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engineering principles and practices

June 1994

Cover Story
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Featured Technology
Designing for 1.5 - 2.5 GHz

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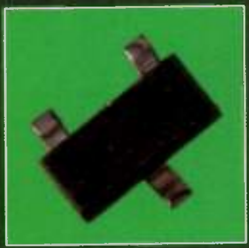

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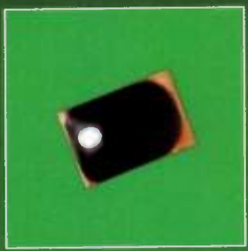

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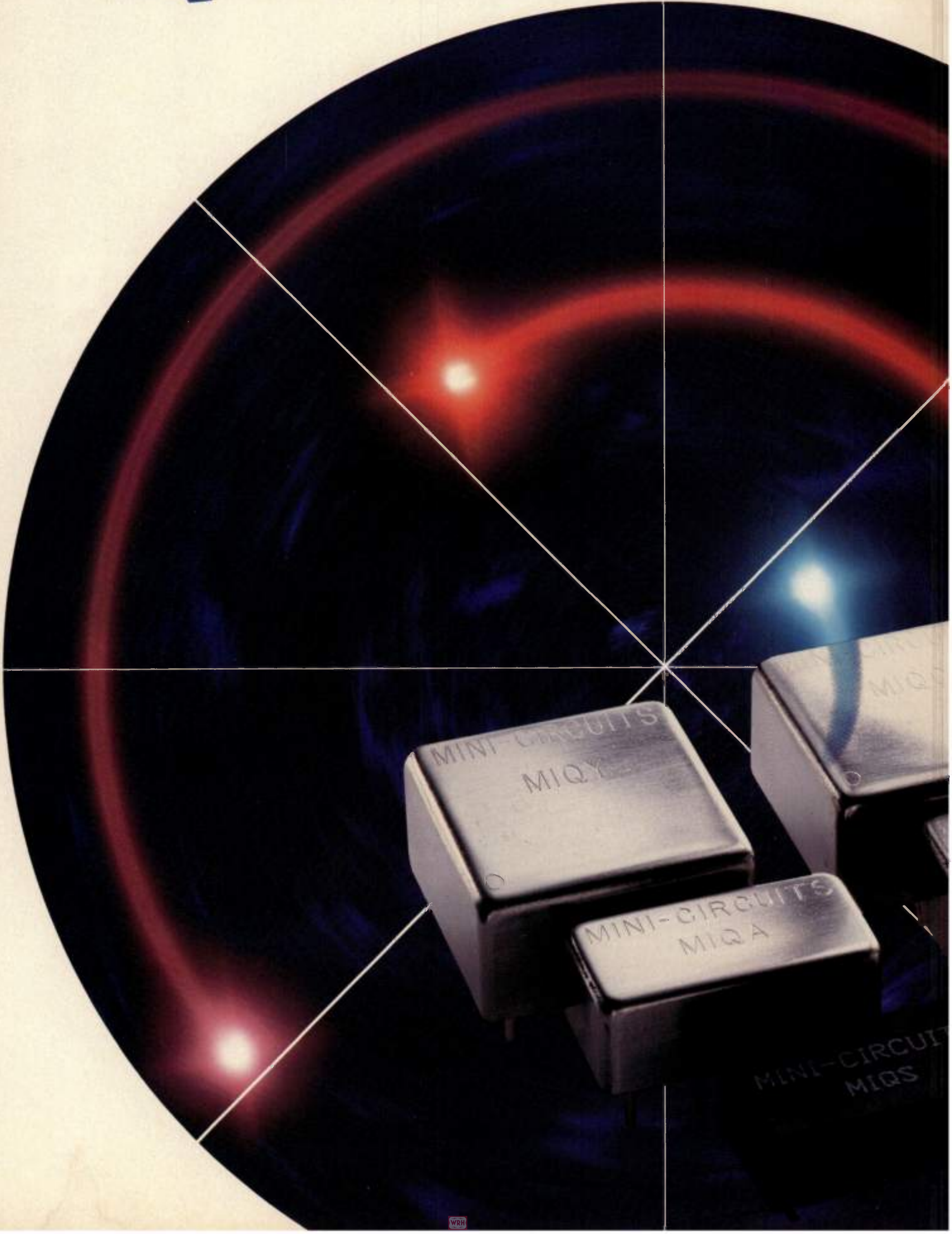
INFO/CARD 2

I&Q MODULATORS



RF design

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June 1994

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RF editorial

An Update on Activities at *RF Design*

By Gary A. Breed
Editor

Every once in a while, we like to take some time to inform you of activities in our own offices, not just in the RF industry at large. Recently, there have been sevral things happening around here that warrant a status report.

First on the report are some plans for a party! When our fifteenth birthday arrived last fall, we noted it without fanfare. But that's no way to mark a milestone, so we're going to celebrate just before we turn sweet sixteen! The October issue will include a review of *RF Design*'s growth and development since the publication of the first issue in November/December 1978.

To remind everyone of the event, we have added a new emblem to our cover that proclaims our "15 Years of Wireless Leadership". We're proud of our history of covering of radio technology, which started long before "wireless" became today's hot buzzword!

Next is a change to our Product Report column. This column reports on marketplace and technology developments, looking at a different product type each month. Through our editors, companies making those products have offered their perspectives.

Now, those perspectives will come to you directly, without editorial interpretation, in short reports authored by those companies. They are the specialists, and you will be able to compare their views side-by-side. Retitled as the Product Forum, the first new column appears next month in the July issue. (Our independent editorial analysis will continue in the Industry Insight column).

We have had some changes worth noting in our staff assignments. I have been promoted to Associate Publisher, and although my responsibilities as Editor are pretty much unchanged, my role has grown in the areas of policy-making and fiscal responsibility.

Although his title hasn't changed, our Technical Editor, Andy Kellett, has been promoted in his level of responsibility, as well. Andy handles all new product information, and will be managing the process of selecting articles, including all communications with authors.

Our advertising sales staff, after a rearrangement of territories several months ago, has been working especially hard to bring new companies into our family of advertisers. These companies have products you need to know about, many addressing new RF applications.

Watch for information on next fall's RF Expo East — there are some new additions being planned, and the technical program is coming together quickly. If you are interested in presenting a technical paper or moderating a session, let me know; there are just a few spots left that we'd like to fill quickly. Also at RF Expo East, we will have the formal presentation of the Grand Prize in this year's RF Design Awards Contest!

Things are always busy in the RF business, whether you are designing products, answering customer questions, or managing an engineering group. It's no different in the technical publishing business either! We stay busy trying to get you the information to help get your job done right!

RF design

Established 1978

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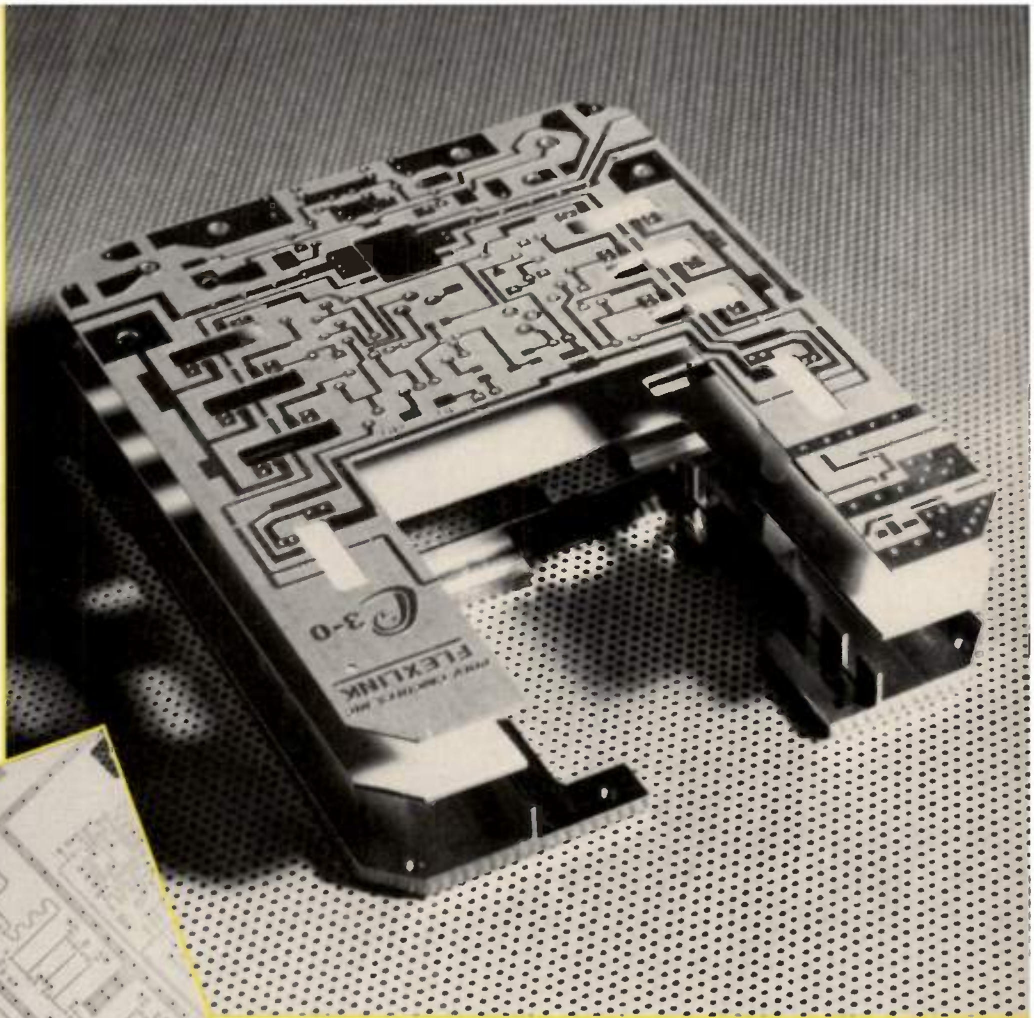

