly easy to determine the exact path loss between two antennas if they are not far apart. This is commonly done to calibrate a three- or 10-meter test site to allow accurate measurement either of intentional or unintentional radiation from electronic equipment.

If the signal source cable is directly connected to the spectrum analyzer or milliwattmeter cable, the variations of all of these test elements are directly accounted for in the resulting measured level, P(direct).

If each cable is then connected to its respective antenna and a new measurement is made, the result will provide the loss (or gain) present caused by antenna gain(s) and path loss (L_p). L_p can be found using

$$L_P = PD - PR - AF_1 - AF_2$$

where
PD = Power Direct (directly connected)
PR = Power Received
AF_1 = Antenna Factor of Antenna 1
AF_2 = Antenna Factor of Antenna 2

This is the heart of techniques spelled out by the American National Standards Institute (ANSI) in document C63.4 for certifying the accuracy of three- and 10-meter sites and other field strength test sites. ANSI C63.4 is also specified by the FCC in Part 15 as an appropriate standard to use when making measurements for certifications and verifications of electronic equipment.

Another formula useful when trying to determine field strength (FS) is

$$FS = 107 + PR + AF \tag{6}$$

where
107 = A Constant derived from the fact that 0 dBm is 107 dB above 1 μV at 50 Ω.

If a 315 MHz tuned dipole is connected to a spectrum analyzer that receives 0 dBm from a radiating source, FS is

$$FS = 107 + 0 + 18.04 = 125.04 \text{ dB}\mu\text{V/m}$$

However, by the time UHF frequencies go above 300 MHz, there is significant loss in even a fairly short coaxial cable. Because there will be cable loss between the antenna and the analyzer, the field strength must actually be slightly higher than 125.04 dBμV/m. In fact, it will be higher by the amount of loss in the cable.

Modifying the formula to account for cable loss adds a factor, LC, or loss of cable. The simplest thing to do is to add the cable loss as a positive number (absolute value) so the formula becomes

$$FS = 107 + Pr + AF + |LC| \tag{7}$$

If cable loss is –0.76 dB at 315 MHz, FS becomes 0.76 dB higher than the previous calculation

$$FS = 107 + 0 + 18.04 + |-0.76| = 125.8 \text{ dB}\mu\text{V}$$

When working with Part 15 devices, it is common to add a peak-to-average factor as allowed by FCC rules. If, for example, the FCC allows 67 dBuV/m average FS at a distance of three meters, they also allow a duty factor (DF) to be used according to the formula

$$DF(dB) = 20\log\left(\frac{T}{100}\right) \tag{8}$$

where
T = Time (milliseconds)

This allowance accounts for the lower potential for interference that short pulses have compared to continuous wave (CW) signals but the FCC sets a limit on this factor of –20 dB. They also require that DF must be calculated for the worst case 100 ms period within any given transmission.

Three examples give the CW, 50% duty and 10% DF

T = 100 ms; DF = 20 log (100/100) = 0 dB
T = 50 ms; DF = 20 log (50/100) = –6 dB
T = 10 ms; DF = 20 log (10/100) = –20 dB

Note that the DF is a negative value. Average power is lower than peak power by the amount shown in the formula. Because a spectrum analyzer actually displays *peak* values adding DF to this will give the average value.

Thus, for the previous examples, average power when the analyzer gives 67 dBuV/m as the peak value are:

FS = 67 dBμV/m + 0 dB = 67 dBμV/m
FS = 67 dBμV/m + (–6) = 61 dBμV/m
FS = 67 dBμV/m + (–20) = 47 dBμV/m

Putting it all together produces a final formula

$$FS = 107 + Pr + AF + |LC| + DF \tag{9}$$

A level of 0 dBm and a 10% (10/100) DF of –20 dB gets

$$FS = 107 + 0 + 18.04 + |-0.76| + (-20) = 105.8 \text{ dB}\mu\text{V/m}$$

When trying to meet various regulations for products, the level the analyzer will see at the allowed limit is what needs to be known. This is done by re-arranging the formula

$$PR = FS - (107 + AF + |LC| + DF) \tag{10}$$

Because the previous example was giving a level of 125.8 dBμV/m, it is clearly in violation of the FCC level that was cited earlier. Even if data transfer rate requirements allow use of a 10% DF (–20 dB), the result is still 105.8 dBμV/m average FS—well above the levels allowed at 300 MHz by the FCC.

If requirements for data transfer allow use of a 10% DF and several other restrictions can be tolerated (see Part 15.241e), the full peak-to-average advantage of –20 dB means that to achieve 67 dBμV/m average field strength, as high as 87 dBμV/m peak signal (i.e., 87 dBμV/m peak – 20 dB duty factor = 67 dBμV/m average) can be used.

However, when Equation 10 is used to get PR, it is the wrong answer.

$$Pr = 87 - (107 + 18.04 + 0.76 + (-20))$$
$$= 87 - (105.8) = -18.8 \text{ dBm}$$

This is because the spectrum analyzer reads peak, not average, so ignore DF when back tracking to the expected PR because it is already th

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peak level. The formula should be

$$Pr = FS - (107 + AF + LC) \tag{11}$$

The level for the example becomes

$$Pr = 87 - (107 + 18.04 + 0.76)$$
$$= 87 - (118.8) = -38.8 \text{ dBm}$$

Thus, to remain below required FCC levels, the peak signal seen on the spectrum analyzer must remain below –38.8 dBm, which is the correct answer.

FSCALC

The second program has a number of selections for general use when working with antennas and field strength. FSCALC also uses a menu to select the calculation desired. The same function can be reused by pressing the same letter used to select the function initially or make another selection. Return to the menu by pressing "m" or exit the program by pressing "q."

FIELD STRENGTH CALCULATOR

SELECT DESIRED FUNCTION:

a.) PATH LOSS:	Given Frequency & Distance
b.) EIRP:	Given Po, Lc, Ga
c.) dBuV/m:	Given EIRP, Distance
d.) dBuV/m:	Given Po, Lc, Ga, Distance
e.) dBuV/m:	Given Pr, Lc, AF
f.) Required EIRP:	For Desired dBuV/m @ Distance
g.) Required Po:	For Desired dBuV/m @ Lc, Ga, DI
h.) Required Pr:	For Desired dBuV/m @ Lc, Ga
i.) Required Ga:	For Desired dBuV/m @ Lc, Pr, DI
j.) Antenna Factor:	Given Frequency & Gain

Selection <a – j, <m> = menu, <q> = quit

Selecting a letter from "a" to "j" selects one of the calculations described below.

• *PATH LOSS*—The following formula is used to calculate "path loss"

$$Lp = 27.125 - (20\log(F) + 20\log(R)) \tag{12}$$

where
Lp = Path Loss (dB)
R = radius (distance in meters)
K = 27.125 (constant assuming MHz and meters)

A sample calculation is:

$$Lp = 27.125 - (20\log(315) + 20\log(3)) = -32.4 \text{ dB}$$

The program will trap certain errors such as zero distance or a distance too close to obtain an accurate path loss calculation.

• *EIRP*—The following formula is used to calculate EIRP

$$EIRP = Po + Lc + Ga \tag{13}$$

where
Po = Power Output (dBm)
Lc = Loss of Cable (dB)
Ga = Gain of Antenna (dBi)

• *INPUT UNITS*—The program also allows selection of either Watts or dBm for most calculations. Whichever is selected, the program makes the appropriate conversion(s) first and the final answer is given in both units.

The formula used to convert from Watts to dBm is

$$Po = 10\log(1000 \times PW) \tag{14}$$

where
PW = Power Output (W)
K = 1,000 (Converts W to mW)

The formula used to convert from dBm to Watts is

$$PW = \frac{10^{\frac{EIRP}{10}}}{1000} \tag{15}$$

where
1,000 = conversion to W from mW

The answer is given in dBm as well as µW, mW or W, depending on the final value. If an answer is less than 0.001 W, it will be given in µW. If the answer is less than 0.01 W but greater than 1 mW, the answer is given in mW. Otherwise the answer is given in W.

• *FIELD STRENGTH*—Three ways are provided to calculate FS with the answer given in dBµV/m.

1. *dBµV/m given EIRP and distance*—If EIRP and distance are known, Equations 16 and 17 are used

$$VM = \frac{\sqrt{30 \cdot PW}}{DI} \tag{16}$$

where
VM = V/m
DI = distance

Then,

$$DBUV = 20\log(1 \times 10^6 \cdot VM) \tag{17}$$

where
DBUV = dB above 1 µV per meter
1×10^6 = 1,000,000 µV per V

2. *dBµV/m given PO, LC, GA and distance*—By combining the EIR[P] equation with the equations just use[d] FS can be found by filling in th[e] blanks on antenna gain, cable los[s] power out and distance.

First EIRP is calculated the sam[e] way as was done with Equation 1[5] The same selection for dBm or W i[s] provided and the same conversion take place as required. This value i[s] then used in Equations 16 and 17 t[o] calculate the answer.

3. *dBµV/m given PR, LC, AF*—Thi[s] calculation determines FS from mill[i]wattmeter or spectrum analyzer read[-]ings. A word of caution, however. Mos[t] spectrum analyzers read peak powe[r] although most wattmeters read av[-]erage power so when measuring any[-]thing having less than 100% dut[y] cycle, a conversion must be made t[o] get the desired value.

If using an averaging wattmete[r] average field strength will be provide[d] by this calculation. To convert to pea[k] field strength, add 20 log (Period/Puls[e] Width). In this case, period is the tim[e] between pulses (or pulse chains[)]. Pulse width is the width of total "o[n]" time for a single or a chain of pulses i[n] a given period (e.g., 10 pulses of 1 m[s] each would be 10 ms total "on" tim[e] regardless of their spacing.) A trans[-]mission having one such pulse trai[n] for each 100 ms period yields 20 lo[g] (10) = 20 dB.

If using a spectrum analyzer, be[-]cause it reads peak, not average, th[e] calculation provides peak fiel[d] strength. To find the average fiel[d] strength subtract 20 log (Period/Puls[e] Width).

$$DBUV = 107 + PR + AF - LC \tag{18}$$

where
DBUV = dB above 1 µV per meter
107 = constant based on 0 dB[m] being 107 dB above 1 µV
LC = Loss of Cable (i[n] dB...watch the sign, it will be –x dB)

This is the formula used earlier.

• *CALCULATING REQUIREMENTS FOR FS*—The program pro[-]vides calculations of four required pa[-]rameters to generate a desired FS.

1. *Required EIRP given dBµV/m, DI*—If desired distance and FS ar[e] known, the EIRP needed to generat[e] the desired goal can be calculate[d] using

$$VM = \frac{10^{\frac{DBUV}{20}}}{1 \times 10^{6}} \quad (19)$$

$$W = \frac{(DI \cdot VM)^{2}}{30} \quad (20)$$

$$EIRP = 10\ \log(W) + 30 \quad (21)$$

where
30 = constant added to power in dBW to obtain dBm
(Note: 10 log (W) produces dBW or dB above one W)

The answer is provided in both dBm and µW, mW or W and is automatically converted to the units that make the most sense (1 mW instead of 0.001 W, for example).

2. *Required PO given dBµV/m, DI*—This calculates EIRP, then determines power output (PO)

$$PO = EIRP - GA - LC \quad (22)$$

where
GA = Gain of Antenna (in dBi)
LC = Loss of Cable (in dB... a minus value)

3. *Required PR for desired DBUV, GA, LC*—To know the received power in dBm expected from a given cable, antenna and FS, use

$$PR = DBUV - 20\log(FR) + 77.213 + LC + GA \quad (23)$$

where
PR = Power Received (dBm)
FR = Frequency (MHz)

4. *Required GA given dBµV/m, LC, PR*—To determine what antenna gain is needed to provide a desired signal level, pr (in dBm) into a receiver and the FS, cable loss and frequency is known

$$AF = DBUV - (LC + PR + 107) \quad (24)$$

a derivative of an earlier formula.

$$GR = \sqrt{\frac{FR}{30.82 \times 10^{\frac{AF}{20}}}} \quad (25)$$

where
GR = Gain ratio of the antenna

$$GA = 10\ \log(GR) \quad (26)$$

(Standard conversion of ratio to dB)

• *AF given GA, FR*—This is don using the equation from AFCALC.

Conclusion

These formulas and description should provide help for those needin to calculate various antenna, fiel strength and related relationships The programs are available for down load on the *RF Design* Web site Although I do not expect to updat this set of programs, I will be happy t correspond with anyone with sugges tions, additions or corrections. M Web site and e-mail address are in cluded. Both programs run usin Windows 3.1 or Windows 95. Be ad vised that because I am not a pro grammer by trade, little error correc tion is present in either of thes programs so that incorrect input re sults in ungraceful exit. Personally, have not had it hang up the computer but the program does need to b restarted if you type in bad informa tion (Garbage In = Garbage Out There is some automatic scaling of cer tain variables (µW, mW or W, for ex ample,) and FSCALC traps things lik division by zero or too close distanc and so forth. RF

Reference

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1. *A Guide to FCC Equipmen Authorizations* by Willmar K. Roberts
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3. ANSI Publication C63.4.
4. FCC OET Bulletin No. 63 October 1993, reprinted Februar 1996.
5. Various EMC and EMI Web sit links via my Web site.

About the author

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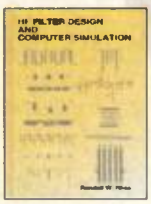

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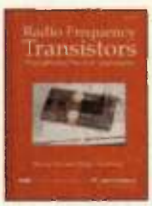

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SAR advances conquer data acquisition challenges

Major developments in software, hardware, signal, imaging and data processing technology have enabled SAR to assume increasing roles in a variety of applications.

By Ernest Worthman, Contributing Editor

The science of radar-based data acquisition, both automated and manned, has been significantly enhanced by the technological developments of the late 20th century. One of the major recipients of this technological revolution is synthetic aperture radar (SAR).

Although some SAR has existed since the late 1970s, it is really coming of age in the 1990s because of rapidly developing computer processing capabilities. SAR can address many of the problems that have plagued optical sensing and real aperture radar. And SAR is impervious to all but the most extreme conditions that obstruct both optical and real radar from acquiring accurate data.

Digital signal processors (DSPs), analog-to-digital converters (ADCs), image processing software and high-speed computer data analysis all contribute to SAR's ever increasing share of the image and data acquisition market. SAR has given markets such as geology, meteorology and ecology, as well as the military, the ability to acquire precise and detailed data, with unprecedented accuracy, from aircraft and satellites.

It is difficult to assess the impact that SAR will have on these and other data acquisition markets over the next few years. SAR is still relatively expensive to use because it is mainly aircraft- or satellite-based. And, because SAR is computer synthesized, it requires high-power processing and analysis hardware to process the image. Just over the last few years, the cost of the electronics required for SAR processing has become more affordable. However, one can reasonably expect to see SAR become eco nomically feasible for smaller scale ap plications as we turn the century Currently, many SAR projects are im plemented by the military, governmen or government-affiliated entities suc as the jet propulsion laboratory (JPL and Sandia Labs.

Two issues will have a significant ef fect upon SAR implementation over th next 10 years:

1. The cost of hardware is expecte to continue to decline.
2. Developments in processin power are making logarithmic leaps i capability.

Bearing these in mind, it is reason able to assume SAR will become a sig nificant source for data acquisition an monitoring in the future.

Technology and theory of operation

Over the years, radar has been re fined from its invention in 1935 by Si

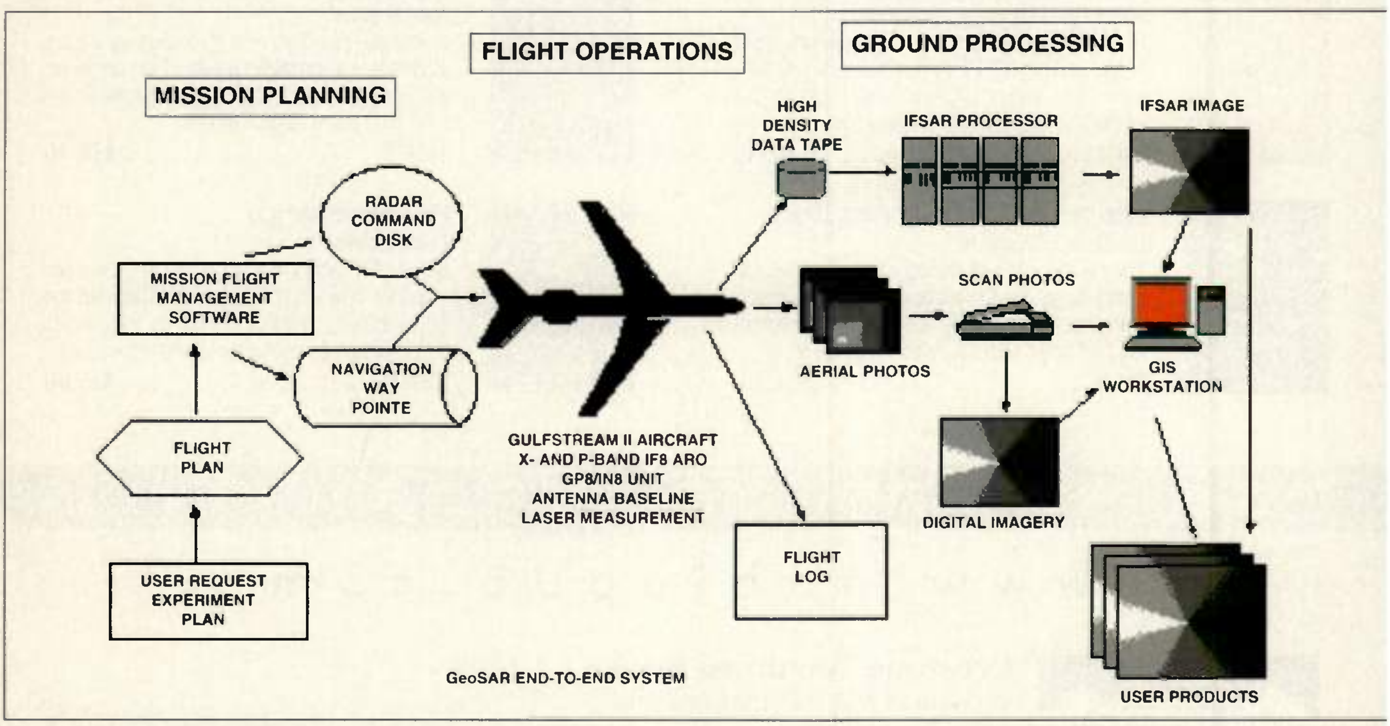

Figure 1. A GeoSAR schematic (courtesy of NASA and JPL).