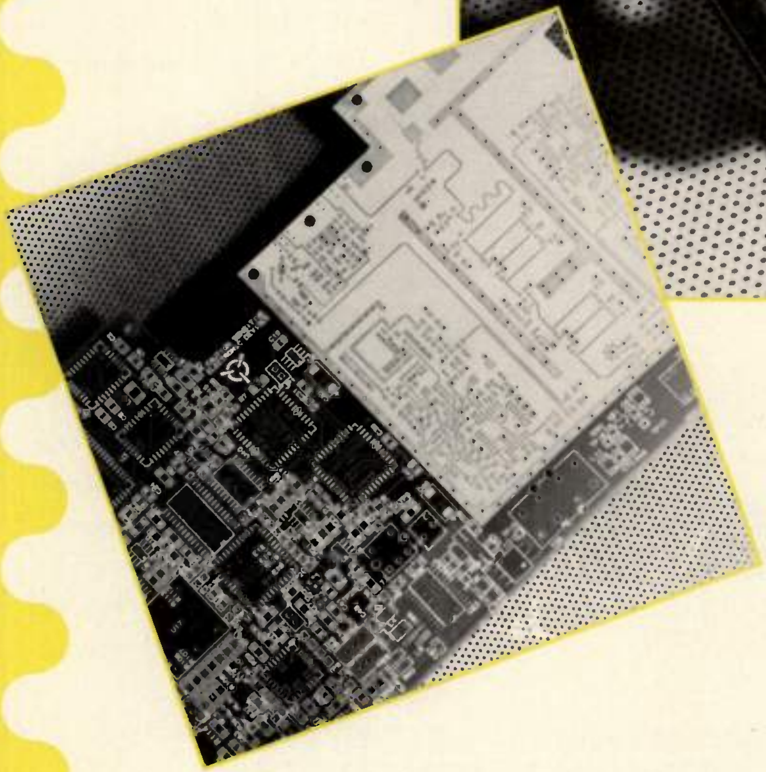

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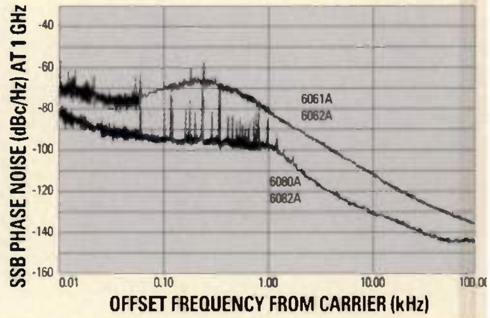
INFO/CARD 9

GIGA-TRONICS SYNTHESIZED RF SIGNAL GENERATORS

Giga-tronics May Not Be The First Name You Think Of For RF Synthesizers, But It Should Be.



Yes, Giga-tronics is a new name in RF synthesizers.

But this new choice may be your best choice.

Because an RF synthesizer from Giga-tronics is an incredible value. You get proven performance and reliability at a low price, backed by an experienced service organization.

Here are the specifics.

Performance.
Check the charts. Giga-tronics offers four instruments, each with its own unique combination of performance and capability.

The 6061A is ideal for accurate and affordable general purpose RF testing.

The 6062A adds the modulation capability you need to generate complex signals for communications and radar testing.

And the 6080A and 6082A give you the performance and capability to test sophisticated systems accurately and quickly.

It's easy to command the power and capability of each instrument from the front panel. And the low profile, standard rack size and IEEE-488 interface make each instrument ideal for ATE applications.

Reliability.
The John Fluke Manufacturing Company initially developed and introduced this line of synthesized RF signal generators. To date, thousands have performed flawlessly in the field.

The instruments incorporate self-testing,

Giga-tronics offers you a full line of RF synthesizers with proven performance and reliability at low prices.

Specifications	Giga-tronics 6061A	Giga-tronics 6062A	Giga-tronics 6080A	Giga-tronics 6082A
Frequency Range Switching speed	.01 to 1050 MHz <100 ms	.1 to 2100 MHz <100 ms	.01 to 1056 MHz <100 ms	.1 to 2112 MHz <100 ms
Spectral Purity* Spurious Subharmonics	<-60 dBc None	<-54 dBc <-45 dBc	<-100 dBc None	<-94 dBc <-45 dBc
Phase Noise* @ 20 kHz offset	<-117 dBc/Hz	<-110 dBc/Hz	<-131 dBc/Hz	<-125 dBc/Hz
Residual FM* (Bandwidth)	<12 Hz (.5 to 3 kHz)	<24 Hz (.5 to 3 kHz)	<1.5 Hz (.3 to 3 kHz)	<3 Hz (.3 to 3 kHz)
Output Range* Accuracy Reverse Power Protection	+13 to -147 dBm ±1 dB >127 dBm 50 Watts/50 Vdc	+13 to -147 dBm ±1.5 dB >-127 dBm 25 Watts/25 Vdc	+17 to -140 dBm ±1 dB >127 dBm 50 Watts/50 Vdc	+13 to -140 dBm ±1 dB >-127 dBm 25 Watts/25 Vdc
Amplitude Modulation Depth Distortion @ 30%	0–99.9% <3%	0–99.9% <3%	0–99.9% <1.5%	0–99.9% <1.5%
Frequency Modulation Max. Deviation* Distortion	100 kHz <1%	400 kHz <1%	4 MHz <1% @ 50% Dev.	8 MHz <1% @ 50% Dev.
Phase Modulation Max. Deviation*	NA	40 Rad.	40/400 Rad.	80/800 Rad.
Pulse Modulation On/off Rise/fall time Minimum Pulse Width	NA	>80 dB <15 ns <2 µs	>40/60 dB <15 ns (Typ 7.5 ns) <30 ns	>80 dB <15 ns (Typ 7.5 ns) <30 ns
Internal Modulation Source Level Range Waveforms Programmable	400, 1000 Hz NA Sine Yes	400, 1000 Hz NA Sine Yes	0.1 Hz to 200 kHz 0 to 4 Vpk Sine/Sq/Tri/Pulse Yes	0.1 Hz to 200 kHz 0 to 4 Vpk Sine/Sq/Tri/Pulse Yes
Memory Locations (NVM)	50 Full Function	50 Full Function	50 Full Function	50 Full Function

*Specifications for the 6061A and 6080A are at 1 GHz, and specifications for the 6062A and 6082A are at 2 GHz. Phase noise is typical for the 6061A and 6062A.

internal diagnostics and modular design for easy fault isolation.

Service.

If a problem occurs, our technical support staff can often help you find and fix it over the phone.

If you need to return an instrument for service, we can take care of it at our factory in California, or at one of our worldwide service centers.

Experience.

Granted, Giga-tronics may be a new name for RF synthesizers, but it's a trusted name to microwave test professionals.

Giga-tronics has a 14-year history of building test and measurement gear for the most demanding requirements. We've shipped thousands of instruments for use in testing communications, radar and EW systems.

Value.

What makes a Giga-tronics RF Signal Generator such an incredible value is its blend of proven performance and reliability at a low price.

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510 328 4650
Telefax: 510 328 4700

INFO/CARD 10

Letters should be addressed to: Editor, RF Design, *6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Letters may be edited for length or clarity.*

Receiver Notes

Editor:

Nice job on your tutorial "Receiver Basics — Part 1: Performance Parameters" [February *RF Design*]. I'd like to add some comments.

Basically, a receiver must produce the desired output for signals you want to receive, and no output for signals you don't want to receive. Furthermore, the reception of the desired signals, which are often weak ones, should not be degraded because some strong undesired signals are present at the receiver input. To the user, "hearing" an unwanted signal (or the noise caused by it) is "interference;" desensitization due to ti often isn't noticed, or if it is, fault is often assigned elsewhere.

Phase noise in a local oscillator, whether crystal controlled or synthesized, is crucial to off-channel rejection. The noise acts like "local oscillator injection" into the mixer for adjacent channels, High IF rejection does no good if there is a lot of phase noise in the oscillator or synthesizer. In this case, adjacent channel rejection and oscillator phase noise are directly related.

Note also that phase noise in a transmitter is a frequent cause of interference on other channels, and receivers are often wrongly blamed. Antenna locations and properties are other factors in overall receiver performance.

RF Design is my top magazine reading priority. Keep up the good work.

Jack Streater
Mentor Radio Company

Spread Spectrum Clarification

Readers:

I made an incorrect statement in my tutorial "A First Introduction to Direct Sequence Spread Spectrum" [April *RF Design*]. Direct sequence SS *is* inherently a below-the-noise system because of its coherent nature, and does not require a positive signal to noise ratio as I stated. I apologize for the error.

I would like to thank Bob Dixon of Omnipoint Corporation and Chuck Phillips of Dunkirk, MD for gently pointing out my fundamental error. Bob reminded me of a well-known direct sequence SS system, the Global Positioning System, that operates with a signal to noise ratio that is typically –29 dB.

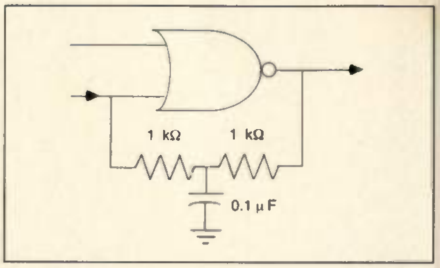
Part of my confusion arose from the fact that some current SS implementations are *not* below-the-noise systems. Some of them trade processing gain for data rate capability. Other systems meet FCC criteria for "spread spectrum" in the unlicensed ISM bands, but they are not true SS techniques. These types of systems *do* "require a total energy that is above the level of noise and other signals," as stated in the article.

Gary Breed
Editor

A Note on ECL for RF

Editor:

In the article on analog use of ECL by R.N. Mutagi [*RF Design*, April 1994], the author discusses several circuits for establishing a voltage level for linear operation (ECL threshold bias). A method not discussed in the article, used by RCA in the TRADEX and ALCOR radar systems, is shown in the figure below. It has proven to be simple, effective and reliable.



The basic idea is to connect the inverting digital gate for 100% DC feedback. This is obtained with the two resistors connected from the inverting output to the input. This forces the circuit to remain at its linear region. The resistors shown are 1kΩ, but can be adjusted to perform other functions such as transmission line termination. Due to the inverse feedback, the voltage level will automatically track the required threshold voltage despite thermal, aging and power supply effects. The bypass capacitor in the center ensures that there will be no AC feedback, thus allowing the gate to swing through its linear region to produce gain for any AC-coupled signal. The capacitor also prevents oscillation at the frequency where the delay is equivalent to 180 degrees (although normally the feedback resistors are too high for this to occur). This circuit has become standard for developing a digital clock from a sine wave reference.

This method should not be used with TTL logic, because the loss of input signal would likely result in the output transistors drawing excessive current.

David Freedman
Exetron

Coupler Addition

Editor:

Your article "The RF Coupler: A 20th Century Workhorse" [Product Report, February *RF Design*] was right on the money. As the article points out, the industry now demands smaller, cheaper, more consistent, surface mount compatible couplers, particularly for commercial applications. EMC has introduced a line line of subminiature couplers based on a proprietary thick film process, covering cellular, GPS and PCN frequency bands. These thick film Lange couplers have integrated crossovers which eliminate the usual bond wires. They handle 250 watts and are compatible with automated assembly processes.

David J. Markman
EMC Technology, Inc.

RF calendar

June

14-16 **Nepcon East '94**
Boston, MA
Information: Michele Filippi, NEPCON East '94, Reed Exhibition Companies, 999 Summer Street, P.O. Box 3833, Stamford, CT 06905–0833. Tel: (203) 352–8254. Fax: (203) 964–0176.

19-24 **1994 IEEE AP-S International Symposium and URSI Radio Science Meeting**
Seattle, WA
Information: Jan Kvamme, Conference Manager, UW Engineering Professional Programs, 3201 Fremont Avenue North, Seattle, WA 98103. Tel: (206) 543-5539. Fax: (206) 543-2352.

June/July

27-1 **Conference on Precision Electromagnetic Measurements**
Boulder, CO
Information: D.B. Sullivan, Time and Frequency Division, NIST, 325 Broadway, Boulder, CO 80303. Tel: (303) 497-3772 or (303) 497-3295.

July

4-7 **HF Radio Systems and Techniques**
York, UK
Information: HF '94 Secretariat, IEE Conference Services, Savoy Place, London WC2R 0BL, UK. Tel: 44-071-344 5478/5477. Fax: 44-071-497 3633.

6-8 **2nd International Conference on ADDA**
Cambridge, UK
Information: ADDA '94 Secretariat, IEE Conference Services, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071 344 5478/5477. Fax: 071 497 3633.

August

9-12 **IREECON 1994 - International Electronics and Telecommunications Exhibition**
Melbourne, Australia
Information: TWI, International Exhibition Logistics, 3190 Clearview Way, San Mateo, CA 94402. Tel: (415) 573-6900. Fax: (415) 573-1727.

25-28 **IEEE Electromagnetic Compatibility Symposium**
Chicago, IL
Information: Thomas Braxton, Vice-Chair, AT&T Bell Laboratories, Room 2B-217, 2000 N. Naperville, Road, Naperville, IL 60566. Tel: (708) 979–1299. Fax: (708) 979–5755.

September

27-29 **16th Piezoelectric Devices Conference**
Kansas City, MO
Information: Electronic Industries Association, 2001 Pennsylvania Avenue, N.W., Washington, DC 20006. Tel: (202) 457–4930. Fax: (202) 457–4985.

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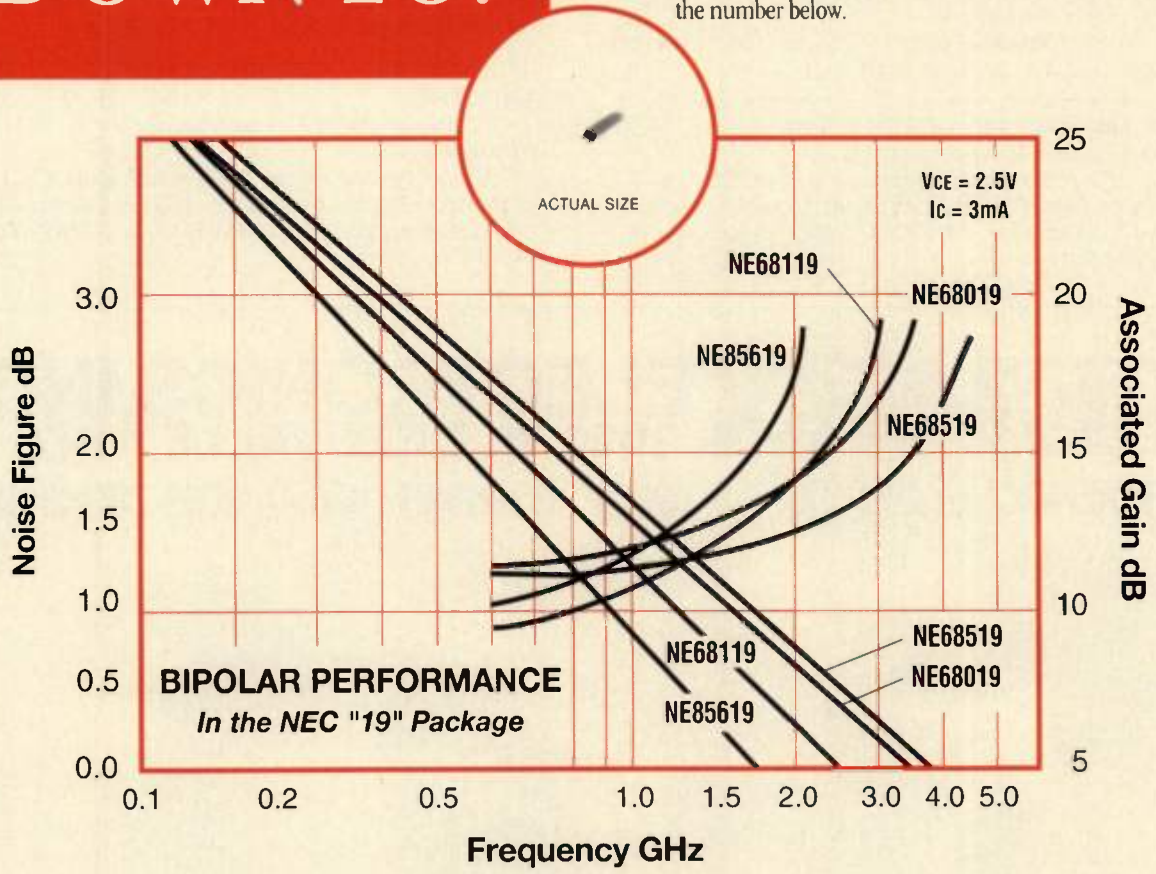
They're available now, on tape and reel, and priced to meet your needs. For Data Sheets and a copy of *Automated Assembly of Miniature Parts*, call your nearest CEL Sales Office or circle the number below.



NEC Bipolars are available in these packages (shown approx. 2X size)

NEC "19" Pkg Ultra Super Mini Mold	NEC "18" Pkg 4 Pin Super Mini Mold	NEC "30" Pkg SOT-323 Style	NEC "33" Pkg SOT-23 Style	NEC "39" Pkg SOT-143 Style

NEC

California Eastern Laboratories

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INFO/CARD 15

Introduction to Electromagnetic Compatibility Design Practices
July 7-8, 1994, Milwaukee, WI
Information: University of Wisconsin - Milwaukee, Non-Credit Registration Office, Drawer No. 491, Milwaukee, WI 53293. Tel: (414) 227–3200 or (800) 638–1828. Fax: (414) 227–3146.

NIST Time and Frequency Seminar - Level-I
June 20-21, 1994, Boulder, CO
NIST Time and Frequency Seminar - Level-II
June 22-24, 1994, Boulder, CO
Information: Patsy Tomingas or Wendy Ortega, Division 847, NIST, 325 Broadway, Boulder, CO 80303. Tel: (303) 497–3276 or (303) 497–3693.

High-Frequency Modeling and Simulation Seminar
June 15, 1994, Los Angeles, CA
June 17, 1994, Phoenix, AZ
Information: HP-EEsof, 5601 Lindero Canyon Road, Westlake Village, CA 91362. Tel: (800) 343–3763.

Design for Testability and for Built-in Test
July 11-14, 1994, Los Angeles, CA
Information: UCLA Extension, Engineering Short Courses, 10995 LeConte Ave., Ste. 542 Design for Testability and for Built-in Test, Los Angeles, CA 90024. Tel: (310) 825-1047. Fax: (310) 206-2815.

Microwave Antenna Measurements: Far-Field, Near-Field, Compact Ranges and Anechoic Chambers
June 13-17, 1994, Northridge, CA
Information: Shirley Lang, Center for Research & Services, School of Engineering & Computer Science, California State University, Northridge, CA 91330–8295. Tel: (818) 885–2146. Fax: (818) 885–2140.

Avionics & Weapons Systems Flight Test
August 22-26, 1994, San Diego, CA
Navstar/GPS: Design Applications
August 1-3, 1994, Washington, DC
High Speed & Microwave Devices & Applications
October 24-27, 1994, Boston, MA
Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995–6335. Fax: (818) 995–2932.