Robert Watson-Watt and its early WWII 100+ MHz frequency band, to today's large aperture radar (LAR), Doppler-effect radar and SAR at as much as 40 GHz.

Essentially, radar is the science (and art) of measuring the distance of one object in reference to another. Radar uses the time delay that occurs when a pulse of energy is transmitted to the time it is received back as a reflection of the object.

A far cry from the past's relatively low frequency and short (a few thousand meters at best) range, today's SAR can be implemented at a number of frequencies and can range thousand of kilometers The wavelengths around the center fre quencies of the most popular SAR mi crowave bands are K~ 0.3-1 cm; X~ 1-: cm; C~ 5-6 cm; L~ 24-25 cm; and P~ 65 75 cm. The fundamental technology i the same, regardless of frequencies.

SAR differs from real aperture rada (RAR) and LAR by using the Doppler shift effect. Comparing the Doppler shifted frequencies (because of th movement of the emissions source) to : reference frequency allows many re turned signals to be "focused" on : single point as the real aperture move through a series of positions along th projected track.

SAR uses narrower pulses compare to conventional radar. This particula technique increases the azimuth reso lution and is commonly called azimut compression and virtualizes the aper ture to appear larger than it really is.

The way that SAR synthesizes an tenna aperture, using the Doppler ef fect, is by first modulating a chirp mod ulation of the received pulse, the taking the reversed characteristics o the chirp modulation and applying it t a matched filter scheme. This effec tively increases the length of the an tenna that is imaging that particula point and is commonly called *synthe sizing the antenna's size or aperture*.

Because this increases the azimut resolution for the same technique adopted for range direction in RAR, i reduces the physical antenna siz while maintaining the precision o LAR systems.

But without the high-tech, high speed digital processing hardware, de veloping SAR images would b painstakingly time consuming becaus SAR processing requires that the varia tion in Doppler frequency for each poin is correctly matched to image. This re quires precise knowledge of the relativ motion between the platform and th imaged objects and requires tremen dous mathematical analysis.

SAR's moving platforms are, as pre viously stated, almost exclusively air plane or satellite-based (I say *almos* because there is also some implementa tion with moving ground-based vehi cles). SAR image areas can be as smal as a few meters, or as large as severa hundred km.

Typical pulse widths are 10-50 us and typical bandwidths are 10-20(MHz around the center frequencies.

Because SAR uses wavelength be

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tween 1 cm and 1 meter, it sees through clouds and smoke, as well as works without external illumination (it is often referred to as self-illuminating imaging), such as what is required by optical data acquisition systems. Still, SAR remains a basic radar technology, however; so in spite of its precise capability, it uses conventional radar techniques, such as side-looking sensors, and acquires its data through standard radar range measurement and resolution techniques.

A bit more detail

If we take a cursory look at the theory behind SAR, we find that fundamentally, SAR produces a 2-D image.

One image dimension is along th track, commonly called *range*. Becaus SAR uses narrow pulses as opposed t conventional radar, SAR yields fine range resolutions. Although this has significant impact upon resolution, thi is not the unique parameter of SA that yields the greatest benefit.

The parameter largely responsible fc SAR's precision is the azimuth, the d mension on the perpendicular axis t range. The relationship between thi axis and the antenna parameter is quit simple—the finer the azimuth resolu tion, the larger the antenna needs to b Thus, to obtain a fine enough resolutio to obtain highly detailed data for suc applications as environmental mon toring, earth resource mapping and mil tary reconnaissance, a large antenna i needed to provide a sharp beam Unfortunately, such antennas are in practical for airborne and space vehicles

However, synthetic aperture eng neering is a relatively complicate process of analyzing received signal and phases from moving objects with small antenna. What is really bein done is mathematically converting th effects of a larger antenna, i.e. synthe sizing a smaller aperture length.

The azimuth resolution can be de rived using a formula. By applying th formula, the resolution of the azimut direction ends up being ½ of the rea aperture radar. However, it is not tha simple.

As with all virtualization and synthe sization (as well as real data acquis tion), error is introduced. With SAF data must be enhanced and correcte based upon the extrapolations involve Remember that the aperture is en hanced mathematically, thus, a lo closer attention must be paid to ensur the data interpolation is accurate Unfortunately, a full discussion of th mathematics and error analysis is be yond the scope of this article. Howeve major issues such as errors introduce by azimuth range compression tech niques, velocity effects upon th Doppler effect, and the fact that co recting for the synthetic aperture lengt is more challenging at lower altitude must all be addressed. Essentially, SA satellites that gather extremely detaile data must be very high, and airplane based SAR requires tremendously accu rate interpolation, as well as having lower detail capability.

However, even with these issue

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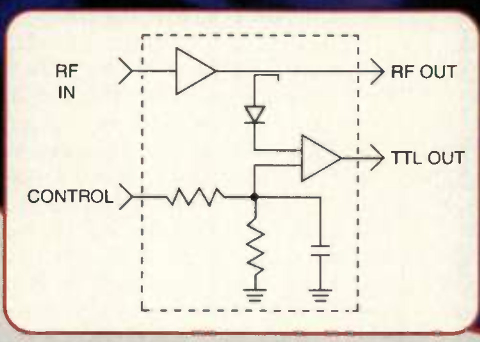

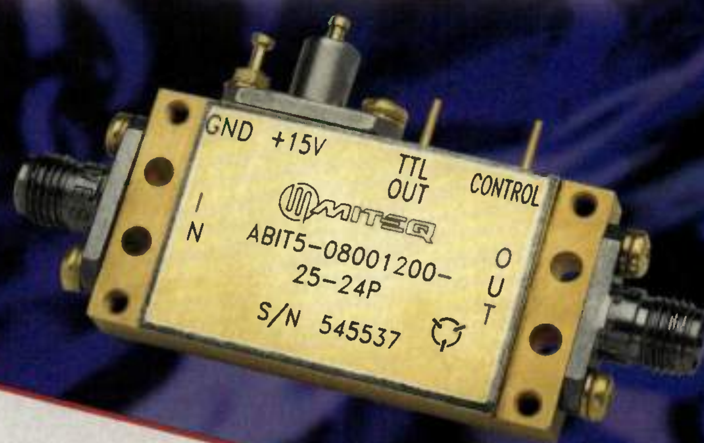

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BIT detection format	TTL single ended	TTL single ended
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SAR still offer significant resolution advantages over other types of radar and overcomes the visual limitations that optical data collection has.

The next step—Inferometry

Earlier, I mentioned that SAR is, fundamentally, a 2-D image acquisition process. However, there is a great deal of interest in applying Inferometry to SAR. Using Inferometry, SAR is capable of acquiring 3-D images.

Essentially, InterFerometric Synthetic Aperture Radar is SAR collecting data (echoes) from the same target using two spatially-separated receiving antennas. (Does the term diversity come to mind? Because the diversity signal is slightl out of phase, two separate images can b produced. From these images, the "inter ferogram" can be developed and used t determine terrain characteristics Elementally, in SAR, the radar uses in terference phenomena between the refer ence wave and the Doppler-shifted wave to measure indices of refraction, wave lengths and wave velocities. This tech nique can measure small distances an thicknesses of the imaging area.

Through a technique called phas unwrapping (which requires supe computing capability), the two image can be integrated and an A/B analysi can be used to differentiate and resolv height, density, and other 3-D parame ters. After resolution, a 3-D DEM (Digital Elevation Model) digital ter rain model can be built. To oversim plify, IFSAR is, basically, SAR with second antenna—similar to usin binoculars over a telescope.

Another interesting SAR techniqu along the lines of multidimensiona modeling is *multipolarization SAR* Depending on the SAR configuration transmitted pulses can be either hor zontal or vertical and received in th same domain.

If you analyze all of the combination of these polarized pulses, you end u with four possible combinations: hor zontal transmit/receive (HH); horizonta transmit/vertical receive (HV): vertica transmit/receive (VV); and vertica transmit/horizontal receive (VH). A SAR becomes more complex, and with third frequency used, polarization ca be used to give the image yet anothe axis to integrate. Additionally, som SARs can also measure phase differ ences in the returned multipolar echoes

LightSAR—lightweight, accurate and cost-effective

SAR systems, in spite of the de clining equipment costs and muscle-u computing power, are still expensive t build and implement. For this reason an accurate, yet economical alternativ is being researched.

Lightweight Synthetic Aperture rada (LightSAR) is a high-technology, low-cos Earth-imaging SAR designed to provid scientific research on a number of topo logical issues. Some of these objectives in cluded seismic and volcanic deformatio mapping, vector ice sheet and glacier ve locity mapping, topographic mapping and surface characterization. Essentiall

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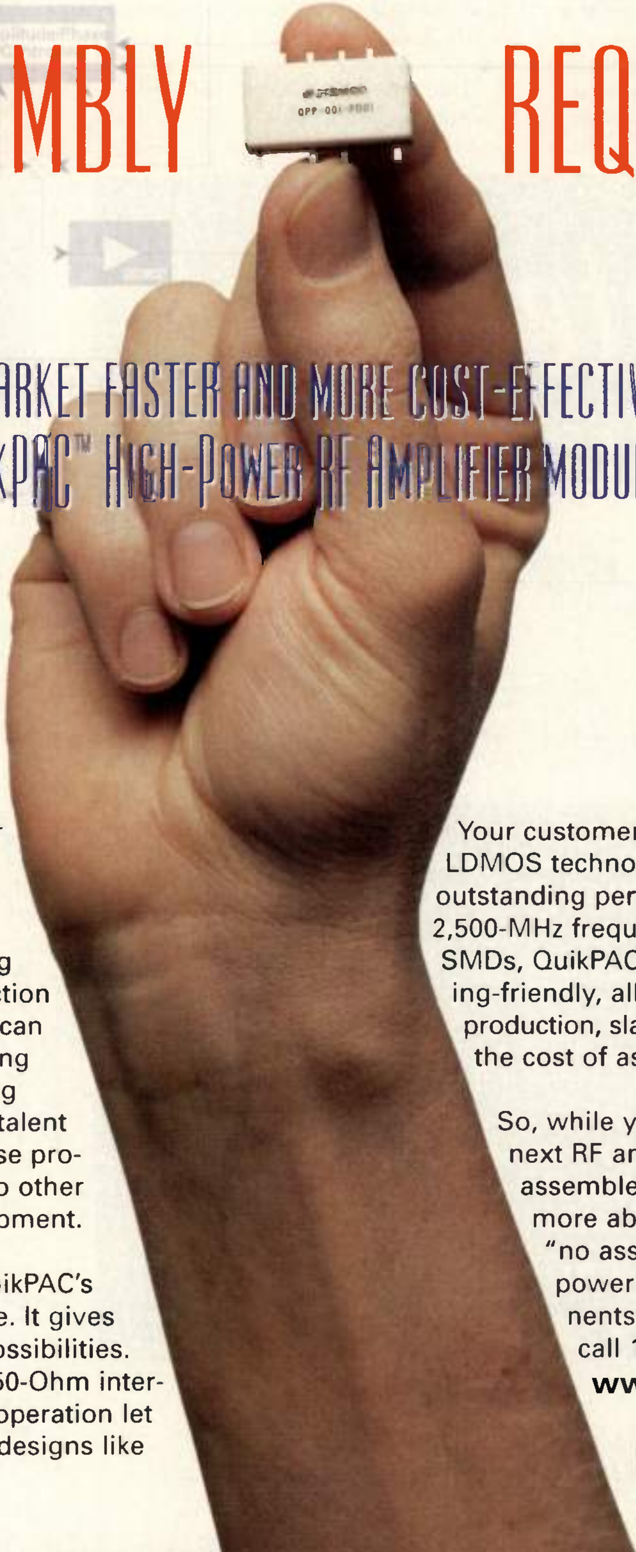

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these are objectives that can be met with repeat-pass Interferometry with single and dual (HH/VV) polarization—an objective that can generally be met with a single-frequency L-Band system.

Because many images don't need the crème de la crème SAR systems to acquire adequate data, LightSAR looks promising. Such a system is a compromise of the multifrequency, multipolarization high-end systems with adequate imaging ability and a reasonable cost to deploy.

It all comes together in GeoSAR

Perhaps one of the most promising

Figure 2. A SAR image taken from the ERS-Satellite (photo courtesy of NASA).

Figure 3. A non-SAR image (photo courtesy c NASA).

projects under current development i GeoSAR. This project is designed t generate high-resolution 3-D maps c the geology of California.

The system is a dual-frequency Inferometric system capable of map ping above, through and below the veg etation layer. Such data is important i seeing how the earth's geography ha formed under the vegetation layer. I will also be of immeasurable value i providing data for geologic seismi analysis and hazard identification.

The GeoSAR system has impressiv technical specification. It will use bot X- and P-Band frequencies, which wi allow DEMs with resolution as low a 0.5 meter.

And the direction is...

Although SAR is not entirely 1990s technology, the technological de velopments of the 1990s have played large part in enabling SAR develop ment and deployment. GeoSAR, for ex ample, will not be operational until a least early 2000.

Other SAR deployments both fror satellite and aircraft will see increase usage as we turn the century. Continue maturation of processing software an digital conversion hardware will rende mountains of acquired data manageable This will provide extremely accurat and timely information for disaste management and research, terrain re source mapping and planning, ecologica and environmental assessment, searc and rescue, security and military intelli gence, just to name a few.

The applications seem endless, an the technological opportunities shoul follow. RF

Measuring multipath in the wireless cable environment

Understanding techniques to measure multipath can get your system off to a quick start.

By Mark Kolber and Marc Ryba

Multipath is one of the most significant transmission impairments in the wireless cable environment. Multipath occurs when two or more propagation paths exist between the transmitter and receiving sites. The transmitted signal may arrive at the receiver from multiple paths exhibiting various amounts of delay and attenuation. The indirect paths may result from reflections from man-made or natural structures, repeaters or the use of multiple transmitters.

Multipath on analog National Television System Committee (NTSC) transmitted signals results in a ghost-like image horizontally displaced from the main image by an amount proportional to the reflected signal's delay. Multipath degradation is not visible in a digitally demodulated picture until the "threshold" of the digital signal is reached, resulting in a loss of demodulator lock.

With digital transmission, uncorrected multipath introduces inter-symbol interference (ISI) that results in a closure of the eye pattern, making the signal more susceptible to decoding errors. Use of an adaptive equalizer in the receiver can minimize the effects of multipath and improve system performance. Multipath outside the time range of the demodulator's adaptive equalizer is perceived as additional noise and causes degradation to the received signal-to-noise ratio (SNR). In severe cases, uncorrected multipath causes a total loss of demodulation. However, there are different methods to recognize and quantify multipath in a digital transmission system to correct such situations.

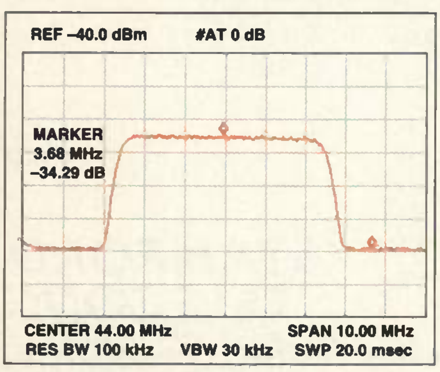

Figure 1. 64-QAM signal with Es/No = 34.3 dB.