Applied RF 1
August 22-26, 1994, Los Altos, CA
Wireless Systems
August 29-September 2, 1994, Los Altos, CA
Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel: (415) 949–3300. Fax: (415) 949–4400.

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INFO/CARD 17

Space-based Imaging Radar Keeps Track of Environmental Trouble Spots

Ball Corporation's Spaceborne Imaging Radar-C/X-band Synthetic Aperture Radar was launched April 8, 1994, aboard the space shuttle Endeavour. The mission, first of two, was to study the damaging effects of pollution and agricultural practices on vegetation to learn how these changes may affect global weather patterns. Scientists are studying photos of 400 environmentally troubled spots including the Galapagos Islands, the Amazon River Basin, Hawaii, Mount Pinatubo, and Death Valley. The photos will be used as a baseline of Earth's health. The radar, an international project with Italy, Germany and the United States, was designed for NASA's Jet Propulsion Laborataory. It penetrates clouds, dense vegetation, ice and dry sand using frequencies in L-, C- and X-Bands.

This image is of Isla Isabella, Galapagos Islands, through the smoke of a huge fire threatening thousands of turtles, according to David Agular, Ball Corporation. On Endeavour's 40th orbit the 47 by 37 mile area was scanned by the L-band radar in HH polarization from the Space borne Imaging Radar C/X-Band Synthetic Aperture Radar.

IEEE Investigates Ambient Noise for PCS Standards – This research on various PCS modulations will identify the maximum on-site field disturbance levels which would allow proper PCS operation. The information will be used to base a standard measurement practice for measuring frequency, amplitude and bandwidth of radio noise environment. The project is called P1385 and titled "IEEE Recommended Practice for Measurement of Radio Disturbances in the Frequency Range of 800 MHz to 40 GHz Capable of Interfering with PCS (Personal Communications Service) Systems". The working group needs volunteers with expertise in PCS and EMC as it relates to RF ambient and analysis. The work will include documentation, analysis, and instrumentation. The meeting will be at Global Product Compliance Laboratory, AT&T Bell Laboratories, Holmdel, NJ 07733-3030 during the summer of 1994. Project Chairman is Dheena Moomgllan, Room No. 11B 184. Tel: (908) 834-1806. Fax: (908) 834-1030.

IEDM Calls for Papers – The 1994 IEEE International Electron Devices Meeting in San Francisco, CA, December 11-14th, at the San Francisco Hilton & Towers, has a call for papers in the following areas: Integrated Circuits, Quantum Electronics and Compound Semiconductor Devices, Device Technology, CMOS Devices and Reliability, Solid State Devices, Modeling and Simulation, Vacuum Electronics, and Detectors, Sensors & Displays. Abstract is due by July 1, 1994. For complete call for papers contact: Melissa Widerkehr, IEDM, 1545 18th Street NW, Suite 610, Washington, DC 20036 USA. Tel: (202) 986-1137.

U.S. 1994 R & D Funding – Battelle forecasts $164.5 billion for a 2.36 percent increase over the estimated $160.8 billion actually spent by the National Science Foundation last year for R&D. The two primary sources are the federal government ($69.8 billion) and private industry ($84.9 billion) for a total of $154.7 billion of which the primary performer of R &D will be private industry ($114.8 billion). Battelle expects continued inflation to absorb the projected increase in total R&D. However, industry shows a diminished interest in actually performing basic research and now that academia, because of financial reasons, is willing to be partners with industry, most R&D will be done by them. This represents a cost saving to industry. Battelle expects no real increases in actual funding will be seen until 1995 and the outlook for federal support is not optimistic. They state that those industries most successful in using R&D will survive and grow.

Georgia Tech Research News – Researchers have produced low voltage magnetic microactuators that operate where more common electrostatic devices can't be used. The iron core wrapped around integrated electrical windings can replace transistors for switching HF signals making voltage microconverters possible. Their fabrication technologies allow these devices to be made fully integrated on a silicon chip surface which allows simultaneous production of multiple devices on the same substrate. The researchers are also working on a magnetic microactuator that would be switched by a pulse of current. The researchers altered standard microelectronics processing techniques to produce thick layers of copper conductor for the winding and a thick plating of nickel-iron for the magnetic core. Magnetic microactuators can generate more force than their electrostatic cousins, operate on lower voltages and

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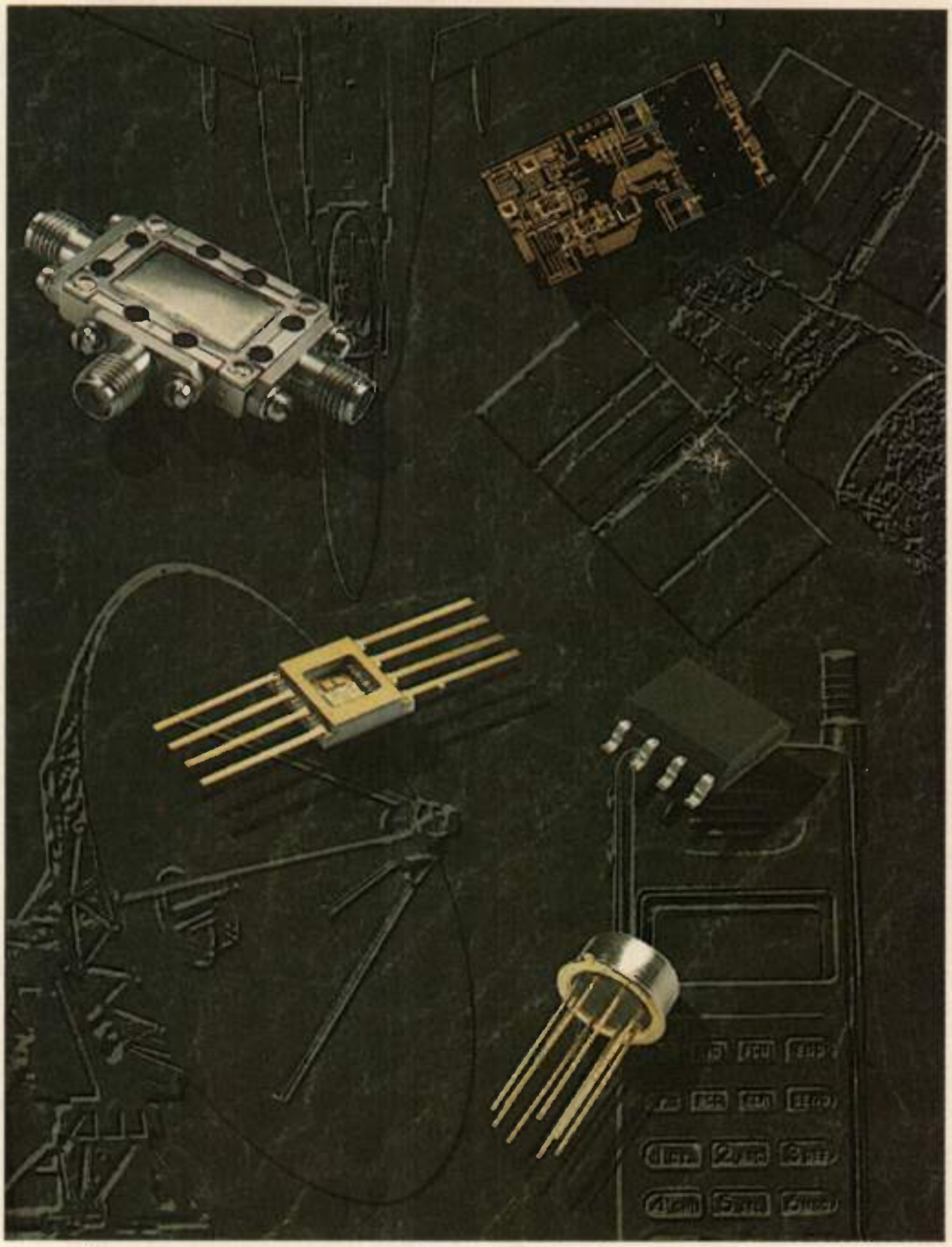

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are compatible with existing microelectronics that have high current and low voltages. The disadvantages are high resistive losses and they are larger than their electrostatic models (100 microns up to one millimeter) and they use more current. This research is sponsored by the National Science Foundation and by the Ford Motor Company.

Multichip Module Design – A Georgia Tech Research Institute group cautions that a shortage of established design rules could create EMI in the new generation of multichip modules (MCM) that involve high-density mixing of analog, digital and microwave devices. Georgia Tech analyzed the electromagnetic susceptibility of some modules. The findings could be the basis for test procedures, simulations and design guidelines to help minimize EMI problems and the expensive network process used when MCMs are not compatible. Their test showed that standard circuit simulation software could be used in analyzing EMI protection and the most resistive components. Supporting this effort, and to see if these MCMs are cheaper to manufacture and repair, is the Air force Office of Scientific Research and the Air Force Rome Laboratory System Technology and Integration Branch.

Digital Links Complete Networks – After testing initial digital radio installations Bell Atlantic Mobile (BAM) is offering digital cellular service in Philadelphia and Pittsburgh, Pennsylvania that is compatible with current analog phones. The digital system paves the way for future improvements in call quality, network accessibility, fraud protection, enhanced voice services, static-free calling,and increased privacy, Also new services such as receiving cellular messages, message waiting indicator for BAM Messenger ™ voice mail, caller ID and battery sleep mode are available.

New Office – Component Distributors Inc. who distribute RF/Microwave and high-tech components has announced a new office in Leesburg, VA. Contact person is Betsy Williammee, Tel: (703) 779-4995. Fax: (703) 779-4996.