Measuring multipath with analog

In the case of analog TV transmission, multipath can be seen on-screen as ghosting, horizontally displaced from the main image. This can be used to provide a rough gauge of both the time delay and amplitude of the multipath. For an NTSC-M system, the duration of each horizontal TV line is about 63.5 µs. Of this, about 11 µs are used for the horizontal synchronization and blanking interval, leaving about 52.5 µs for active video. Typical over-scan will reduce this value somewhat, so we can round off to 50 µs. Assuming that the multipath is less than 50 µs, the percentage of the TV screen in that a ghost is delayed can be used to estimate the multipath's delay time.

$$\text{Delay} = \text{\% of TV screen displaces} \times 50\ \mu\text{s}$$

For example, if a ghost is displaced 25% from the main signal, the multipath delay is approximately 12.5 µs.

The amplitude of the multipath can also be determined by comparing the ghost's amplitude to the original object. This can be measured using a video waveform monitor, but it is difficult to do using active video. A waveform monitor with the appropriate triggering to view only the vertical integral test signals (VITSs) would make the job easier. The magnitude of the multipath can be determined as:

$$20 \cdot \log\left(\frac{\text{\% of amplitude}}{100}\right)$$

For example, if multipath creates a ghost that is 25% the amplitude of the original waveform, it is −12 dBc, or 12 dB below the desired signal. If one or more of the transmitted channels are analog, this method can be an effective way to measure multipath.

Digital SNR vs. C/N

With digital TV transmission, uncorrected multipath does not create visible ghosting on the TV screen but rather degrades the demodulated noise margin by creating ISI. For example, a digital signal may have a carrier-to-noise (C/N) of 35 dB or better, but at the same time the SNR estimate as reported by the demodulator may be only 27 dB because of multipath. This difference is caused by the different definitions of C/N and SNR.

C/N is the ratio of carrier power to noise power. Here, noise power is taken to mean random thermal noise. C/N measurements must also specify a measurement bandwidth. It is convenient to use the symbol rate bandwidth because the measured C/N is then equivalent to Es/No (the ratio of the observed signal density to the observed noise density). Es/No and C/N in 5.05 MHz can be measured directly on the spectrum analyzer with no bandwidth factor corrections needed. For example, Figure 1 shows a 64-quadrature amplitude modulation (QAM) signal with Es/No = 34.3 dB and equivalently with the $C/N|_{5.05\ \text{MHz}}$ = 34.3 dB. To ensure that the measurement is accurate, the spectrum analyzer noise floor must be at least 10 dB below the system noise floor.

SNR or modulation error ratio (MER) as reported by the QAM demodulator, on the other hand, is a measure of the spread of the demodulated constellation points (also known as the constellation cluster variance) expressed as a noise ratio. Here, the SNR estimate includes any impairment that causes a spreading of the received constellation points from the ideal reference constellation points. Phase noise, ISI caused by multipath, as well as

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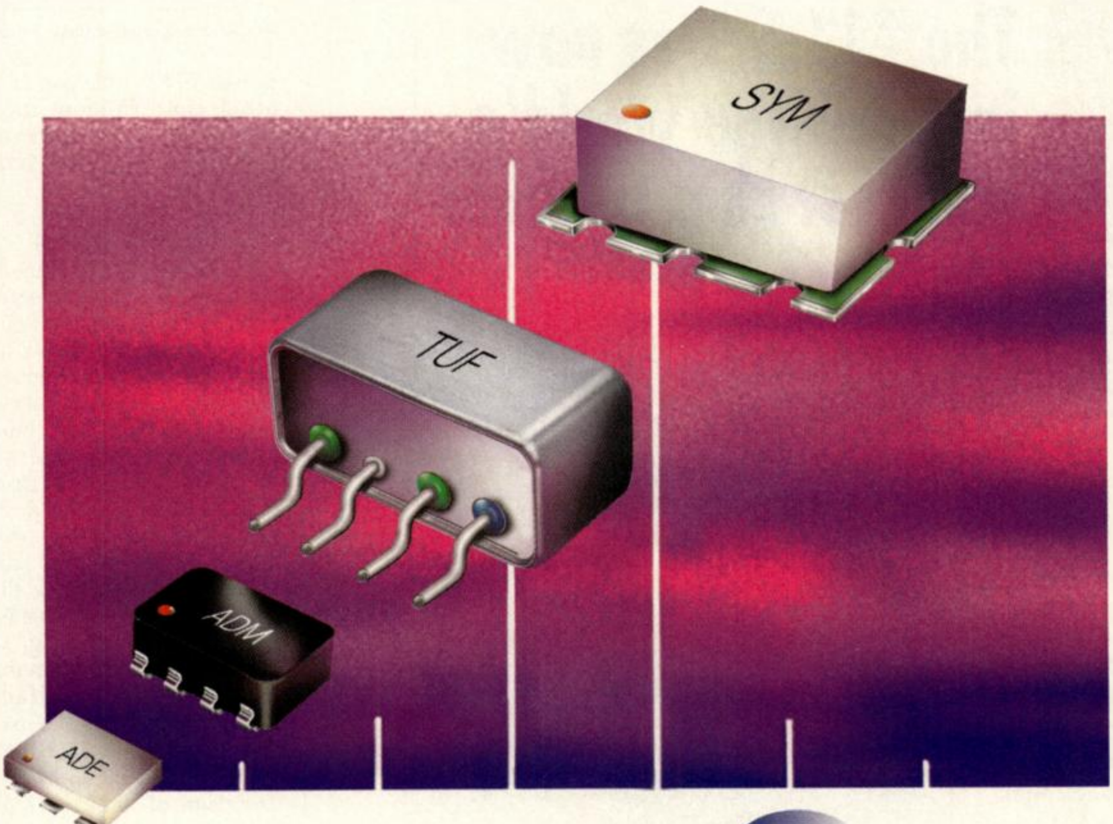

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thermal noise and other transmission impairments will all be reflected in the SNR/MER. If no impairments other than random thermal noise are present, the reported SNR/MER should be nearly equal to the input C/N. The reported SNR/MER is usually worse than the measured C/N because of other impairments besides random thermal noise existent in the signal. Phase noise and multipath are two examples of impairments that do not directly affect C/N but do affect the constellation point spread and therefore decrease the reported SNR/MER. A large discrepancy between the observed C/N and reported SNR indicates that impairments other than random thermal noise is present, and it is often multipath that is beyond the adaptive equalizer's range.

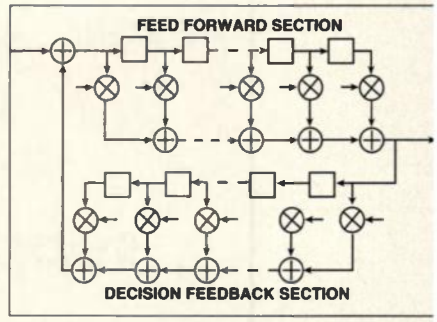

Figure 2. Adaptive equalizer block diagram.

Adaptive equalizers

Practical QAM demodulators must contain an adaptive equalizer (AE) to reduce the effects of multipath and ISI. Think of an AE as a tapped delay line that creates delayed versions of the main signal that are used to cancel out the multipath. A block diagram of a typical AE is shown in Figure 2. The AE can have only a finite number of taps and can only cancel multipath that is within its time delay range. The number of taps, the symbol rate and the tap spacing determine the time delay range. Most AE designs are either T (T = symbol rate), spaced (also called synchronous designs) or T/2 spaced (also called fractionally spaced designs). For example, a 64-QAM ITU-T J.83(B) signal with a symbol rate of 5.05 Msps has a symbol duration of about 0.2 µs. Therefore each tap in a T-spaced AE contributes 0.2 µs of delay range. With a 16-tap decision feedback equalization (DFE) AE, the maximum echo that could be corrected would be $16 \times 0.2\ \mu s = 3.2\ \mu s$. Echoes with a delay greater than this would be beyond the range of the AE and would appear as noise to the system.

Multipath occurs in two forms, pre-echoes (leading the desired signal) and post-echoes (lagging the desired signal), with the latter the most common. The decision feedback section of the AE cancels post-echoes. Usually the feedback is taken after the decision slicer function that determines the bit value. This is DFE and has the advantage of rejecting some noise. Pre-echoes can occur when repeaters are used. They can result

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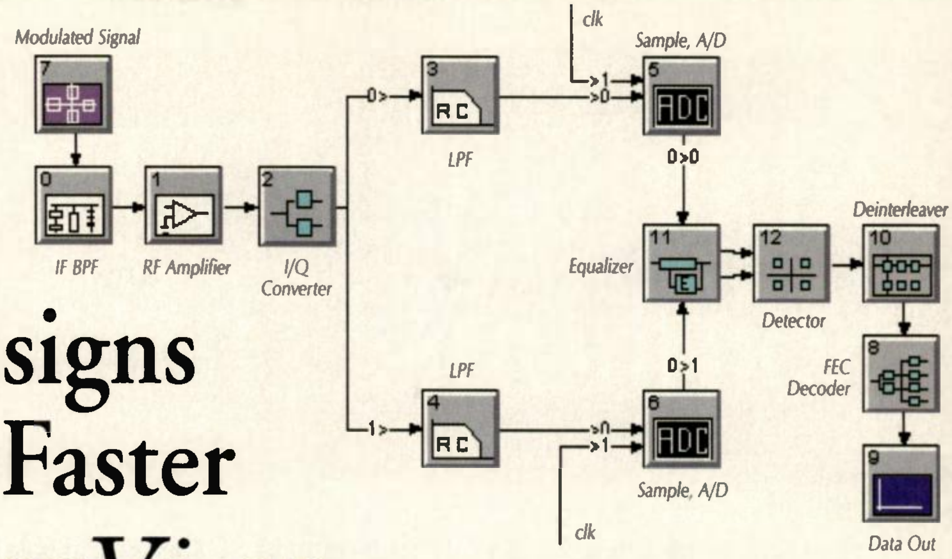

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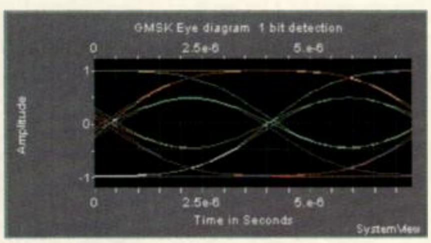
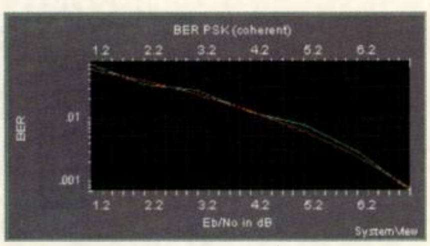
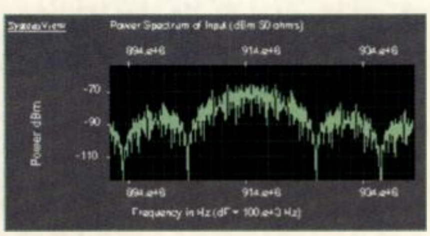


GMSK Eye Diagram (1-bit Detection)



Theoretical vs. Actual BER (Coherent PSK)



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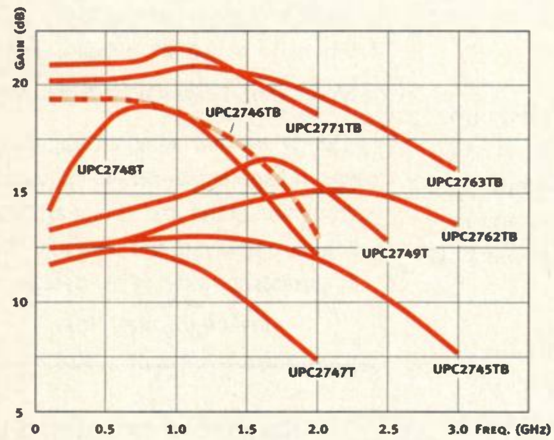

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UPC2708T	50MHz–2.9GHz	15	6.5	+9.2	26	1.0GHz
UPC2709TB	50MHz–2.3GHz	23	5	+8.7	25	1.0GHz
UPC2710T	50MHz–1.0GHz	33	3.5	+10.8	22	500MHz
UPC2711TB	50MHz–2.9GHz	13	5	-2.6	12	1.0GHz
UPC2712TB	50MHz–2.6GHz	20	4.5	-0.4	12	1.0GHz
UPC2713T	50MHz–1.2GHz	29	3.2	+0.3	12	500MHz
UPC2776TB	50MHz–2.7GHz	23	6.0	+6	25	1.0GHz
UPC2791TB	50MHz–1.9GHz	12	5.5	+1	17	500MHz
UPC2792TB	50MHz–1.2GHz	20	3.5	0	19	500MHz

3V WIDEBAND RF IC AMPS—FROM 35¢

PART	FREQ. RANGE	GAIN (dB)	NF (dB)	P_{1dB} (dBm)	I_{CC} (mA)	f_{TEST}
UPC2745TB	50MHz–2.7GHz	12	6	-3.0	7.5	500MHz
UPC2746TB	50MHz–1.5GHz	19	4	-3.7	7.5	500MHz
UPC2747T	100MHz–1.8GHz	12	3.3	-11	5	900MHz
UPC2748T	200MHz–1.5GHz	19	2.8	-8.5	6	900MHz
UPC2749T	100MHz–2.9GHz	16	4	-12.5	6	1.9GHz
UPC2762TB	100MHz–2.9GHz	14.5	7	7	27	1.9GHz
UPC2763TB	100MHz–2.4GHz	20	5.5	6.5	27	1.9GHz
UPC2771TB	100MHz–2.1GHz	21	6	11.5	36	900MHz

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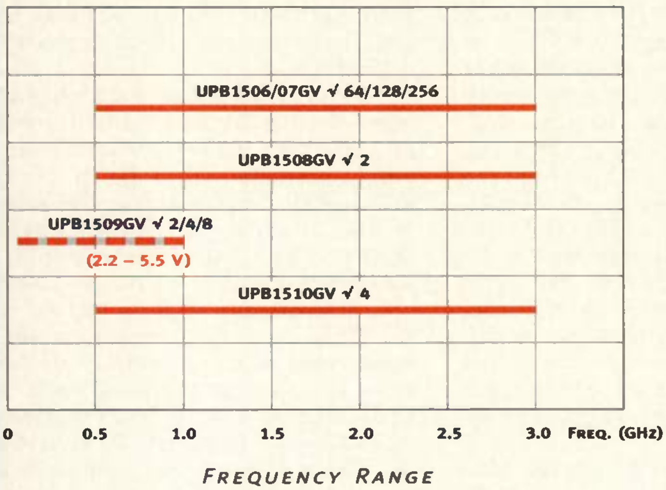

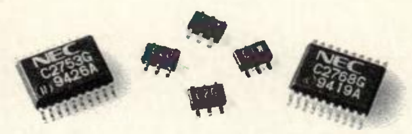

3V Frequency Converters—from 60¢

Part	RF Frequency (MHz)	I_{CC} (mA)	Conversion Gain (dB)	Output IP_3 (dBm)
UPC2756T[1]	100 - 2000	5.9	14	0
UPC2757T[1]	100 - 2000	5.6	13	0
UPC2758T[1]	100 - 2000	11	17	+6
UPC2768GR[1]	10 - 450	7	80	-17
UPC8106T[2]	100 - 2000	9	9	+1
UPC8112T[1]	800 - 2000	8.5	13	-10
UPC8116T[3]	100 - 500	4.1	6.5	—

1. Downconverter 2. Upconverter 3. AM/ASK Receiver IC

UPC8104GR
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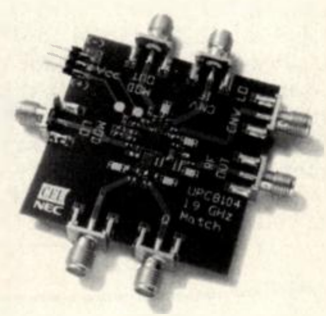

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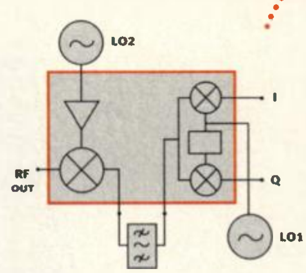

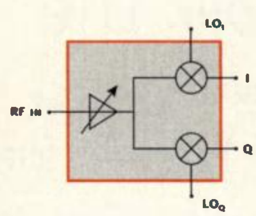

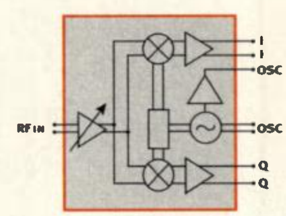

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when the stronger desired signal is received via a longer path through a repeater and the weaker undesired signal is received via a shorter path directly from the main transmitter. The feed-forward section of the equalizer (FFE) cancels pre-echoes. This portion cannot take advantage of the noise rejection properties of the decision slicer.

In both sections, the tap values or weights are adaptively adjusted based on the results of the constellation decisions. These are called decision-directed equalizers (DDE). Generally, the least mean squares (LMS) algorithm is used to adaptively adjust the tap values to minimize the spread or mean square error of the constellation points and thus achieve the highest SNR or MER. Most AEs used in digital TV applications train or adapt using the normal data signal. This is known as blind equalization. Other approaches require that a known training sequence be transmitted.