New Division – Solid State Testing, Inc., has formed a new division called Burlington Microelectronics that will specialize in circuits that are no longer available but need to be reverse-engineered to meet required electrical and/or mechanical parameters. New devices that are not yet available from normal sources can be custom designed, assembled and tested. They do this with monolithic circuits, hybrids, and multi-chip modules. The parent company started as a testing lab for discrete components and integrated circuits 25 years ago then expanded as a procurement and manufacturing partner for its customers.

Subsidiary Purchased – Electro-Magnetic Processes, Inc., a subsidiary of Ferranti International has been purchased by an internal management group. The maker of antenna tracking systems will continue their operations at 20360 Plummer St., Chatsworth, CA 91311. Tel: (818) 882-6333. Fax: (818) 709-4668.

TELECOM 95 Call for Papers – TELECOM 95 will be in Geneva, Switzerland, October 3-11, 1995, and will have an Exhibition and Book Fair, Strategies Summit, and Technology Summit. Abstracts are due by August 15, 1994. Contact: FORUM 95 Secretariat, International Telecommunication Union, Place des Nations. CH-1211 Geneve 20 (Switzerland). Tel: +41 22 730 56 80. Fax: +41 22 730 64 44. Internet: forum-cfp@itu.ch.

FCC Gives Type Acceptance of 220 MHz Linear Modulation – E.F. Johnson Company has received approval for its infrastructure and mobile subscriber equipment using linear modulation at 220 MHz. This equipment utilizes 5 KHz bandwidths, compared to 25 KHz for current trunking systems. It provides five radio channels in the same bandwidth as one typical 800 MHz channel, with no loss of voice quality. In 1987 the FCC reallocated the 220-222 MHz frequencies as a narrow band for private land mobile use. The FCC has issued 3,900 licenses for 220 MHz systems.

High Speed Wireless LAN Proposal Approved – The IEEE 802.11 Standard Committee approved a high-speed physical layer (PHY) option for frequency hopping spread spectrum wireless LANs based on a proposal submitted by Proxim, Inc. The proposal called for four level Gaussian Frequency Shift Keying (4GFSK) modulation to achieve high-speed data transmission in frequency hopping wireless LAN environments with a fallback to two-level GSFK (2GSFK) for communications with lower speed frequency hopping systems. This dual mode approach, which does not cause interference between higher and lower speed wireless environments, has been implemented in Proxim's LAN2 product family and are FCC approved. In evaluating potential high-speed PHY layer implementations, the 802.11 Committee ratified a 2.0 Mbps/1.0 Mbps dual data rate standard for both frequency hopping and direct sequence spread spectrum systems. This decision follows the Carrier Sense Multiple Access with Collision Avoidance Media Access Control layer foundation protocol approved at the last plenary meeting in November 1993. With the specification of this new high-speed PHY layer modulation and data rate standard now decided, the path is cleared for the 802.11 Committee to meet its target of completing a Draft Standard document by late 1994.

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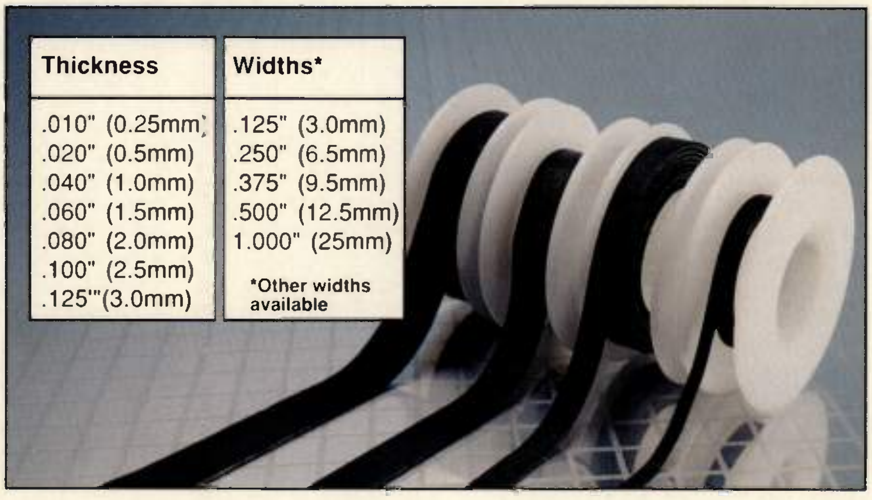

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INFO/CARD 19

RF Radiation Hazards

By Andy Kellett
Technical Editor

How hazardous is RF radiation? There are probably more emergency room visits due to toothpick mishaps than RF accidents, and any link between RF and disease (mostly cancer) is many many times weaker than, for instance, the link between fat intake and heart disease. But for RF engineers and others who are exposed to more than typical ambient RF fields, it is prudent to try to assess the risks. This report will review some RF exposure standards, examine the facts surrounding the concern about links between RF and cancer, and review some of the situations that are most likely to place someone in strong fields.

Current RF exposure standards address the thermal effects of RF radiation. The same losses that occur in circuit elements at RF frequencies occur in human tissue, and those losses eventually take the form of heat.

The latest ANSI standard for RF radiation exposure, C95.1-1991, specifies maximum power densities, electric and magnetic field strengths, and contact currents. A new feature of C95.1-1991 is the definition of separate limits for "controlled" and "uncontrolled' exposures. Situations in which people intentionally expose themselves to RF radiation, are considered controlled, whereas those situations where a person unwittingly is exposed to RF radiation are considered uncontrolled.

Taking the limits for power density as an example, a person in a controlled environment can be exposed to 10 W/m^2 from 100 to 300 MHz, while a person in an uncontrolled environment can only be exposed to 2 W/m^2 in the same frequency range. Above 300 MHz, both exposure limits ramp up to 100 W/m^2, but the controlled environment limit reaches that density at 3 GHz, while the uncontrolled limit does so only after reaching 15 GHz. After 15 GHz both controlled and uncontrolled exposure must be no greater than 100 W/m^2. Below 100 MHz, C95.1-1991 specifies exposure limits in terms of independent electric and magnetic field strengths.

Whether it is due to the relatively new ANSI standard or not, manufacturers of field measurement equipment have noticed some changes in what their customers demand. Users of industrial RF equipment such as induction heaters are purchasing more monitoring equipment says Bob Johnson, Manager of Instrument Applications at Narda Microwave Corp. Other standards such as Sweden's video display terminal emissions standard are also driving sales says Burton Gran, President of Holaday Industries.

RF and Cancer

Epidemiologist Dr. David Savitz of the University of North Carolina School of Public Medicine says that a link between electromagnetic (EM) emissions and cancer is certainly not dismissable, but not convincing either. Most of the epidemiological research regarding links between EM fields and cancer have been done for very low frequency signals such as the 60 Hz radiation from power lines. To determine what link exists between cancer and EM radiation at RF frequencies, the Scientific Advisory Group on Cellular Telephone Research has commissioned several studies. Among them is an epidemiology study under the direction of Dr. Kenneth Rothman and Dr. Nancy Dreyer of Epidemiology Resources, Inc., that will study cellular users in several parts of the U.S. "This study is unique in that it is a very large cohort study," says Dr. Dreyer.

In cohort studies, a population of disease-free subjects is selected, and as they naturally expose themselves to different circumstances they are monitored for disease. Most prior studies were case control studies, in which researchers selected patients already stricken, and grouped them according to similar exposure circumstances. While cohort studies avoid biases in selection of patient grouping, they also take much longer than case control studies. "This study is being done in the spirit of surveillance, which means it will likely continue for long time," says Dreyer.

Assuming there is a link, by what mechanism could RF cause cancer? RF fields do not have enough energy to ionize atoms, and by definition, athermal fields produce minuscule heating.

A mechanism that casts RF radiation in the role of a cancer promoter has been suggested by Dr. W. Ross Adey. Adey points to research that has shown enhancement by RF and ELF fields of chemical cancer promoters that act at cell membranes. Based on studies in his group and at Oxford University, Adey suggests that RF radiation extends the life of free radicals set free in the course of all chemical reactions. This alters the rate and products of chemical reaction at the cell surface, interfering with intercellular communication and allowing mutated cells to begin uncontrolled growth.

Hot Spots

Whether in the thermal or athermal regime, RF engineers are more likely to encounter unusual RF radiation levels. What are the "hot spots" around which engineers should be careful? Jim Maginn of Amplifier Research points out that major sources of RF and microwave leakage are waveguide flanges and bad cable connections. In a bad cable connection, the most radiation emanates not from the end of the cable, but from the cable, just before the connector notes Maginn.

Another hot spot that RF engineers are likely to encounter is high power microstrip circuitry says Narda's Johnson. Working close to an energized microstrip circuit, to inspect a transistor for instance, can expose a person to high field intensities.

Such seemingly obvious hot spots as powered radar and broadcast transmitters have been known to catch the unwary. Randy Ridley, Technical Editor for *RF Design*'s sister publication *Communications*, tells the story of how he and his colleagues worked in jackets at the base of one tower, while a tower maintenance man climbed down from an adjacent FM broadcast tower drenched in sweat and complaining of the heat. It didn't take long for the maintenance man to realize he had climbed an energized tower.

Sometimes the largest danger in RF equipment are phenomena that are one step removed from RF phenomena. Electrons accelerated across potentials greater than a few thousand volts or more emit ultraviolet and x-ray radiation when they decelerate, as when electrons in a high power tube hit an anode. Some hazards are even twice removed, for instance, Amplifier Research's Maginn described how a wrench was thrown into a wall by a steam explosion caused by a failed cooling system.

Whether you work in high intensity fields or merely in the ambient RF environment, assesing the risks associated with RF depends on individual circumsatnce. However, it seems that very few people have much to fear from RF. **RF**

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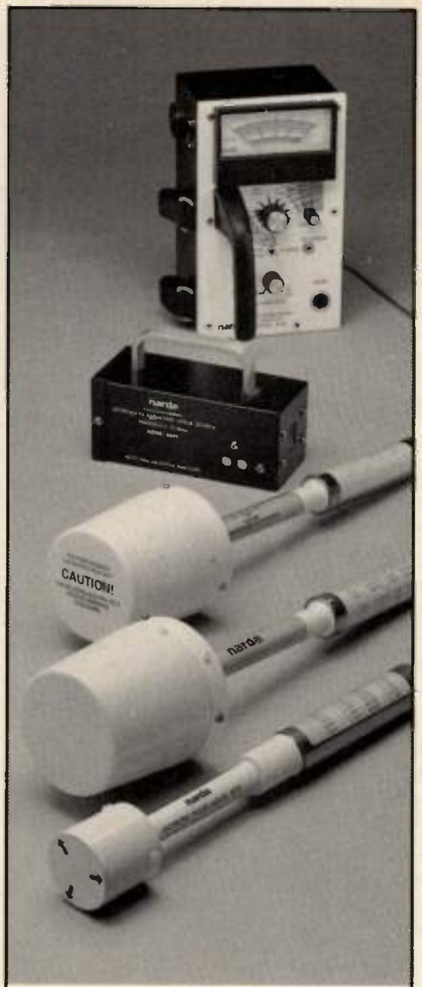

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Transceiver Chip Simplifies GSM Cellular Design

By Reynolds E. Jenkins
AT&T Microelectronics

With the advent of digital cellular service, cellular phone designers face increasing pressure to create smaller, lighter and less costly models that provide longer talk and standby times. Until recently, however, designers also faced the problem of meeting these goals using RF sections incorporating discrete components and very low levels of circuit integration. New digital cellular standards compounded the problem by requiring far more complex signal processing than that required in the analog standards. To meet both the demands of the marketplace and the requirements of the digital cellular telephone standards, much higher levels of circuit integration are required — in both the baseband circuits and the radio section.

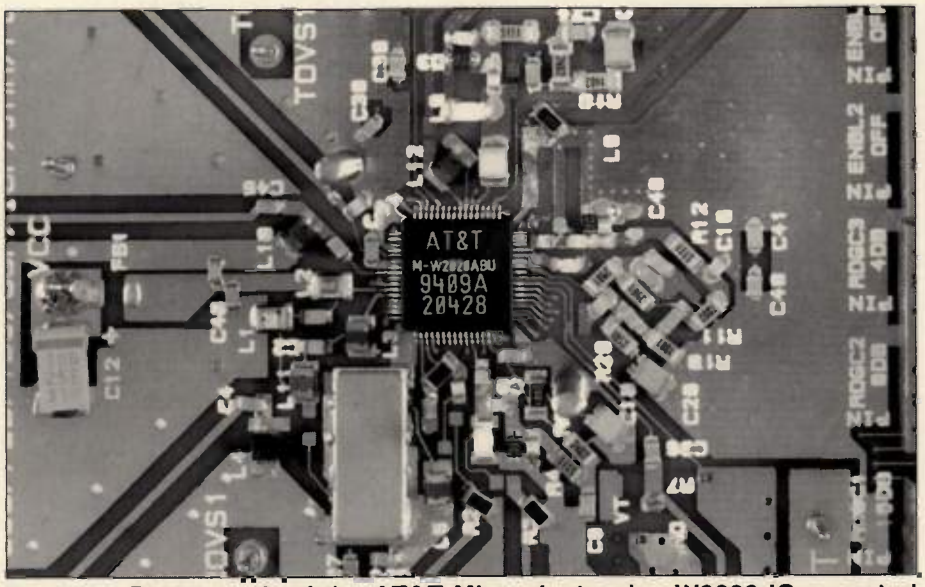

Figure 1. Photograph of the AT&T Microelectronics W2020 IC mounted on the EVB2020 evaluation board.

A digital cellular telephone's transceiver function is an area of the RF section traditionally resistant to higher levels of integration. In designing its W2020 transceiver chip, AT&T Microelectronics achieved an unprecedented level of transceiver circuit integration. The W2020 combines on a single substrate the RF receiver mixer, IF amplifier strip, quadrature demodulator, modulator, UHF synthesizer, two fixed-frequency oscillators and control circuits that optimize power consumption. The addition of a power amplifier, low noise amplifier (LNA) and filters are all that is required to complete the digital cellular telephone's radio section.

Although digital cellular telephones built for North American use follow the IS-54 standard, the GSM standard dominates world-wide digital cellular telephone service. Originally representing "Group Speciale Mobile", the term GSM now refers to the standard being the "Global System for Mobile communications." Originally a European standard, over 40 countries representing a total of 1.5 billion prospective users now accept it as their national digital cellular standard. To date, 25 GSM networks covering 17 counties have been deployed. In Europe alone, the number of digital cellular subscribers has reached the 1.2 million mark. By the end of this year, there should be more than 3 million subscribers worldwide.

Designers of North American IS-54 cellular telephones have several chip sets from which to choose but the levels of integrationrepresented by them is relatively low. However, this is more a result of the IS-54 requirement for dual analog/digital capability than it is a lack of technology on the part of the chip designer. For example, the IS-54 standard requires that when the phone operates in its analog mode, full duplex operation be supported. The need to both transmit and receive simultaneously makes it virtually impossible to integrate both functions on the same chip because of the resulting need for isolation between them. The GSM standard does not suffer from this limitation because it specifies a digital service using time division multiple access (TDMA) in which the transmitter and receiver are never on at the same time. Without the need to consider the isolation between transmit and receive circuits, very high levels of circuit integration can be achieved.

Similar to the digital portion of the IS-54 standard, the GSM standard also makes use of TDMA and speech compression to increase channel capacity. GSM cellular telephones transmit in the 890 MHZ to 915 MHz band and receive in the 935 MHz to 960 MHz band, very close to the bands assigned to cellular service in North America. It should be noted that the 25 MHz transmit and receive bands were recently expanded by an additional 10 MHz with the new spectrum referred to as extended GSM (EGSM). The GSM standard specifies Gaussian minimum shift keying (GMSK) modulation with a 200 kHz channel spacing. Although the standard specifies five different transmitter power classes, ranging from 0.8 W to 20 W, the most commonly used in handportable telephones is Class IV at 2.0 watts.

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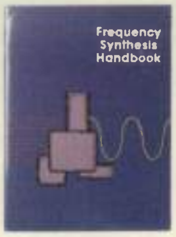

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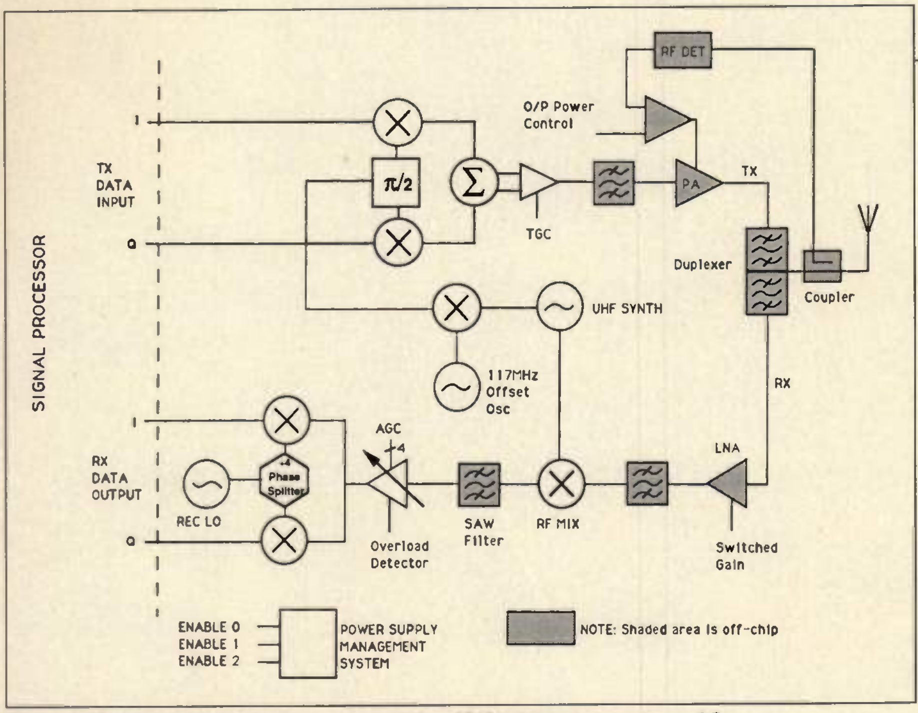

Figure 2. Functional diagram of the W2020 transceiver IC.

Receiver Architecture

The first decision to be made was the architecture of the receiver. There are three accepted choices for a receiver design — direct down conversion, single IF solution and dual IF solution.

For direct down conversion to function properly, the input signal must fall into a relatively narrow range of amplitudes. However, the GSM standard requires the receiver to accept input signals of much greater instantaneous dynamic range than can be handled by direct down conversion techniques. Also, since in a direct conversion architecture the mixer is driven by a local oscillator at the received frequency, leakage from the mixer to the antenna can be a problem. For these and other reasons, direct down conversion is not a practical choice for a GSM cellular telephone. This limits the choice of receiver architecture to single and dual IF solutions.

Because of the many tradeoffs that must be considered, choosing between a single and a dual IF solution can be very difficult and depends to a great extent on the receiver's design goals. For a battery powered cellular telephone, cost and power consumption are critical concerns. A dual IF solution requires a second mixer and IF strip, which adds to the cost of materials and total receiver power consumption. Also, a single IF solution generates fewer spurious signals that can prove difficult to contain in a single-chip transceiver solution. Admittedly, the kind of high performance SAW filter required in a single IF solution does cost more than the lower performance filters that can be used in a dual IF solution. However, it still costs less than the pair of filters that would be required for a dual IF solution. These factors clearly point towards the use of a single IF solution receiver architecture in a GSM cellular telephone.