The tap values of the AE, after it has adapted to a particular echo, can be used to characterize the multipath. It is important to note that the AE taps only indicate the presence of multipath that is within the AE's time range. If multipath is present outside this time range, it will not be reflected in the tap values. In other words, the AE can only report multipath that it is capable of correcting. Multipath that is longer than the correction range of the adaptive equalizer is perceived by the demodulator as random noise and, depending on its amplitude, can sometimes be corrected by the demodulator's forward error correction [1]. However, this places an extra burden on the error-correction budget, thereby reducing the capability to correct errors caused by other impairments.

If the multipath is within the AE's time range, the value of the DFE taps will directly indicate the time delay and magnitude of the post-echo multipath. For the ITU-T J.83B example where the tap spacing is about 0.2 µs, a post-echo that is 1.0 µs in delay will activate the fifth DFE tap. The magnitude of the tap relative t the main tap indicates the amplitud of the echo. Often this can be directl displayed in dB. If the multipat falls between two tap values, th taps on either side of the echo wi activate.

The situation becomes slightl more complex in the FFE for the cas of pre-echoes. In compensating for single pre-echo, the FFE itself cre ates additional pre-echoes. The resul is that a single pre-echo activates series of FFE taps. For example, single pre-echo at –0.5 µs at –10 dB will activate the –0.5 µs tap at –1 dB. This, however, creates an addi tional echo of the echo that also ac tives the –1.0 µs tap at –20 dB. I turn creating another echo that acti vates the –1.5 µs tap at –30 dB and s on. This continues until there are n more taps and the residual is left ove For the purposes of measuring multi path, the first activated tap (in this ex ample, the tap at –0.5 µs) correspond to the actual echo. The DFE does nc suffer from this echo of the ech problem caused by the inherent feed

back in its DFE structure.

Using adaptive equalizer data to measure multipath

The DFE and FFE AE tap values can be used to measure multipath that is within the AE's time range. Figure 3 shows the tap value coefficients of an AE with 8 DFE taps and 8 FFE taps. Note that one of the FFE taps is designated as the "main tap" or "reference tap" that passes the non-reflected desired direct signal. The other 7 FFE and 8 DFE taps are available for echo cancellation. In this example, the main or reference tap has been set to a

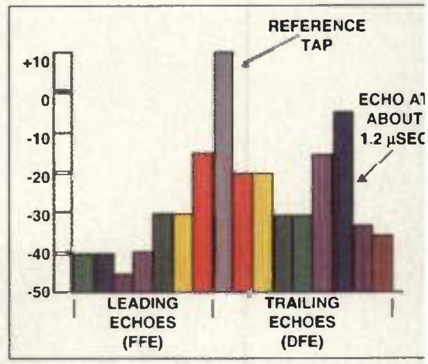

Figure 3. Adaptive equalizer tap values.

value of +10 dB. This value is arbi trary and is a function of the partic ular AE design. To cancel the multi path, the adaptation algorithm has se the sixth DFE tap to –4 dB or –14 d relative to the main tap. This indicate that an echo exits at about 1.2 µs a –14 dBc. The fifth tap at 1.0 µs is als activated at about –25 dBc. This coul be caused by a second echo at 1.0 µ but because the seventh tap is quit low, it probably means that the singl echo is actually occurring between th fifth and sixth taps or between 1.0–1. µs but closer to 1.2 µs. The –0.2 µs FF tap is activated at –25 dBc and most o the taps are also partially activated This could be caused by low-leve echoes or a slight tilt in the channe Multipath and frequency response tilt are closely related, and both are cor rected by the AE. Some AE read-out provide for a frequency response dis play that is calculated as the Fourie transform of the tap values.

Some systems that include a QAM demodulator provide for reading th AE tap values. As mentioned, this fea ture can be used to measure multipat that is within the time range of th AE. A vector signal analyzer such as HP 89441 that is equipped with th AE option can also be used to measur multipath. The AE in this instrumen can be configured for as many as 99 T spaced taps covering a time range o about –3.7 µs to +15.6 µs. The adapta tion in this type of instrument is muc slower compared to that in an actua demodulator AE, typically taking 1 seconds to acquire initial lock and a much as a minute to optimize th adaptation. Unlike the AE used in typical 64-QAM demodulator, the H 89441 uses a finite impulse respons (FIR) FFE for both the pre- and th post-echo section. Because the FI

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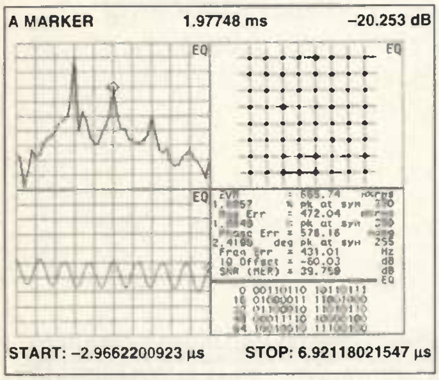

Figure 4. HP 89441 display.

filter structure has no feedback, both pre- and post-single echoes will create multiple tap responses. In the typical demodulator AE, multiple responses to a single echo occur only in the pre-echo FFE section.

Recommended settings for using an HP 89441 to measure multipath on an ITU-T J.83 B signal are given in Table 1.

Figure 4 provides an example of the QUAD display format from the HP 89441 that has been configured as described. The upper-left display is the impulse response of the AE that shows the values of the 51 taps over the range of about −3 µs to +7 µs. For this example, a single echo at +2 µs and −20 dBc was created, as is shown by the marker. Notice the main tap at 0 dBc to the left of the echo. Also notice the "echo of the echo" tap at +4 µs and −40 dBc as well as the additional "echo of the echo of the echo" at +6 µs and −60 dBc. Remember, these additional echoes are not actually present in the channel but are the result of the operation of the FFE-type AE that is used in the HP 89441, which has no inherent feedback. The lower-left display shows the frequency response of the channel based on calculating the FFT of the AE impulse response. The display is 1 dB per division and shows ripples spaced at 500 kHz with peak to valley amplitude of about 1.7 dB. The upper-right-hand display shows the 64-QAM constellation that has been demodulated by the HP89441 using the AE. The lower right hand display shows the demodulated error vector magnitude (EVM), SNR (MER) and other parameters of the demodulated signal using the AE.

Some signals may require changing the AE convergence speed factor. Although a smaller convergence number will provide a more accurate result and is more likely to converge, it will require more time. A larger convergence number will usually converge faster but may fail to converge with some signals and will give a less accurate result. A good strategy is to start out with a larger number such as 5×10^{-6} for fast initial convergence and then reduce the value for a more accu rate result.

Measuring multipath using a spectrum analyzer

In some cases, a HP 89441 or the ta values of the demodulator AE may b unavailable. Also, as mentioned, an A cannot indicate multipath that is be yond its time range. When this is th case, an ordinary spectrum analyze can be used to characterize multipat on digital signals. The spectrum of digital signal, after averaging, is essen tially flat across most of the symbo rate bandwidth. If, however, multipat is present, constructive and destructiv interference of the reflected path wit the direct path will cause ripples in th otherwise flat spectrum. The inverse o the frequency spacing of the ripples in dicates the time delay of the multipath Ripple spacing can be measured fron peak-to-peak or from null-to-null. B measuring the frequency delta betwee adjacent ripples, the time delay can b determined as:

$$\text{Time delay}\ (\mu s) = \frac{1}{\text{frequency}\ \Delta}\ (\text{MHz})$$

For example, if the ripple is observe with peaks spaced at 1.25 MHz, th time delay is 1/1.25 MHz, that equal 0.8 µs. Note that as the time delay in creases, the frequency delta betwee the ripple peaks decreases. A long-tim

delay causes many closely spaced ripples across the channel. A short-time delay causes a small number of widely spaced ripples across the channel.

To understand why multipath cause ripples in the spectrum, consider a con

...continued on page 7

• INST MODE	
Demodulation	
Demodulation Setup	
Demodulation Format	= 64 QAM
Symbol Rate	= 5.056941 MHz
Result Length	= 512 Symbols
Measurement Filter	= Root Raised Cosine
Reference Filter	= Raised Cosine
Alpha/BT	= 0.18
EQ Filter	= ON
EQ ADAPT	= Run
EQ Filt Length	= 51 symbols (this sets the time range of the AE)
Convergence	= 1e-07 (this sets the speed of adaptation SEE TEXT)
EQ Reset	= (select this to restart AE)
• AVERAGE	
Average	= OFF
• TIME	
Result Length	= 512 symbols
Pulse Stretch	= OFF
Sync Search	= OFF
Points/Symbol	= 1 (this establishes the AE as a T spaced AE)
• FREQUENCY	
Center	= (Center Freq of QAM signal)
Span	= 6 MHz
• SWEEP	
Continuous	
Sweep Mode	= AUTO
• TRIGGER	
Free Run	
Trig Holdoff	= OFF
Ext Arm	= OFF
• SOURCE	
Source	= OFF
• Res BW/Window	
Main Window	= Flat Top
• INPUT	
Input Z	= 50 or 75 Ohm (dependent on system)
Ch 1 In Im	= 50 or 75 Ohm
Ch 1 Coupling	= AC
Ch 1 Alias LPF	= IN
• RANGE	
Ch 1 Autorange	= OFF
Ch 1 Range	= (set to the minimum value that does not indicate overload)
• DISPLAY	
4 Grids Quad	
• A DISPLAY	
A Measurement Data	= Equalizer Impulse Response
A Calculate	= ON
A Data Format	= Log (dB)
A X Axis	= Linear
A Ref Lvl/Sale	
y per division	= 10 dB
• B DISPLAY	
B Measurement Data	= Channel Frequency Response
B Calculate	= ON
B Data Format	= Log (dB)
B X Axis	= Linear
B Ref Lvl/Sale	
y per division	= 1 dB
y Ref level	= 6 dB
• C DISPLAY	
C Measurement Data	= Time
C Calculate	= ON
C Data Format	= Polar Constellation
C Ref Lvl/Sale	
y Ref level	= 0
x Ref level	= 0
y per div	= 180m
Range Tracking	= ON
• D DISPLAY	
D Measurement Data	= Symbol Table/Error Summary
D Calculate	= ON

Table 1. HP 89441 configuration for measuring multipath with ITU-T J.83B signals.

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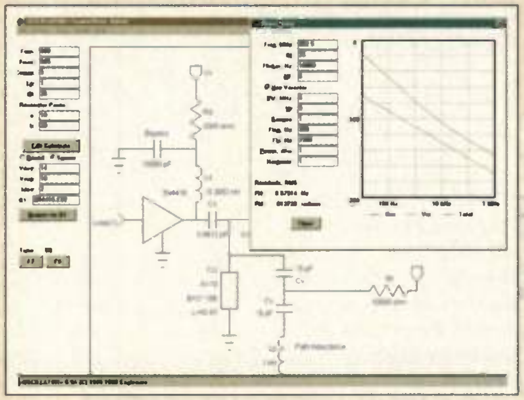

SYNTHESIS

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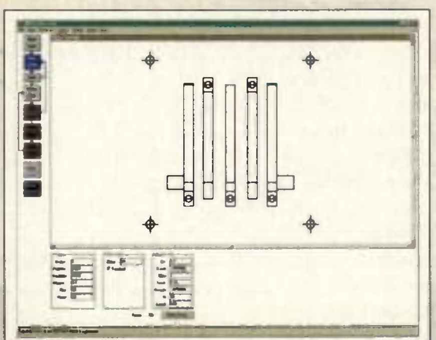

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Hairpin
Combline
End Coupled
Stepped-Z
Elliptic Lowpass
Stub Lowpass
Elliptic Bandpass
Stub Bandpass
Stub Bandstop
Edge Bandstop
Microstrip, Stripline, Stabline and Coaxial

=FILTER=

All-pole and elliptic
Top-C and -L coupled
Shunt-C coupled
Tubular
Blinchikoff flat delay
Zig-Zag
Eagleware Symmetrical
6 Group Delay Equalizers

=A/FILTER=

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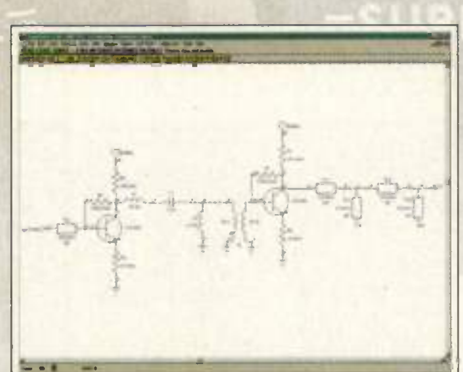
State variable (bi-quad)
VCVS
Dual amplifier
Low Sensitivity
3 Group Delay Equalizers



=MATCH=

L-C pi and L
L-C tee
T-line quarter wave
T-line single/double stub
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COMPANY NAME	PAGE NO.	READER SVC. NO.
Aerotron-Repco Systems	88	149
Amplifier Research	83	134
Analog Devices	93	128
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Artech House	93	127
AVX	93	129
Boonton Electronics	82	168
California Eastern Labs	89	153
California Micro Devices	81	166
Comtech PST	90	157
Control Concepts	93	130
Densitron Microwave	90	158
Electronics Surveillance Comp	91	163
EMCA-REMEX	80	164
EMC Test Systems	93	131
Ericsson Components	89	155
FEI Communications	91	161
Flann Microwave	86	145
Frontline Test Equipment	92	118
Gage Applied Sciences	93	124
Harmonic Software	92	116
Hewlett-Packard	83,93	133,132
Hewlett-Packard EEsof	92	117
IFR Americas	84,93	138,123
Kepco	82	167
KOA Speer Electronics	89	154
Kota Microcircuits	90	156
Linx Technologies	88	151
Lucent Technologies	90	159
Maxim Integrated Products	88,93	148,126
Mini-Circuits	84	141
Narda Microwave-East	86	143
National Instruments	92	115
Optotek	92	121
Philips Semiconductors	84	140
Piezo Technology	91	162
Pletronics	91	160
Power Trends	80	165
Princeton Electronic Systems	83	136
Programmed Test Sources	84	139
Renaissance Electronics	83	137
RF Micro Devices	84	142
Riedon	93	125
RLC Electronics	86	144
Tanner EDA	92	119
Toko America	89	152
Wavetek	92,93	120,122
Z-Communications	87	147

.continued from page 64

Echo Amplitude (dB)	Destructive Interference (dB)	Constructive Interference (dB)	Peak to Valley Delta (dB)
–0.001	–78.78	6.02	84.80
–1	–19.27	5.53	24.81
–2	–13.74	5.08	18.81
–3	–10.69	4.65	15.34
–4	–8.66	4.25	12.91
–5	–7.18	3.88	11.05
–6	–6.04	3.53	9.57
–7	–5.14	3.21	8.35
–8	–4.41	2.91	7.32
–9	–3.81	2.64	6.44
–10	–3.30	2.39	5.69
–11	–2.88	2.16	5.03
–12	–2.51	1.95	4.46
–13	–2.20	1.75	3.96
–14	–1.93	1.58	3.51
–15	–1.70	1.42	3.12
–16	–1.50	1.28	2.78
–17	–1.32	1.15	2.47
–18	–1.17	1.03	2.20
–19	–1.03	0.92	1.96
–20	–0.92	0.83	1.74
–21	–0.81	0.74	1.55
–22	–0.72	0.66	1.38
–23	–0.64	0.59	1.23
–24	–0.57	0.53	1.10
–25	–0.50	0.48	0.98

able 2. Tabulated echo amplitudes.

inuous wave (CW) signal on 2,500 IHz, for example. Also consider that ıultipath is present with a delay of 1 so that the reflected path signal is elayed by exactly 2,500 cycles of the ignal. Because the delay corresponds o exactly 2,500 cycles, the peak of the elayed signal carrier will correspond o the peak of the direct signal, but isplaced by 2,500 cycles. Because the eaks of the two signals line up, this vill result in constructive interference, and the amplitude of the signal t 2,500 MHz will be increased. The ituation is the same at 2,501 MHz; he signals differ by exactly 2,501 cyles, so constructive interference still esults. Now consider the situation at ,500.5 MHz. Here, the signals differ y 2,500.5 cycles. The peak of the relected signal arrives with the trough f the direct signal. This causes detructive interference, so the amplitude at 2,500.5 MHz is reduced. This attern repeats across the entire hannel, resulting in constructive interference every 1 MHz and destructive interference at the ½ MHz points etween. When a QAM signal is ransmitted, ripples in the spectrum esult.

Short multipath (less than 0.2 μs) reates less than one ripple in each 6 IHz channel. This short multipath caused by relatively close objects has the effect of attenuating or amplifying entire channels; i.e., it creates ripples across the MDS band rather than ripples across individual channels. This can actually be helpful to those channels that are increased in level, but it is harmful to those that are attenuated. If the attenuation brings the channel below the noise threshold, it will not be receivable despite the AE.