With only a single intermediate frequency, its choice plays a large part in the ultimate success of the design. Its selection depends, in part, on economic concerns and, in part, on performance. Choosing a relatively high IF implies difficulties with packaging and isolation between circuits implemented in the same substrate. These problems point toward an IF of less than 100 MHz.

Another consideration is the cost of the IF bandpass filter. This filter, which must significantly attenuate blocking signals near the intermediate carrier, must have steep-sided bandpass characteristics. Only SAW filters can economically provide the needed performance. Choosing a non-standard intermediate frequency implies the use of a custom SAW filter, which means higher cost and single sourcing. However, most major SAW filter manufacturers offer 71 MHz filters as standard models. Thus, by choosing 71 MHz as the IF, packaging and isolation problems are minimized and high performance SAW filters can be obtained from several sources, ensuring lowest cost and highest availability.

Signals received by the GSM telephone's antenna flow into the receiver circuits through a LNA and RF filter. Consideration was given to including the LNA in the W2020 transceiver to simplify the overall telephone design. However, doing so would also require additional I/O pins for connections to external inductors, capacitors and ground connections. It would also require an interface to the RF filter between the amplifier and RF mixer. Providing these I/O pins would not be possible if the W2020 was to be packaged in the 64-pin TQFP originally envisioned. The only practical solution is to use a discrete LNA. Fortunately, there are many low-cost discrete devices that can be used very effectively in a GSM telephone. The required RF filtering can be achieved with a combination of dielectric and SAW filters.

Transceiver Design

Decisions made during the design process were based on several factors, but two of the more important were the need to provideoptimum performance and to operate well on a 2.7 V supply. A block diagram of the resulting GSM transceiver is provided in Figure 1. Incoming signals received at the telephone's antenna are amplified by an LNA and passed through a filter to the RF mixer, which is a single balanced design. Though a double-balanced mixer might provide marginally better performance, it cannot readily be implemented in an integrated circuit intended to operate at 2.7 V.

A UHF phase-locked loop synthesizer drives the mixer at a frequency 71 MHz above the received frequency. With the exception of the varactor, resonator and loop filter, the synthesizer is completely integrated into the W2020 transceiver. All of the counters used to set the channel frequency and to divide the reference frequency are realized using ECL techniques. This reduces power supply noise to levels low enough to permit the synthesizer to be integrated into the W2020 substrate. The varactor could not be implemented on-chip because the bipolar process used is not well suited for high-Q varactors. The loop filter components and the resonator are passive and cannot economically be implemented in the bipolar substrate.

Digital cellular telephony uses quadrature modulation to carry the information being transmitted. On receive, the baseband circuits require an in-phase "I" and a quadrature "Q" signal, which is phase shifted precisely 90° from the I signal. To provide the baseband circuits with the I and Q signals they require, the

transceiver's receive section includes a quadrature demodulator and local oscillator. The demodulator implemented in the W2020 transceiver uses a divide-by-four strategy to generate the required phase shift. Unlike the divide-by-two strategy commonly used, dividing by four provides a 90° phase shift independent of duty cycle. Dividing by four requires a local oscillator frequency four times the IF, which in this implementation equals 284 MHz. The 71 MHz IF generated in the RF mixer passes through the IF SAW filter into the quadrature demodulator, which directly down converts the IF signal into baseband I and Q signals.

Among the specifications in the GSM standard is a requirement that the receiver section operate properly with input signals ranging from –102 dBm to –15 dBm. The standard also requires the cellular telephone to report the received signal strength to the base station. Measurement of the received signal strength is performed in the baseband, which continuously monitors the amplitude of the I and Q signals received from the demodulator and the ACG setting.

To prevent saturation when the received signal is at the upper end of input amplitude range, an AGC circuit adjusts the gain of the IF amplifier to maintain its output within the desired range. The actual gain setting is determined by the baseband, which digitally controls the AGC in 4 dB increments over a four-wire bus. Overall, the system provides 60 dB of dynamic IF amplifier gain control range.

Also included in the receiver is an overload detection circuit. An overload detector built into the AGC circuit provides a logic signal through an output pin when the received signal is so strong that the AGC circuit cannot adequately reduce the gain. The signal provided by the overload detector is fed to the baseband for processing. If the telephone designer chooses to exploit the overload detector signal, the baseband can be programmed to reduce the gain of the receiver's LNA when necessary to maintain the appropriate receive path signal levels.

During transmit, the baseband circuits provide I and Q signals, which are applied to a quadrature modulator integrated into the W2020 transceiver chip. The modulator uses a direct modulation technique to reduce the filtering requirements in the transmit path. Direct modulation also eliminates an up conversion mixer, reducing bill of materials costs and power consumption. Moreover, small adjustments in the output's center frequency can be made in the baseband

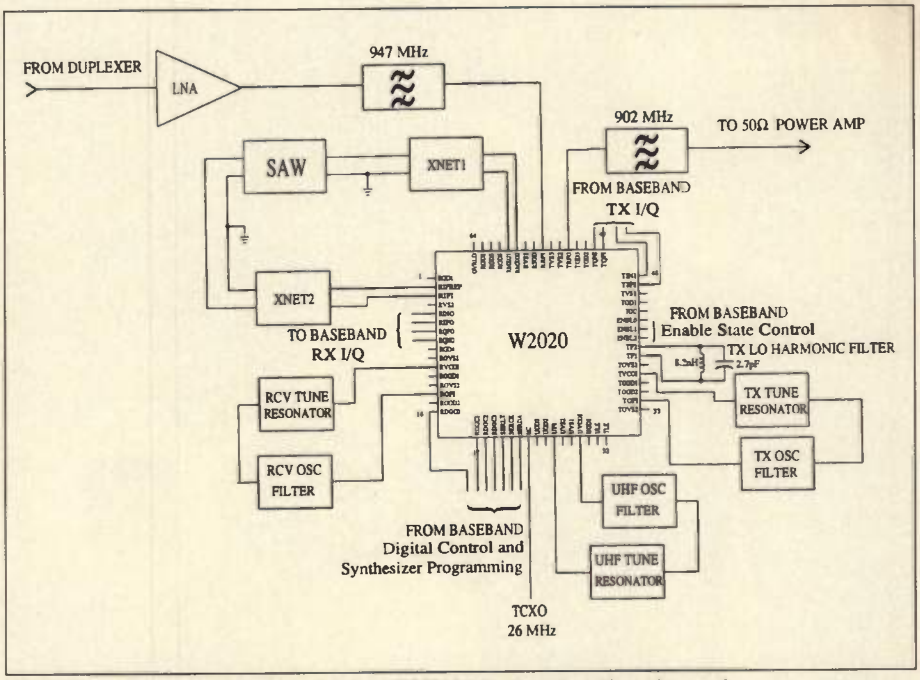

Figure 3. Block diagram of the EVB2020 evaluation board.

circuits, which permits the relaxation of the reference frequency precision. Some external filtering is required to avoid noise in the receiver, with implementation left to the telephone designer.

Transmit channels are offset from the corresponding receive channels by 45 MHz. Ordinarily, a transceiver would contain separate synthesizers for the transmit and receive channels. Instead, the W2020 transceiver contains just one channel-selecting UHF synthesizer, which is locked to a frequency equal to the sum of the received frequency plus the 71 MHz IF. The transmitter's carrier frequency results from mixing the UHF synthesizer's output with a 117 MHz offset oscillator.

The offset oscillator's frequency is 1 MHz above the desired offset frequency, which is equal to the 71 MHz IF plus the 45 MHz transmit-receive separation (a total of 116 MHz). However, the decision was made to use a 117 MHz local oscillator because that frequency could be generated in a more elegant and cost effective circuit than could 116 MHz. Moreover, the frequency agility of the UHF synthesizer allows it to shift 1 MHz during the interval when the telephone is switching between transmit and receive.

Like the UHF synthesizer itself, with the exception of the varactor and passive filter and resonator, the offset oscillator is implemented on-chip. Moreover, the offset oscillator draws power only during transmit, which means that it cannot generate spurious signals when the telephone is receiving. It also reduces power consumption — an important consideration in a battery powered portable telephone.

It should be noted that thought was given to simply shifting the UHF synthesizer down 116 MHz to directly produce the required transmit frequency. However, analysis showed that the synthesizer could not reliably shift back and forth 116 MHz in the allotted time when operating at 2.7 V. Also, the presence of an offset mixer isolates the UHF synthesizer from the modulator, thereby reducing pulling on the synthesizer.

Another feature built into the W2020 transceiver chip is a transmitter gain control circuit that permits the drive to the power amplifier to be adjusted dynamically. The power amplifier's drive is controlled through the application of an externally generated control current. This arrangement affords the telephone designer considerable flexibility in implementing the control circuit.

Evaluation Board

In an era when product life cycles continue to decrease, bringing a product to market as quickly as possible is critically important to that product's success. Even with the latest in design aids, there are still bottlenecks in the product development cycle that can significantly lengthen the process. Fabricating a prototype printed circuit board, for example, can require several weeks when the product is as complex as a cellular telephone.

Recognizing the need to shorten development time, AT&T created the EVB2020 evaluation board, shown in Figure 2, which provides an immediate vehicle that designers can use to ana-

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lyze the W2020's performance. The evaluation board provides a realistic system-level operating environment and includes many of the filters, matching networks and other external components required to implement a functioning radio. Using the board, a designer can evaluate the chip's performance as a stand-alone circuit or the interaction between the chip and the baseband.

The evaluation board's block diagram is shown in Figure 3. The external networks XNET1 and XNET2 are passive interface networks that impedance match the RF mixer's output to the SAW filter input, and the SAW filter's output to the IF amplifier's input.