The amplitude of the multipath can be determined by measuring the magnitude of the amplitude ripples. In this case, we will measure the difference in amplitude between the peak and the valley. If, for example, the multipath is –12 dBc, this corresponds to a voltage ratio of 0.25. The constructive interference results in a peak level of 1 + 0.25 = 1.25 or +1.9 dB, and the destructive interference results in a valley of 1 – 0.25 = 0.75 or –2.5 dB. This is a peak to valley of 1.9 + 2.5 = 4.4 dB. The following equation can be used to calculate the magnitude (in dB) of the echo with respect to the main signal from the destructive (negative) trough and constructive (positive) peak delta from the spectrum analyzer:

$$\textit{Echo Level}\,(dB) = 20\log\left[\frac{10^{\left(\frac{\text{Pk-trough}\Delta}{20}\right)}-1}{10^{\left(\frac{\text{Pk-trough}\Delta}{20}\right)}+1}\right]$$

where Pk-TroughΔ = Peak to trough delta in decibels read from spectrum analyzer. Table 2 shows tabulated results corresponding to the peak to trough delta on the spectrum analyzer.

Figure 5 shows a simulation of a normal 64-QAM spectrum along with a simulation of a spectrum effected by multipath with a 0.8 μs echo at 12 dB below the main signal. Note the frequency spacing between the peaks is 1.25 MHz, and the peak to valley amplitude is about 4.5 dB. In practice, it can be difficult to see and measure small ripples on a QAM spectrum caused by the normal random variations in the signal. These random effects can be reduced by the use of the averaging function on the spectrum analyzer, allowing the ripples to be seen and measured. The span and resolution bandwidth must also be set to an appropriate value to see the ripples. It is useful to set the spectrum analyzer resolution and video bandwidth modes to AUTO. These parameters will then automatically track as the SPAN setting is varied when looking for various delays of multipath. Once ripples have been found, the span, resolution bandwidth, and amplitude settings can be varied as needed to make an accurate measurement. Table 3 shows some suggested spectrum analyzer settings useful for searching for multipath of various delays.

The spectrum analyzer plots in Figures 6 and 7 show examples of rip-

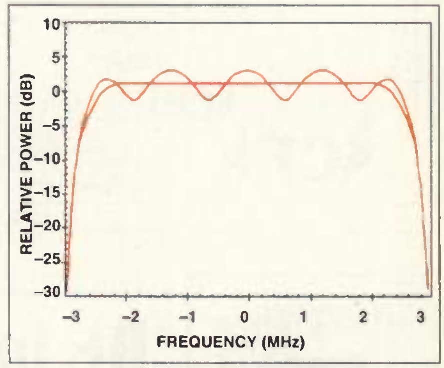


Figure 5. Multipath with 0.8 μs delay at –12 dBc.

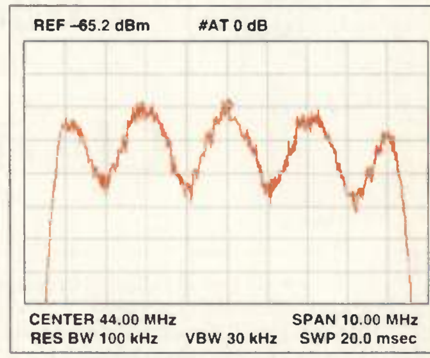


Figure 6. Spectrum analyzer plot with 0.8 μs echo at –12 dBc.

Multipath Time Delay	Frequency Spacing of Peaks	Span	RBW/VBW
0.2 to 1 μs	5 to 1 MHz	10 MHz	100 kHz / 30 kHz
1 to 10 μs	1 MHz to 100 kHz	1 MHz	10 kHz / 10 kHz
10 to 100 μs	100 to 10 kHz	100 kHz	1 kHz / 1 kHz
AVERAGE ON = 100 AMPLITUDE 2 dB/ DIV			

Table 3. Suggested spectrum analyzer setting for searching for multipath.

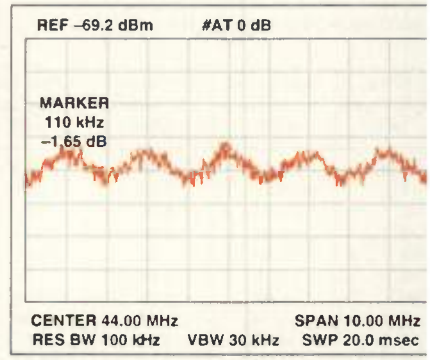


Figure 7. Spectrum analyzer plot with 5 µs ech
at –22 dBc.

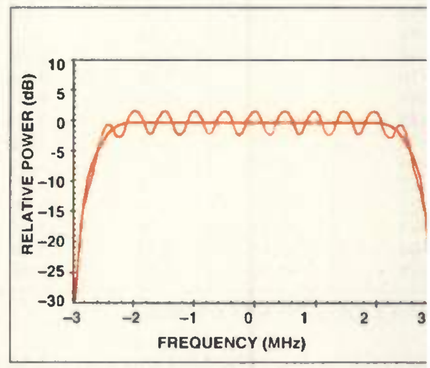


Figure 8. Simulation of a single 2 µs echo at –1
dBc.

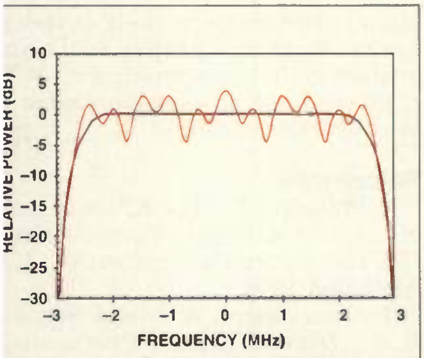
ples on a 64-QAM spectrum caused b
multipath.

Multiple echoes

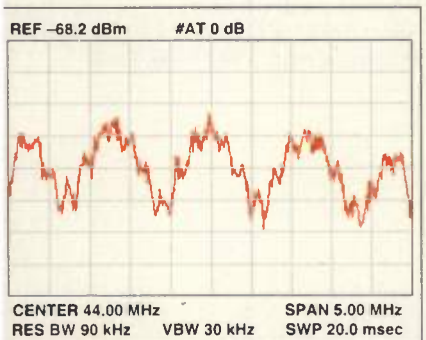
In general there can be more tha
one echo at any given receiving loca
tion. On an analog signal, multipl
echoes are seen as multiple ghosts, an
each can be measured separately. Wit
digital transmission, the AE comper
sates for each echo and the DFE ta
values can be used to resolve multipl
post echoes within the AE time range
Because the FFE itself creates add
tional echoes, it can be difficult to di
ferentiate between these additiona
echoes and multiple pre-echoes in th
propagation path.

It can also be difficult to interpre
the spectrum analyzer ripples whe
multiple echoes are present. Figure
shows a simulation of a 64-QAM spec
trum for the case of a single 0.8 µs ech
at –12 dBc. Figure 8 shows a single
µs echo at –12 dBc. Although the resul
is simply the superposition of the tw
echoes, it can be difficult to try to recog
nize and separate the two patterns a
seen in Figure 9.

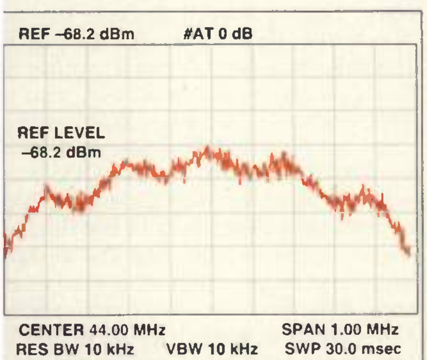
If there are only two echoes and the



igure 9. Simulation of double echo: 0.8 μs @ 12 dBc and 2 μs @ −12 dBc.



igure 10. Multiple echoes: 0.8 μs at −12 dBc and μs at −22 dBc.



igure 11. Zoom view of 5 μs echo from Figure 0.

re fairly widely spaced in time, it is ossible to recognize and measure the wo patterns, demonstrated in Figures 0 and 11. Both the 0.8 μs and the 5 μs cho can be measured from the spectral lots below. Zooming in on the 5 μs cho is required to obtain an accurate neasurement.

If there are more than two echoes nd they are not spaced widely in ime, it can be difficult to recognize he individual ripple patterns. The disrete Fourier transform (DFT) is a nathematical technique that can be used to reveal repetitive patterns. The fast Fourier transform (FFT) is a computer algorithm that is used to calculate the DFT in an efficient manner. The spectrum analyzer data can be captured, and an FFT algorithm can be used to resolve the individual echoes. The process of taking an FFT of a spectrum is called Cepstral analysis [2].

Conclusion

Various techniques for measuring multipath have been examined. The techniques presented can be used by wireless cable operators to characterize signal conditions in their coverag areas. With the proper tools an analysis techniques, multipath can b minimized and possibly eliminated a the subscriber's home. **RF**

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Novel designs for elliptic bandstop filters

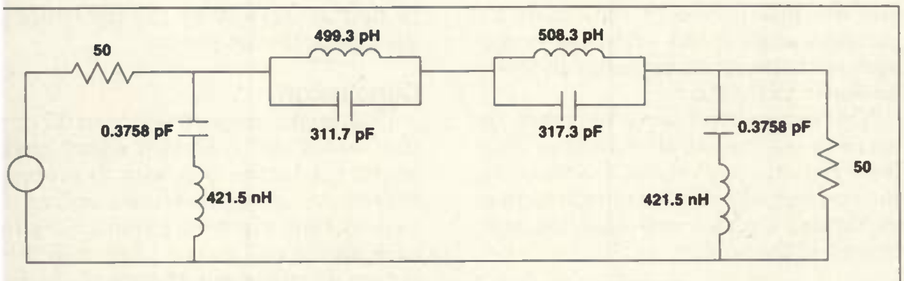
A new way to design elliptic bandstop filters may lessen frustrations.

By Philip R. Geffe

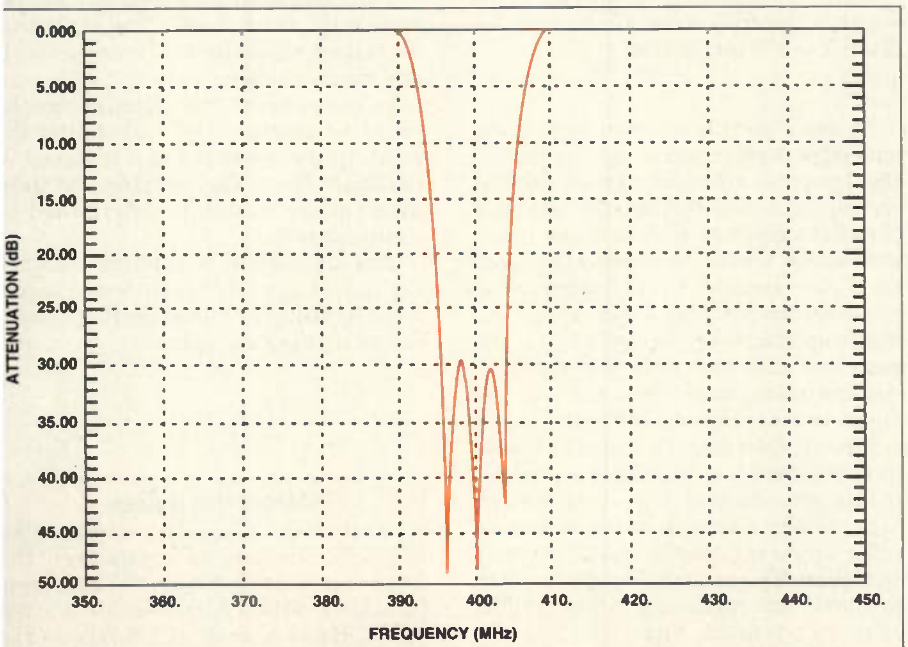
Designers of RF bandstop filters are often frustrated by the discovery hat their numerical design cannot be uilt as a practical lumped element lter because of extreme element pread.

The design of a 50 Ω bandstop filter centered at 400 MHz, with a bandwidth of 20 MHz, illustrates this point. If an elliptic lowpass prototype having three ladder branches, 0.1 dB passband ripples and a 30 dB stopband is chosen, it is apparent that no choices other than those already made are available (not counting the dual circuit, which offers no improvement). Every calculation is determined in advance, so that only one final design is possible. This design is shown in Figure 1, with the simulated ideal frequency response in Figure 2.

If this were a bandpass filter, various transformations would be available to make the circuit feasible, but that is not available here. The design is totally inflexible. Note that the element spread for both inductors and capacitors is an appalling figure of 844. At this frequency, inductors that are near 20 nH are desired, but the shunt branches of the filter use inductors that are more than 400 nH, and the series branch coils are less than 1 nH. Any coils built with these values would have a poor Q at 400 MHz.



igure 1. Conventional bandstop filter.



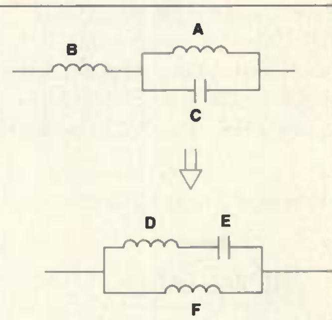
igure 2. Bandstop filter response.

Making transformations possible

Transformations that are desirable—

$$D = B + C \qquad F = C\left(1+\frac{C}{B}\right) \qquad E = \frac{A}{\left(1+\frac{C}{B}\right)^2}$$

Figure 3. Shunt circuit.



$$D = B\left(1+\frac{B}{A}\right) \qquad F = A + B \qquad E = \frac{C}{\left(1+\frac{B}{A}\right)^2}$$

Figure 4. Series circuit.

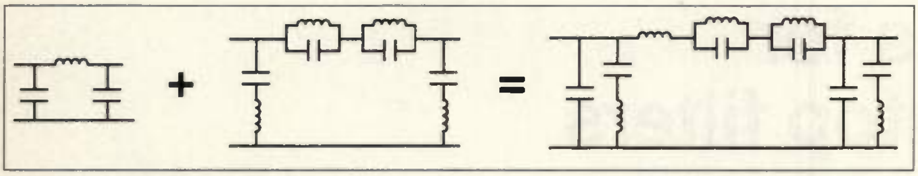


Figure 5. Example of interleaving.

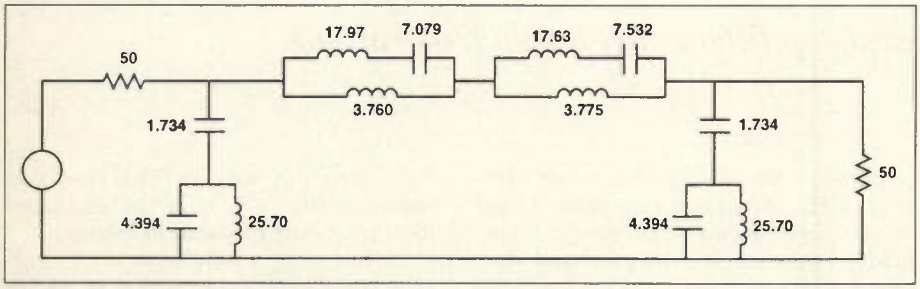


Figure 6. Filter with low inductance spread.

because they are useful in other circuitry—are the dipole transformations of Figures 3 and 4. Observe that they both require two-terminal networks containing three elements rather than two. This suggests combining the notch circuit with a lowpass or other filter circuit to obtain three-element dipoles. This method is illustrated in Figure 5: a process called "interleaving." This means that two ladders of the same size are interleaved by combining the corresponding shunt branches in parallel, and the series branches in series, as shown in the figure.

This process will serve to embed the notch in the passband of another filter. Now that three-element branches are in the filter, the dipole transformations of Figures 3 and 4 will apply to every branch of the ladder.

Optimizing the element spread

Now let y = inductance spread, and x = bandedge frequency of the lowpass filter. Y is a function of x:

$$y = f(x)$$

To evaluate this function for a given bandedge frequency, x, begin by finding the lowpass element values for the chosen value of x. Because the bandstop element values of Figure 5 are fixed, writing the lowpass element values into the figure completes the design. Now, partition the lowpass series coil of the bandstop filter into two equal parts, and associate each part with one of the series antiresonances. Then perform the dipole transformations of Figures 3 and 4. Finally, calculate the element spread thus obtained (i.e., the function value). If this procedure is performed enough times to plot a smooth curve, it will become apparent that it has a definite minimum. The bandedge frequency that produced this minimum is the optimal value for a practical filter.

Applying this procedure to the pre-

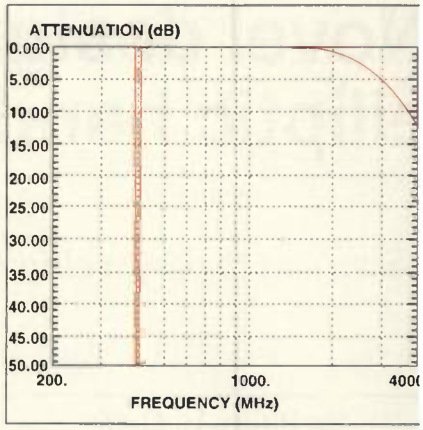


Figure 7. Attenuation of low spread filter.

vious numerical example gives the de sign of Figure 6, with its response show in Figure 7. The inductance spread is re duced from 844 to 6.8, and the capac tance spread is reduced from 844 to 4.: Although the upper passband is no limited, it appears to be wide enough fo many applications. It would, of course be limited anyway by the parasitics o the circuit, though less so.

Conclusion

One might suspect that partitionin the series coil into two equal part might not be the best way to proceed Maybe using multivariable optimiza tion to find the best partitioning fo each series coil in the filter would b better. It turns out that a slight im provement can be achieved this way but the largest improvements obtaine were only about 3–4%. The algorithm described previously is more reliabl and much simpler.

It is apparent that similar result could be obtained by embedding th notch in the passband of a highpass o bandpass filter. The usefulness of thes alternatives would be determined b circumstances.

The numerical minimum obtaine for the design of Figure 6 was calcu lated by using a Golden Section search not by plotting a graph. RF

About the author

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ECL line receivers used as amplifiers: How do they work?

The ECL differential line receiver has been used in amplifier and limiter applications. But did you know about its analog capability?

By Tom Balph
and Bill Morgan

DEVICE	SUPPLY VOLTAGE	STAGES /RCVR	GAIN @ 100 MHz (dB)	INPUT RESISTORS	RCVRS/ PACKAGE	PACKAGE
10H116	5.2	1	17	NO	3	PLCC 20
10E116	5.2	1	19	YES	6	PLCC 28
10EL16	5.2	2	30	YES	1	SOIC 8
100LVEL16	3.3	2	29	YES	1	SOIC 8
100LVEL17	3.3	2	29	YES	4	SOIC 20

Table 1. Several line receiver characteristics.

Figure 1 shows a circuit schematic and symbol of a traditional emitter coupled logic (ECL) line receiver. The receiver is a simple emitter-coupled current switch with emitter-follower output buffers to provide low-impedance drive. Resistors RB are sometimes not shown because they offer low-impedance (50–75 W) and are present to keep the input transistor base complex impedance a positive real value over frequency and switching conditions. An ECL line receiver is normally specified with VEE @ –5.2 Vdc and VCC @ Gnd, but it is not uncommon to operate an ECL with VCC @ +5.0 Vdc and VEE @ Gnd (commonly called positive-ECL or PECL).

Although esentially the same circuit, more modern devices will vary somewhat from the traditional model. Perhaps the most significant deviation is that some receivers have two DC-coupled stages (shown symbolically in Figure 2). The net result is higher gain as well as important implications on biasing techniques because of the DC-coupling. Other deviations can include devices rated for 3.3 VDC operation, better common mode range, fewer receivers per package (lower cost and smaller package) and better electrostatic discharge (ESD) protection with input resistor bias and pulldown networks.

With the model shown in Figure 1, which has no input resistor network, any unused line receiver in a package should have one input tied to the provided V_{BB} bias voltage. This supplies base current to one transistor of the switch. If this is not done, the current source transistor will be starved for current that loads the internal V_{CS} voltage source and disrupts proper operation of the rest of the line receivers. Alternately, the input resistor networks on newer devices provide a known bias condition to any line receiver not in use. The switch current source is not disrupted, and the unused line receiver is in a known condition. Figure 2 also shows a typical bias network with resistor values nominally 75 kΩ. Perhaps the main implication for RF applications is the impact on input impedance.

Table 1 provides a listing of several available ECL line receivers.

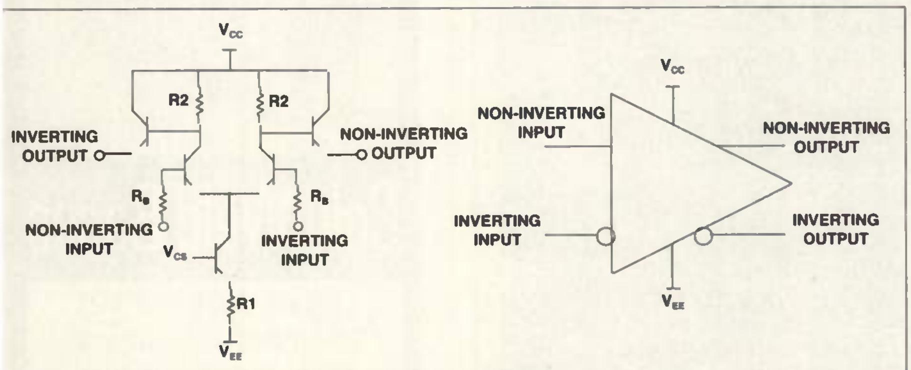

Figure 1. Simple circuit schematic and symbol of a traditional ECL line receiver.

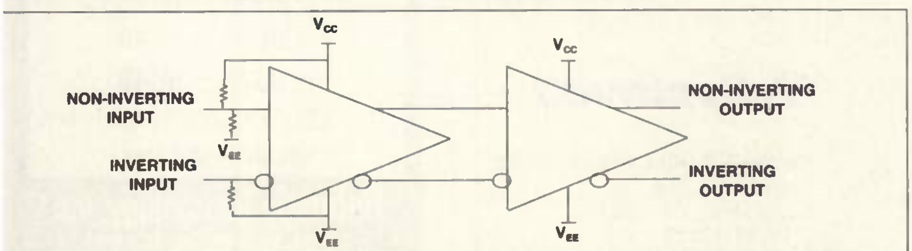

Figure 2. More recent line receivers can be comprised of two DC-coupled stages and include input bias networks.

Line receiver AC test configuration

To test the response curves of the ECL line receivers, a network analyzer was used in the configuration shown in Figure 3a. The analyzer output drove a 20 dB pad, connected to the device-under-test (DUT). The output of the DUT drove a 6 dB pad that was in line with the analyzer input. The pads were required to put the test voltages in the desired range.

The test circuit for the DUT is shown in Figure 3b. The input signal from the 20 dB pad is terminated with a 50 Ω load and AC-coupled to the input of the re-

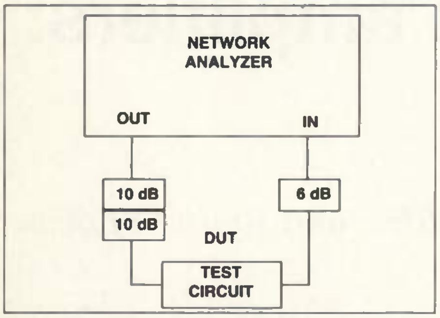

Figure 3a. ECL line receiver test configuration.

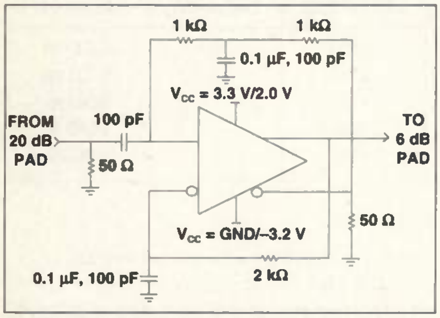

Figure 3b. Test circuit for ECL receiver.

ceiver under test. The non-inverting receiver output drives the 6 dB output pad.

The line receiver requires a proper biasing network as shown in Figure 3b for AC operation. Because of the DC-coupling of the two-stage receivers, feedback is required in the bias scheme to overcome the input differential offset voltage so that the line receiver can amplify low level signals. The bias network has a 2 kΩ feedback resistor from the non-inverting output to the inverting input, and two equivalent 1 kΩ resistors in series from the inverting output to the non-inverting input. On a DC basis, this differential feedback biases the amplifier in the center of the linear region and overcomes the input offset voltage. To allow AC operation, the AC feedback is bypassed by capacitance to ground at the inverting input and a similar network at the junction of the 1 kΩ feedback resistors. Note that the bypass networks consist of a 0.1 μF and a 100 pF capacitor in parallel. Finally, a 50 Ω resistor to ground on the inverting output is added to balance the 50 Ω loading caused by the dB pad on the non-inverting output.

The supply voltages in Figure 3b als require some comment. When testin ECL, the power supply voltages ar split to allow the emitter-follower out puts to drive 50 Ω to ground. In a 5. VDC supply situation, V_{CC} is +2.0 VD and V_{EE} is –3.2 VDC (equivalent to normal total supply voltage of 5. VDC). For the 3.3 VDC supply devices V_{CC} is 3.3 VDC and V_{EE} is ground. Th power supply voltages were altered a required by the DUT.

Test results

Using the test configuration discusse previously, curves were run for the lin receivers listed in Table 1. Both lo level input voltage gain and phase angl versus frequency were obtained. Figur 4 shows low-level gain versus frequency and Figure 5 shows phase angle versu frequency. Observe that the two-stag devices have higher gain than the tw single-stage devices. Also, most of th later technology devices have usabl ıandwidth as high as 0.9–1.0 GHz.

Conclusion

Since its inception, the ECL line reeiver has been useful as an economical ow-level amplifier and limiter. The newer levices provide higher gain (25–30 dB) ınd bandwidth out to the GHz region. For nany applications, the 10LVEL16 and he 10EL16 provide the best solution with ı single amplifier in an SOIC 8 package; he main difference being the need for a 3.3 Vdc vs. a 5.0–5.2 Vdc power supply. All of these devices provide 50 Ω drive caıability (with a maximum of about 22–25 nA drive). The user must note that the ECL outputs are limited to about 600–800 nV swings depending on the device.

For users familiar with line receivers ıs amplifiers, be sure to note the discusıion about the bias network used in the ests. Older bias techniques that do not ıse feedback will not work well with the wo-stage line receivers. The offset voltage of the first stage is amplified by the gain of the second stage such that simple nonfeedback biasing is not suitable. RF

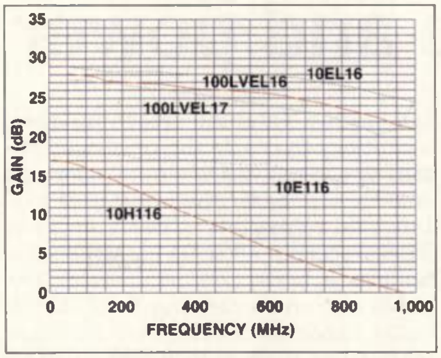

Figure 4. MECL voltage gain vs. frequency at low level input voltage (1.8 mVrms).

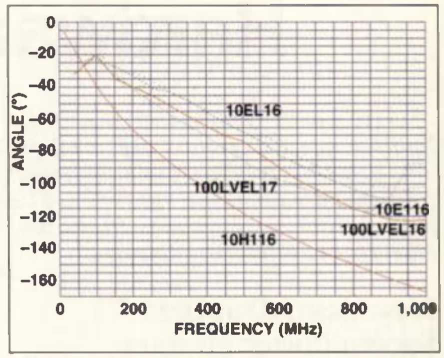

Figure 5. MECL phase angle vs. frequency high or low level.

About the authors

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Bill Morgan, recently retired, spent 33 years at Motorola as an RF lab technician. He was heavily involved in the evaluation and applications support of PLL and frequency synthesizers. Other experience includes significant RF small signal device evaluation and support.

All the devices mentioned in this article are available from Motorola.

Hardware and power products

Each month, the product focus highlights a specific area of RF products and provides product announcements from that area in a special section. This month, the product focus highlights hardware and power technologies.

Advanced dielectric for multilayer circuits

The MULTISYST 7061D is an advanced thick-film multilayer dielectric for the construction of multilayer circuitry. The 7061D dielectric is designed to meet the needs of increasingly complex multilayer circuits, providing dense and hermetic film at a fired thickness of 40 microns. This dielectric is compatible with a wide range of gold conductors as well as silver and palladium-silver conductors. The 7061D dielectric represents the most advanced combination of high-temperature glass chemistry in conjunction with a patented screening agent technology to enable the consistent processing of 8 mil vias while still offering a smooth surface for the printing of fine-line gold connectors. Closely matched TCE properties allow for construction of complex hybrids, and the dielectric provides manufacturers the flexibility to build complex and low-cost packages with either silver or gold connectors for the inner and top layers.
EMCA-REMEX
INFO/CARD 164

Dual output power modules for DSPs

The PT6920 series of dual output, integrated switching regulators (ISRs) is designed to power the latest generation of dual voltage digital signal processors (DSPs) including Texas Instrument's C6000 series. These dual 25 W power modules are available wit 3.3 V/6 A and 2.5 V/2.2 A fixed outpu voltages. The second output can also b selected for 1.8 V to accommodate th next generation of DSP chips. Both out put voltages are independentl adjustable with an external resisto with the second output voltag adjustable to as low as 1.2 V. The 5 input voltage makes it easy for inclu sion into existing power systems Housed in a 23-pin SIP package, th PT6920 series is available in surface mount, vertical and horizontal through hole configurations. The series is price at $24.20 in quantities of 1,000.
Power Trends
INFO/CARD 165

Integrated dual supply power management circuit

The CMPWR-100 is a fully integrate

ual supply power management circuit. Jnlike other integrated products that equire an external switch, the CMPWR-00 integrates the regulator, switch and ontroller into a single 8-pin SOIC package, eliminating the need for additional omponents. This component integrates ll the functions required to deliver 200 nA output current from either an onhip 5 V to 3.3 V LDO regulator or from 3.3 V auxiliary supply via an internal witch. The CMPWR-100 is priced at 95 ents in quantities of 10,000.
California Micro Devices
INFO/CARD 166

.inear, low-noise power supplies

The ATE-DMG series of 1,000 W nstrument-grade, linear low-noise ower supplies feature a built-in IEEE 88-2 bus interface for full talk-listen est-system interact ability. These ower instruments respond to both the SCPI command set and to analog nputs that range from DC to high audio signals. A 24-key keypad augments the rotary voltage and current controls and a two-line, 16-character liquid crystal display (LCD) replaces the meters. This display shows both the settings and actual voltage and current readings. The output ripple and peak-to-peak noise are around 100 μV root mean squared (RMS). The source-load effects are less than 0.001% with 12-bit digital resolution. Their fast-mode bandwidth also allows users to operate them as true current stabilizers with fast load recovery. The series is priced from $3,868–4,100.

Kepco
INFO/CARD 167

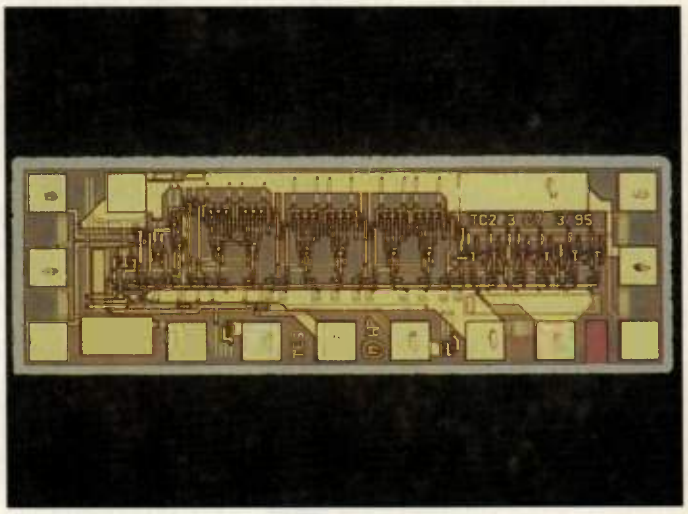
RF peak power meter for wideband modulation formats

The 4530 series RF power meters can make high-speed peak and continuous wave (CW) power measurements from 10 kHz to 40 GHz, depending on the sensor, across a 20 MHz video bandwidth. Designed for measuring signals based on wideband modulation format the 4530 series has the capability having two independent channels th allow a variety of measurements to b made, including maximum power, pea power, average power, peak-to-averag power ratio, dynamic range, minimu power, CW power and RF voltage. I addition, the peak power meter pr vides channel math functions that allo the sum ratio to be displayed, as well the difference between channels between a single channel and a refe ence measurement. The meter als includes statistical analysis feature such as histograms and cumulative di tribution functions (CDFs), so it ca conduct fast analysis of complex signa including code-division, multiple acces (CDMA) and high-definition televisio (HDTV). The peak power meter can als perform time domain power measure ments for global system for mobile con munications (GSM) and time-divisio multiple access (TDMA) signals.

Boonton Electronics
INFO/CARD 168

irst microwave ICs using aAs HBT technology available

A series of eight mi-rowave integrated circuits Cs) feature indium-gal-ium-phosphide (InGaP)-mitter, gallium arsenide GaAs) heterojunction ipolar transistor (HBT) echnology. The benefits of IBT technology include the otential for high gain per tage, operation from a ingle power supply voltage, ow phase noise, high lin-arity and good uniformity nd repeatability to ensure igh production yields. The ight products include six rescalers, three operating at 60 mA current in low output power mode (one divide-by-two, one divide-by-four and one divide-by-eight) and three operating at 30 mA current in low power mode, and two Darlington-feedback HBT amplifier chips, one having a DC–15 GHz frequency range and the other a DC–20 GHz frequency range. Prices for the ICs are $27.35 for the prescalers and $15.10 and $21.40 for the amplifiers, respectively, in quantities of 1,000–2,499.
Hewlett-Packard
INFO/CARD 133



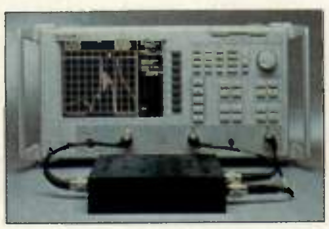
mplifier meets ellcore specification

The 15T4G18 travelling vave tube (TWT) mi-rowave amplifier tests to he higher frequency radi-ted RF requirements of he Bellcore specification or telecommunications. The amplifier has a fre-uency response from .2–18 GHz and offers 15 W ninimum output standard, .0 W minimum in a choice

of three low-harmonic sub-bands. A digital front panel hows extensive system tatus information that can be accessed (like the sub-bands) through a series of menus using soft keys, or remotely via the built-in EEE-488 interface.
mplifier Research
NFO/CARD 134

Vector network analyzers

The MS462x family of RF vector network analyzers is the first such instrument to



incorporate noise-figure measurement capability. Features include a tune mode that maintains full 12-term calibration while optimizing sweep speed and fully error-corrected measurements that can be made at sweep speeds as high as 150 µs per point. Four models are available. Two are equipped with non-reversing transmission.reflection test sets with frequency ranges from 10 MHz to 3 GHz and 10 MHz to 6 GHz. The other two incorporate reversing test sets for automatic forward and reverse S-parameter measurements from 10 MHz to 3 GHz and 10 MHz to 6 GHz.
Anritsu
INFO/CARD 135

Fixed frequency and smart synthesizer

The PSF-2510 fixed frequency synthesizer for wireless applications consists of a basic frequency synthesizer and an internal microcontroller that is used to program the synthesizer's high-performance phase lock loop (PLL) chip. The synthesizer is suitable for applications ranging from fixed frequency operation to other complex modes without the need of external control. The synthesizer operates at a frequency of 2,510 MHz having a cur-

rent of less than 38 mA with a 5 V supply. The phase noise of the synthesizer is –105 dBc/Hz at a 10 kHz offset. The synthesizer can be built using any voltage-controlled oscillator (VCO).
Princeton Elec. Systems
INFO/CARD 136

Boost channel capacity

This multichannel, high-power transmitter combiner

can be configured in nine combinations covering a frequency range of 850–870 MHz with no tuning required. Packaged in a standard EIA 3.5″ high, 19″ rack mounting, the unit has integral fan cooling for as much as 400 W dissipation with wiring provided for 18–70 VDC operation. Designed for either code-division, multiple access (CDMA) or time-division, multiple access (TDMA) applications, typical performance for the unit features greater than 70 dB isolation between transmitters with less than 7.0 dB insertion loss for the four-way and 3.8 dB insertion loss for the two-way.
Renaissance Electronics
INFO/CARD 137

TEST EQUIPMENT

Portable spectrum analyzer designed for TV/wireless

The 2398 portable spectrum analyzer combines high performance and portability in an instrument. Designed for use in wireless infrastructure or television broadcast field-testing or for use in laboratory environments, features include a frequency range of 9 kHz to 2.7 GHz. Other features include a split screen mode to allow for simultaneous viewing of two display functions, full marker capabilities and the ability to store traces and setups internally. The 2398 is priced at less than $7,000.
IFR Americas
INFO/CARD 138

Frequency synthesizer with ultra-fine resolution

The PTS 620FR frequency synthesizer features ultra fine resolution. It has an operating range of 1–620 MHz, with a choice of resolution from 100 kHz down to 1 mHz. Output level is +3 to +13 dBm, single sideband (SSB) phase noise is –100 dBc at 100 Hz and –120 dBc at a 10 kHz offset, spurious outputs are –70 dBc and frequency switching speed is 5–10 microseconds. Available with remote parallel BCD control, the unit operates from an internal 10 MHz or an external 5 or 10 MHz standard. The price is a function of resolution. A typical version with 1 mHz resolution is priced at $10,375.
Programmed Test Sources
INFO/CARD 139

Evaluation board supports wireless home networks

This wireless local area network (WLAN) frequency hopping spread spectrum (FHSS) radio evaluation board targeting home network applications supports data rates as high as 2 Mbps with a range of more than 300 meters. Designed for designers developing applications based on the HomeRF shared wireless access protocol-cordless access (SWAP) specification, the board is compliant with the IEEE 802.11 specification designed for home and office wireless network environments. The board uses a highly integrated four-chip configuration. It is designed to implement the front end of a 2.4 GHz FHSS subsystem and als includes the analog circuitry from th antennas to the analog baseband sig nals. The board is priced at $250.
Philips Semiconductors
INFO/CARD 140

SIGNAL PROCESSING COMPONENTS

Low profile splitter/combiner

Model BP2G is a low cost, low profil (0.077″ high) surface mount two-way 0 power splitter/combiner. The unit oper ates in the 1.42–1.66 GHz frequenc range with a typically low 0.6 dB inser tion loss (above 3 dB), a high 28 dB iso lation and a maximum voltage stand ing wave ratio (VSWR) of 2:1 on al ports. The unit can handle an absolut maximum input power of 1.5 W as divider and 0.75 W per port, divided b two, as a combiner. Additional feature include 50 Ω operation and tape an reel availability. The BP2G is priced a 99 cents each in quantities of 10–49.
Mini-Circuits
INFO/CARD 141

Dual-mode upconverter for CDMA/FM cellular

The RF2628 is a complete dual-mod code-division multiple acces (CDMA)/FM upconverter that can als be used as a direct bipolar phase shif keying (BPSK) modulator. Operatin from a single 2.7–5.0 V power supply the RF2628 features a +9 dBm outpu intercept point with 0 dB conversion gain. The unit contains a double-bal anced mixer stage and an output buffe amplifier. The RF2628 is priced a $1.75 in quantities of 10,000.
RF Micro Devices
INFO/CARD 142

Cellular and PCS VSWR alarm modules

Models 8460-N1 and 8480-N1 ar voltage standing wave ratio (VSWR monitors for use in cell site transmi antenna cables. These monitors contin uously measure root mean square (RMS) forward and reflected RF powe and trigger an internal single pole, dou ble throw (SPDT) form "C" latchin alarm relay when the antenna VSWI exceeds a user adjustable threshold

The alarm level can be set from 1.2:1 to 2.5:1. They also feature a one-second alarm activation time that eliminates false alarms caused by rapidly fluctuating RF transmit power. The modules come with built-in, dual-directional sampling couplers for monitoring the transmit line. Model 8460-N1 is designed for advanced mobile phone service (AMPS) band applications while the 8480-N1 covers personal communications service (PCS) band applications.
Narda Microwave-East
INFO/CARD 143

Surface-mount transfer switches

A miniature surface-mount transfer switch combines high performance in a compact configuration. The switch is available with a choice of four operating frequencies and two coil voltages. The SR-TMIN-MIN-X-X features an impedance of 50 Ω, a power rating of 10 W and a switching time of 15 ms maximum. Frequency options range from DC to 18 GHz, insertion los ranges from 0.2–0.7 and isolatio ranges from 40–70.
RLC Electronics
INFO/CARD 144

Fixed attenuator for radio applications

The 22081-06 WT is a 6 dB fixe attenuator that offers ±0.2 dB frequen cy sensitivity over the full WG 22 (W 28) band. Features include an attenua tion loss variation with temperature ±25 dB and a return loss of better tha –28 dB. 3, 6, 10, 20 and 30 dB attenua tion models are available.
Flann Microwave
INFO/CARD 145

4×4 hybrid matrix for power amplifier applications

A 4×4 hybrid matrix is available tha uses four amplifiers in parallel typicall to cover three channels, providing relia bility and performance for the user wit ess overall amplifier capacity. Features nclude a minimum isolation of 20 dB ınd a maximum insertion loss of 0.5 dB, vhich means increased efficiency and hannel isolation for the user. The unit ılso has a maximum voltage standing vave ratio (VSWR) of 1.25:1, a peak ımplitude balance of 0.50 dB and a peak ›hase balance of 3.0 degrees. It operates n the 750–960 MHz frequency range ınd measures 2.0″ × 2.5″.

\naren Microwave
NFO/CARD 146

/CO enhances ıp/downconverters

The V844ME01 is a low-cost, high- ›erformance voltage-controlled oscilla- or (VCO) designed for low noise ıp/downconverter applications. This /CO generates frequencies from ‚,200–3,400 MHz within a control volt- ıge range of 0.4–4.5 VDC. Features nclude spectral purity of –88 dBc/Hz ypically at a 10 kHz offset while oper- ıting off a 5 VDC supply and drawing 22 mA typically. Other features include a 1.1:1 linearity over frequency and temperature, output power of 5±3 dBm into a 50 Ω load while suppressing the second harmonic to better than –15 dBc, and pulls less than 25 MHz with a 14 dB return loss, any phase. The V844ME01 is priced at $15.95 each in prototype quantities.

Z-Communications
INFO/CARD 147

SUBSYSTEMS

Transceivers extend distance and increase data rate

The MAX3291/MAX3292 RS-485/RS-422 transceivers feature driver preemphasis circuitry, which increases the maximum distance and data rate for reliable communications by reducing intersymbol interference (ISI)—a form of data-dependent timing jitter caused by cable parasitics. Typically, preemphasis allows either double the data rate at

a given distance or double the distance at a given data rate as compared to other RS-485/RS-422 transceivers. The MAX3291 is optimized for data rates of 5–10 Mbps, while the MAX3292 is programmable for data rates as high as 10 Mbps using a single external resistor. Each device operates from a single +5 V supply and features a shutdown mode that reduces the supply current to 100 nA. Prices for the transceivers start at $2.70 in quantities of 1,000.
Maxim Integrated Products
INFO/CARD 148

Spread-spectrum radio modem

The Blue Streak is a high-speed digital communications spread-spectrum radio modem capable of linking various electronic devices. Features include unlicensed operation, frequency hopping spread spectrum (FHSS) technology, 50 user selectable sequences, adjustable data rates from 1,200–115,200 baud operations and multiple repeater capabilities.
Aerotron-Repco Systems
INFO/CARD 149

Aluminum feeds enhance performance of antennas

Aluminum feeds featuring wideband (7.1–8.5 GHz) capabilities are available on terrestrial microwave antennas. Formed from heavy duty aluminum, the feeds feature improved cross polarization (35 dB), a low 1.06 voltage standing wave ratio (VSWR), a return loss of –30.7 dB, gain improvements from 0.1 to 0.5 dBi and front-to-back ratio improvements from 2 to 3 dB.
Andrew
INFO/CARD 150

Planar antenna for portable products

The Splatch is a low-cost planar antenna that uses a proprietary grounded-line technique to extract outstanding performance from a tiny surface-mount element. The Splatch i immune to proximity effects, making i suitable for handheld applications suc as remote controls. The antenna mea sures 1.1″ × 0.5″ × 0.062″, exhibits a 5 Ω characteristic impedance, a voltag standing wave ratio (VSWR) of les than 1.9 and is available in standar or custom frequencies within th 300–900 MHz range. The Splatch i priced at less than $1 in productio quantities.
Linx Technologies
INFO/CARD 151

Low-profile GPS antenna module

The TMM869 global positioning sys tem (GPS) front-end antenna is pack aged in a compact low-profile radome c less than 13 mm that houses the entir front end of the antenna element, th low-noise amplifier (LNA) and the pres elect bandpass filter. The right han circular polarized antenna receives th GPS signal of 1,575.42 MHz with

passband width of 9 MHz. The antenna gain is +5.0 dBi typical with an elevation angle of 90°. Operating off 5 V, the module has a low current consumption of 15 mA with no signal. The power gain of the gallium arsenide (GaAs) field effect transistor (FET) LNA is 31 dB, and the noise figure is 0.9 dB typical. Several mounting methods are available. The TMM869 is priced starting around $18 in quantities of 1,000.
Toko America
INFO/CARD 152

DISCRETE COMPONENTS

100 W GaAs MESFET for L-band transmitters

The NES1823P-100 is a twin-transistor device consisting of two pairs of gallium arsenide (GaAs) metal semiconductor field effect transistor (MESFET) chips that can be externally combined in either a push-pull or balanced configuration. The unit is designed for 2.1–2.2 GHz base stations, but can be modified to be used in 2.3–2.4 GHz wireless local loop (WLL) and digital audio broadcast (DAB) applications. The NES1823P-100 delivers 100 W of output power with a linear gain of 11.0 dB typical and drain efficiency of 50% typical at V_{DS} = 10 V, I_{DSS} = 6 A and f = 2.2 GHz. The NES1823P-100 is priced at $306 in quantities of 100.
California Eastern Labs
INFO/CARD 153

Open type inductor for signal line filtering

The KQ 0603 is an 0603 size chip inductor that features a high Q factor and is used in signal line filtering. The inductor's flat top design allows for increased placement and its electrical and mechanical characteristics offer the ability to be used as a direct replacement for other open-type inductors. The inductor is available in both ±5% and ±10% tolerances, with a nominal inductance range of 1.8–120 nH.
KOA Speer Electronics
INFO/CARD 154

High power RF transistor

The PTF10112 is a high-power, high-frequency RF transistor intended for code-division, multiple access (CDMA) and time-division, multiple access (TDMA) applications in the personal communications service (PCS) band. This laterally diffused metal oxide semiconductor (LDMOS) device typically has a gain 3 DB higher than bipolar equivalents, with a minimum output power of 60 W at 1 dB compression. Designed to operate from a 28 V supply, this device has a typical power gain of 12 dB while exhibiting a gain flatness of ±0.2 dB over the 1.93–1.99 GHz PCS band. Class AB two-tone third-order intermodulation distortion is –40 dBc at 25 W. Efficiency is typically 41%.
Ericsson Components
INFO/CARD 155

AMPLIFIERS

Low distortion hybrid operational amplifiers

The KH232 and KH207 are two low distortion hybrid operational amplifiers designed specifically for wide dynamic range systems. Using current feedback topology and offering high slew rates and fast settling times, the KH207 is designed for high-gain applications while the KH232 is designed for low-gain applications. The KH207 features a gain range of +7 to +50 and –1 to –50 V/V. Other features include a small signal bandwidth of 170 MHz and a large signal bandwidth of 100 MHz at a gain of 20V/V. It operates on supply voltages of ±5 V to ±15 V and provides a maximum output current of 150 mA. The KH232 features a gain range of ±1 to ±5 V/V. Other features include a small signal bandwidth of 270 MHz and a large signal bandwidth of 95 MHz at a

gain of 20V/V. It operates on supply voltages of ±5 V to ±15 V and provides a maximum output current of 100 mA.
Kota Microcircuits
INFO/CARD 156

Class A linear amplifier

Model AR88258-50 is a class "A" linear amplifier that operates over the full 800–2,500 MHz frequency range with 50 W of output power at the 1 dB compression point. Gain is 47 dB minimum with typical two-tone intermodulation distortion at –30 dBc. The unit measures 8.75" × 19" × 22".
Comtech PST
INFO/CARD 157

Power amplifier for intermodulation testing

The DMS 7055 is a high power amplifier that delivers 50 W of RF output power over the 2,000–2,2000 MHz frequency range. Designed for intermodulation distortion measurements, the amplifier operates in class "A" from a 240 V AC power supply with integral output isolator for added protection. Features include a small signal gain of 48 dB minimum, saturated output power of 47 dBm typical and spurious outputs of –60 dBc maximum.
Densitron Microwave
INFO/CARD 158

SEMICONDUCTORS

Chip set for spread spectrum digital cordless phones

The Everest 900 MHz digital cordless chip set will enable cordless phone manufacturers to deliver products with world-class clarity and range at as much as half of today's typical costs. The chip set provides digital cordless phones with 200 MIPS of processing power. The set consists of two B900 digital signal processor (DSP) chips, two, four-channel codec CSP1009 communications signal processor chips an two W9009 RF transceiver chip designed for the 900 MHz industria scientific and medical (ISM) band. On of each chip is placed in the handse while the other is placed in the bas station. The chip set enables six hour of talk time and seven days of standb time, and is priced at $14.95 in quant ties of 100,000.
Lucent Technologies
INFO/CARD 159

SIGNAL SOURCES

Line of surface-mount VCXOs

A line of surface-mount voltage cor trolled crystal oscillators (VCXOs) ar available for the frequency range fror 1.5–60 MHz. Part numbering i VC2XXX for the non tri-state parts and VC3XXX for the tri-state parts Features include frequency stabilitie from ±15 ppm to ±100 ppm over a

operating temperature range of 0–70°C. Pullability is from 25–300 ppm. Input voltage can be either 5 or 3.3 V ± 5%. Pricing starts at $12.75.
Pletronics
INFO/CARD 160

Low phase noise crystal oscillators

The FE-103A series of low phase noise crystal oscillators is designed for wireless applications such as cellular phone base stations. Available in the 5–20 MHz frequency range, features include typical phase noise of –142 dBc at a 10 Hz offset, -148 dBc at a 100 Hz offset, –155 dBc at a 1 kHz offset and –158 dBc at a 10 kHz and 100 kHz offset.
FEI Communications
INFO/CARD 161

Low profile, high performance OCXO

Model XO5008 is a low profile, high performance oven compensated crystal oscillator (OCXO) that features a height of 0.75″ with a footprint of 1.5″ × 1.5″. The standard center frequency is 10 MHz with options available over the 3–50 MHz frequency range. The unit features an SC-cut resonator, sinewave output, five-minute warm up and moderate current consumption. AT-cut options are also available.
Piezo Technology
INFO/CARD 162

Low noise 13 GHz oscillator

The ESC 206-121 is a 13 GHz dielectric resonator oscillator that has a typical phase noise of –90 dBc/Hz at an offset of 10 kHz. Output power is 19 dBm minimum. Pulling into a 1.5:1 voltage standing wave ratio (VSWR) is less than 0.01% and harmonics are –20 dBc. The oscillator is housed in a low profile flanged package measuring 1.6″ × 1.7″ × 0.6″ including the flange.
Electronics Surveillance Components
INFO/CARD 163

RF software

Development software updated with Web apps

National Instruments' Labview 5.1 offers users an improved application development by streamlining the creation of Web-enabled applications. The new modular application architecture allows users to create smaller executables that improves system performance and decreases memory usage. The software also extends the use of Active X (COM) to allow for seamless integration of the math and interactive analysis capabilities of the Mathworks Matlab and National Instruments HiQ.
National Instruments
INFO/CARD 115

Design software can be client or server

O-Matrix 4, from Harmonic Software, offers a client/server capability that enables O-Matrix to communicate with other Windows programs and perform interprocessing computing. The interprocessing capabilities enable the software to act as either a computing client or server with other applications. Both data and programs can be transferred between instances of O-Matrix or other applications. The process allows the users to develop a Mathlink package for communicating with Mathematica.
Harmonic Software
INFO/CARD 116

Software designed for EM, layout, linear analysis

Hewlett-Packard (HP) EEsof introduces its Momentum circuit designer. The software enables engineers to perform electromagnetic (EM) analysis as well as layout and linear analysis in a single, integrated environment. The software includes strip-slot computation, adaptive frequency sampling, edge mesh and other capabilities. It is designed for use with PCs using Windows 95 and Windows NT 4.0.
Hewlett-Packard EEsof
INFO/CARD 117

PC-based asynchronous data and protocol analyzers

Frontline Test Equipment has released its Serialtes Async and Serialtes Spy asynchronous data and protocol analyzers for Windows NT environments. Previous releases support Windows 95/98. The software is designed to allow a PC to function as a full-featured serial data and protocc analyzer, enabling the user to troubleshoot asynchronous data transmissions on the fly, reviewing and searching the data as it is being captured.
Frontline Test Equipment
INFO/CARD 118

Circuit design tool updated with optimization options

Tanner EDA's T-Spice Pro is an integrated circuit design capture, simulation and data viewing system fo Windows NT/95/98. The software ha been updated to include new optimization options, remote simulation over network, parameter sweeping, mor flexible behavioral modeling an advanced post processing support.
Tanner EDA
INFO/CARD 119

CDMA developed for base station test platform

Wavetek has developed version 2. code-division multiple access (CDMA software for the Model 4032 base station tester. Technicians using the software with the Model 4032 can determine the PN offset of a sector within 3C seconds.
Wavetek
INFO/CARD 120

Software on the Web

Design software demo offered on Web

Optotek offers a demonstration version of its MMICAD suite of CAE/CAT software. The suite allows for precise circuit simulation and device modelling, contributing to design realization and high-yield, affordable manufacturing of RF and microwave circuits.
Optotek
INFO/CARD 121

To access this Web site, and other Web sites offering downloadable software, check out **RF Design Online** for direct links.

Booklet describes CDMA history, growth, technology

Wavetek offers a booklet that offers readers an overview of code-division multiple access. *CDMA Introduction* describes the history of CDMA, its growth and the differences between CDMA, time-division multiple access (TDMA) and frequency-division multiple access (FDMA). The booklet also describes advantages of CDMA, codes in CDMA, logical channels in CDMA and basic CDMA specifications

Wavetek
INFO/CARD 122

Short form catalog features testers and power supplies

IFR Americas' 1998/1999 short form catalog features a quick reference to Kikusui power products and detailed product descriptions. Products include Kikusui's full line of safety testers, DC power supplies, AC power supplies, electronic loads and battery testers.

IFR Americas
INFO/CARD 123

Catalog and application notes offered on CD-ROM

Gage Applied Sciences has issued its first full line catalog and application notes CD-ROM. The CD-ROM contains descriptions, specifications and technical information on Gage products. Up to 100 application notes, articles, and white papers are offered as well as a question and answer section. All files are saved in the PDF format and are readable on Windows 95, Windows NT, Macintosh and Unix computers.

Gage Applied Sciences
INFO/CARD 124

Catalog features foil and film resistors

Riedon's new line of foil and film resistors are described in a 12-page catalog. The catalog features as many as 35 power and precision resistors and networks using advanced metal foil, metal film and thick film elements. Included are precision low ohm foil resistors for current sensing and shunt applications.

Riedon
INFO/CARD 125

Design guide offers A/D converters

Maxim offers an analog design guide that features the company's 18-bit sigma-delta analog digital converters (ADC). The MAX1400 family of ADCs can sample to 4.8 ksps and can maintain 16-bit performance at 480 samples per second.. The devices also offer a 4.8 ksps max sample rate.

Maxim
INFO/CARD 126

Guide offers insights into DECT, PWT

Personal Wireless Communications with DECT and PWT, from Artech House, is a guide to two personal wireless communications systems: digital enhanced cordless telecommunications (DECT) and personal wireless telecommunication (PWT). The book offers background material and technical principles, basic protocols and implementations, plus features and a wide range of applications. For more information visit Artech House's Web site.

Artech House
INFO/CARD 127

Newsletter dedicated to digital communications

Analog Devices "Communications Direct" is dedicated to digital communications issues. The current 8-page edition, Vol 3, No. 3, presents an overview of ADSL and cable modem technologies and the viability of each in the marketplace. To request a copy visit Analog Devices Web site.

Analog Devices
INFO/CARD 128

Catalog highlights circuit protection products

AVX offers its TVS catalog, featuring the company's Transguard, Staticguard and Multiguard multilayer ceramic transient voltage suppressors. The catalog provides users with detailed product specifications in addition to information on applications. Schematic diagrams for typical circuits and application notes are also provided.

AVX
INFO/CARD 129

Product catalog features filters, surge suppressors

Control Concepts' product catalog features the company's line of power conditioning equipment. Included are a variety of high-performance filters and surge suppressors designed to help eliminate power fluctuations at the source. The 20-page catalog offers technical specifications, typical installation, test results and other information.

Control Concepts
INFO/CARD 130

Catalog offers SMT, thru-hole PCB components

Keystone's 16-page catalog supplement offers information on the company's surface mount (SMT) and thru-hole printed circuit board (PCB) components and hardware. Product groups include coin cell holders, retainers and clips, vertical 20m coin cell holders, surface mount battery clips, SMT and thru-hole test points, and PC screw terminals. The catalog also addresses surface mount tape and reel packaging availabilities.

EMC Test Systems
INFO/CARD 131

Online

HP offers end-of-production equipment on its Web site—Hewlett-Packard (HP) has listed end-of-production equipment and more than 1,200 refurbished test and measurement products on its Web site. The refurbished equipment comes with a one-year limited warranty covering all components covered by HP's warranties on new products. To access the site go to **www.hp.com/go/refurbished**.

Hewlett-Packard
INFO/CARD 132

RF Design Online

For direct access to those companies offering information through their Web sites, go to **www.rfdesign.com**.

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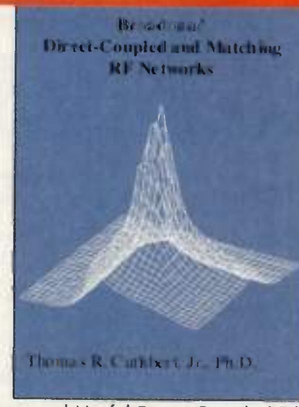

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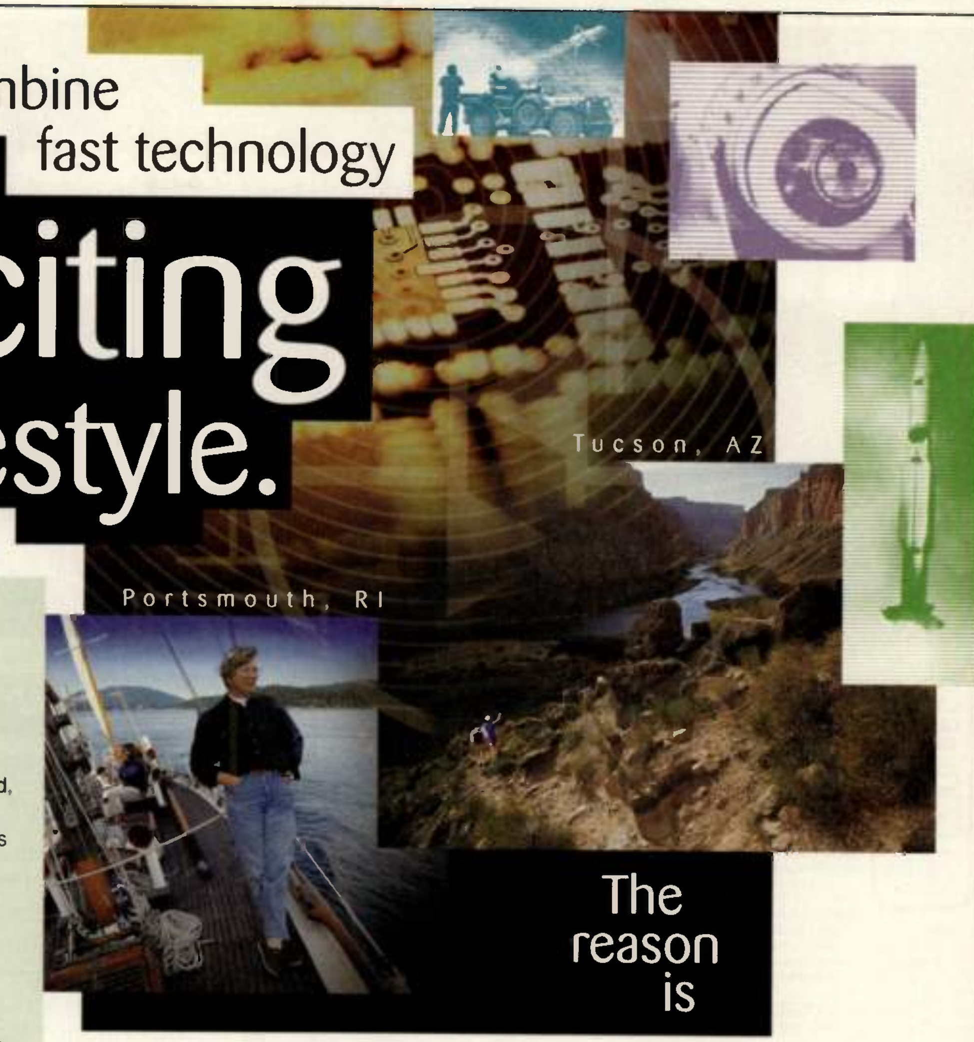

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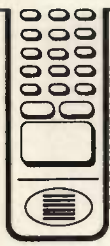

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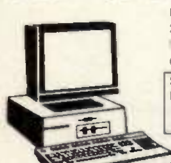

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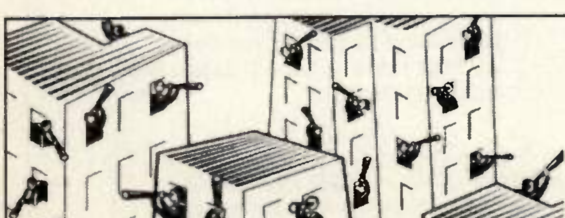

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Dr. Bami Bastani

Chief Executive Officer, Anadigics

Dr. Bastani has previously worked with a number of semiconductor companies including Intel, National and Fujitsu. He joined Intel in 1980 as the senior device engineer where he was involved in memory and microprocessor technology development.

In 1985 he joined National Semiconductor where he served in several executive positions. In 1996, Dr. Bastani joined Fujitsu Microelectronics as executive vice president for System LSI Group, where he lead semiconductor business units in networking and processing information for worldwide markets.

Dr. Bastani holds a B.S.E.E. from the University of Arkansas College of Engineering. In 1977, he received his M.S.E.E. and earned his Ph.D. in 1980 from Ohio State University.

Dr. Bastani has written over a dozen publications on semiconductor devices, and holds three US patents in semiconductor technology.

Anadigics manufactures high-performance gallium arsenide integrated circuits (GaAs ICs) used in a variety of high-volume communications applications.

The interview was conducted by Senior Associate Editor Roger Lesser.

RF Design: *Where is third generation (3G) headed?*

Bastani: I was at the worldwide conference. It was interesting how 3G is a "beauty in the eyes of the beholder." While one company sees it as an evolutionary path others see it as a common standard. Yet, others are saying we already have dual-mode, triple-mode, dual-band and we can accommodate diversity. There is room for diversity. We don't have to have everything under the same standard. So, I think the only thing that will be common is the name. However, I would like to see us converge on a standard.

RF Design: *How does a company, like Anadigics, prepare?*

Bastani: First of all, 3G gives us opportunity. But you also have to have good partnership with industry leaders. We have worked with companies like Ericsson and Qualcomm in the past. We will be using them as sounding boards and then positioning Anadigics. We also will be participating in the standards bodies and conferences to make sure we are well prepared.

RF Design: *What does this hold for the future of Anadigics chip technology?*

Bastani: I think it will just be an evolution of what we do. It is going to be a digital and linear standard. Coupled with our gallium arsenide (GaAs) capability, which lends itself to linearity and efficiency and power management. These are all the elements required for the RF deck of the third generation handsets.

RF Design: *While GaAs technologies offer benefits, isn't there room for silicon-based technologies within 3G?*

Bastani: When you're dealing with the high end, the linearity and efficiency you get out of gallium arsenide will still be leading the pack. Especially in the critical transmitter side of the radio module.

RF Design: *When do you anticipate producing 3G chip technology?*

Bastani: At least a couple of years. We are keeping abreast of developments. But I think what is going to be developed is industry leader dependent. And we are fortunate to have the strong ties with these players.

RF Design: *How can you do this?*

Bastani: There are two things we are doing. One is to ensure we understand what the customer wants. This is where we need to be part of the roadmap, part of the interface with the company we are supporting. The second part is to anticipate the future. There is nothing unique about the gallium arsenide industry compared to the rest of the semiconductor industry. But, you need to move up the integration and performance path. So, what you will see will be modules that can integrate multiple functions into the substrate. Also, higher performance FETs (field-effect transistors) where you can move up the performance bridge. Another area is shrinking the gate so you can get higher performance and higher gain. As development tools progress, we will progress our performance.

RF Design: *What higher performance areas can we anticipate?*

Bastani: The biggest will be in linearity and efficiency. This is especially true for power amplifiers. Also, another area will be how fast a raw transistor runs. In this area, we see a tremendous amount of interest from our broadband customers. For example, when we deal with 10 Gbits per second transimpedance amplifiers. we are introducing new pseudomorphic high-electronmobility transistor (pHEMT) technology. This will drive our 10 GBits per second product developments and offerings. In the broadband area, 2.5 Gbits per second is the sweet spot of our current MESFET (metal semiconductor field-effect transistor) technology. So, broadband is pushing transistor performance.

RF Design: *What is wireless pushing?*

Bastani: Wireless is heavily pushing linearity, power efficiency and cost.

RF Design: *How do you view Anadigics place in the market?*

Bastani: When you look at wireless from Anadigics perspective, power amplifiers have been our dominant place. I view Anadigics as an analog RF company in the communications market. We have segmented the market into wireless, and broadband communication. Cable addresses set top boxes, cable modems, and digital TV broadcast, while fiber represent our participation in telecomm and datacomm. 40% of our revenue is wireless, and 60% is broadband communication with 40% in cable and 20% in fiber market segments. RF

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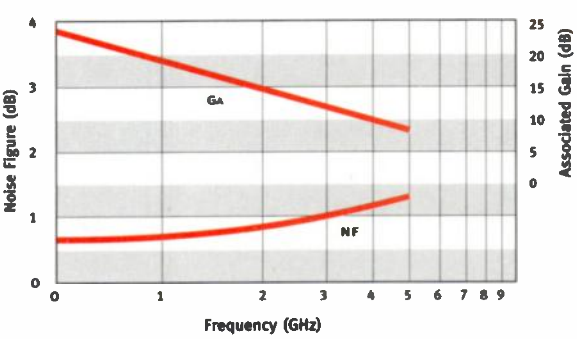

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