Isolation between the SAW filter's input and output nodes is critical if the desired filtering is to be achieved. Both the W2020's internal layout and its pinout assignments support the required level of isolation. Here the RF mixer output and IF amplifier input are located across one corner of the W2020's package. This allows the SAW filter to be mounted close to the W2020's package. Even so, the designer must still provide a printed circuit board layout that maintains high levels of isolation at critical nodes such as these. The SAW filter's input to output isolation, for example, can be enhanced by surrounding it with the printed circuit board's ground plane. By so doing, the isolation between the XNET1 and XNET2 components can be significantly improved.

The RX and TX filters and resonators shown in Figure 3 consist of varactor diodes with tuning provided by passive LC networks and loop filters implemented with RC networks. The UHF tune resonator consists of varactor diode and either a printed microstrip inductor or ceramic resonator and capacitor. The UHF loop filter consists of an RC network. The use of small surface mount components is critical in these applications because it allows the resonator components to be placed very close to the W2020's VCO inputs. Keeping these close to the W2020 minimizes the parasitic capacitance to ground and the inductance of the printed circuit board's copper traces. Keeping the parasitic reactances to a minimum affords the UHF synthesizer a tuning range wide enough to eliminate the need for frequency trimming during assembly.

In any RF design, care must be taken to eliminate ,as much as possible, signal leakage paths. A common trouble spots, for example, are IF and RF local oscillator synthesizers. Because the counters and prescalers are digital, their signals are rich in higher order harmonics that can find sneak paths into other circuits. However, since the IF and RF synthesis is performed on-chip in the W2020, the associated harmonics are contained within the chip and, therefore, do not radiate to other circuits on the printed circuit board.

Keeping the IF and RF synthesizers on-chip reduces, but does not eliminate, potential interference to other circuits from signals produced by the synthesizers. For example, the synthesizer's PLL phase comparator signals must come off-chip to the external loop filter components. These components form networks that filter and integrate the phase comparator signals, thereby significantly reducing their harmonic content. To be fully effective, however, these components must be located as close as possible to the phase comparator I/O pins. The recommended positioning of these components is demonstrated on the EVB2020. Another off-chip component is the synthesizer's tuning resonator but since it serves as the VCO's tank circuit it produces a nearly-pure sine wave with very little harmonic content.

Isolation between the transmitter output and receiver input is not ordinarily a design concern. Because GSM uses TDMA to increase channel capacity, the transmit and receive functions occur in different time slots and are, therefore, not active at the same time.

Because of the complexity of a GSM cellular telephone, the use of a multilayer printed circuit board is essential. An added benefit of using a multilayer board is that isolation between baseband and radio signals can be enhanced through careful placement of ground traces between various signal lines. In the EVB2020, for example, all the RF signal lines reside on the top layer with most of the external impedance matching network, oscillator and filter components mounted on the top surface. The board's bottom layer is used to carry all the digital I/O signals between the radio and baseband sections. The bottom layer also carries the I and Q signals, but these are isolated from the digital lines by the board's ground plane.

The EVB2020 includes 50 ohm microstrip lines for the transmit and receive RF and IF signals. Microstrip lines are also provided that permit the on-chip oscillators to be overdriven by an external signal generator. Matching networks provided for the RF mixer input and output as well as the transmitter output permit the use of standard 50 ohm laboratory equipment.

The W2020 transceiver provides an ideal foundation for the radio section of a power-efficient GSM cellular telephone. Most of the critical design challenges ordinarily faced by an RF designer have already been overcome in the circuit implementation. The proven design and performance of the chip reduces engineering costs and helps shorten time to market.

Additional information on this chip set can be obtained by contacting the author at the address below, or by circling Info/Card #101. *RF*

About the Author

Reynolds E. Jenkins received a BSEE from Pennsylvania State University and an MSEE from Clemson University. He has been with AT&T Microelectronics for 10 years. Originally an applications engineer, he is now product manager for wireless products. He can be reached at AT&T Microelectronics, 2525 North 12th Street, Reading, PA 19612-3396; tel. (610) 939-7447, fax. (610) 939-6069.

Designing a Low Cost GPS LNA Using the NE68519

By Terry Cummings
California Eastern Labs

This article is intended to demonstrate the NE68519 in a low cost GPS LNA. The NE685 is a low cost NEC silicon bipolar transistor which has an f_T of 12 GHz and yields a typical NF of 1.5 dB at 2 GHz when biased at 3V and 3mA. The purpose of this note is to show how one can design a simple L band LNA for low voltage power supply conditions. The design approach will be discussed. While this design may not yield the optimum design solution for all GPS applications, it does introduce the reader to a few important RF and microwave techniques that can be applied to GPS as well as other applications.

The amplifier is a single stage design that yields a 1.8 to 1.9 dB NF and 10.5 dB gain at 1575 MHz. A second LNA will be presented with a ceramic filter at the input resulting in a NF around 2.6 dB and 10 dB of gain. Bias conditions are 3V and 3mA for power consumption per dB that is one quarter that of commonly used GaAs MMICs for this application. Low cost and small size are considered throughout this design using commonly available SMT components.

The entire design was initially done on CAD, using EEsof's non-ideal elements section-by-section in the "Tune" mode. This approach yielded a solution that was extremely close to the desired solution. Noise and gain circles are not considered since a narrowband design approach was used to achieve best noise match to the device. Stability circles are discussed and studied through CAD and not derived.

Distributed lines were used to avoid costly SMT inductors. A 14 mil FR4 substrate was chosen for cost considerations and its ability to be laminated to a multilayer board as used in the DSP sections of a typical GPS receiver. Minimum trace width was set to 10 mils for easy PCB processing and the entire amplifier occupied a space which was less than a 1/2 x 1/2 inch. The single stage LNA used a minimum number of 0603 SMT capacitors and resistors for repeatability and minimal size.

Characteristics of LNAs

An LNA is a design that minimizes the NF of the system by matching the device to its noise matching impedance, or Gopt. Gopt occurs at an impedance where the noise of the device is terminated.

All devices exhibit noise energy. To minimize this noise as seen from the output port, one must match the input load to the conjugate noise impedance of the device. Otherwise the noise will be reflected back from the load to the device and amplified.

While this gives a minimum noise figure, it often results in slightly reduced gain as well as possibly increasing the potential for instabilities. Noise match often comes close to S_{11} conjugate ($S_{11}{}^*$) under non-feedback conditions. As a result, the input impedance to the amplifier will not be matched to 50 ohms.

Γ_{opt}, as presented in data sheets, is the actual measured load at which the minimum noise figure is found. The output load is rarely given since it can be easily derived through a simple equation. The reduced form of this transform is given in equation 1.

$$\Gamma_L{}^* = \frac{S_{22} - \Gamma_{opt}\Delta}{1 - \Gamma_{opt} S_{11}} \qquad (1)$$

$$\Delta = S_{11}S_{22} - S_{12}S_{21}$$

A further complication on LNA design is that the input load of the amplifier is usually less than ideal. It is either connected to an antenna, which can change its impedance with its changing environment, and/or to a filter which by the very physics of a reflective network will have a horrible match out of band. These mismatches could cause the device to become unstable out of band and in some cases in band. As the gain of the device increases, the difficulties in yielding a stable design become increasingly more challenging.

To avoid overloading the LNA, an input filter is commonly used. Since the device is not matched to S11*, the input of the LNA will not be 50 ohms. This can cause distortions in the pass band of a filter when connected to the input of the LNA, as filters are intended to operated in their characteristic impedance, typically 50 ohms. Fortunately the NE685's optimum noise match and S11* are not too far apart. As a result, the filter passband is only slightly distorted.

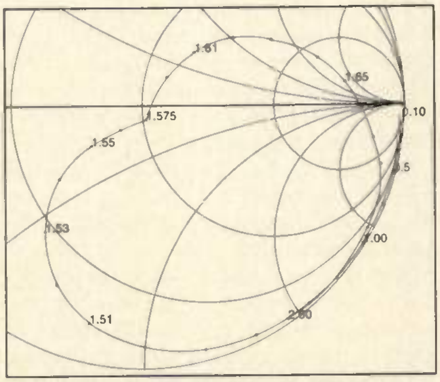

Figure 1. S_{11} of ceramic filter.

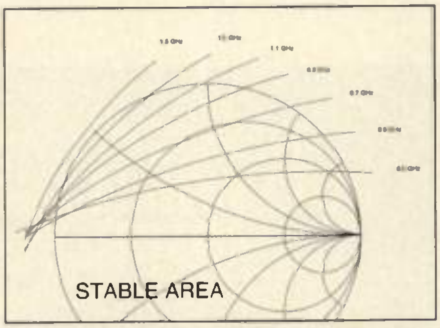

Figure 2. Input stability circles for NE68519 loaded in 50 ohms.

Input structure

The LNA uses a simple input structure that provides a noise match at 1575 MHz at the expense of gain flatness and improved stability. The first step of the design is to plot the stability circles of the device as a function of Γ_{opt}. While plotting the circles, keep in mind the out-of-band gain response of the device and its impact on S_{11}.

In our case, a ceramic filter's impedance plot is a cardioid rotated about 45 degrees clockwise when measured at the leads of the filter. Note that the input

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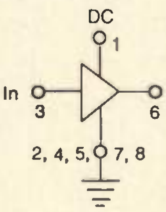

Freq.(GHz)	.5-.8	.8-1.0	1.0-2.0	2.0-2.5
Gain (dB) typ.	14.0	17.0	18.0	16.0
Max. Output (dBm) @1dB Comp. typ.	+18.0	+18.5	+17.5	+17.0
I P 3rd Order (dBm) typ.	+27	+27	+27	+27
VSWR Output typ.	1.5:1	1.7:1	1.7:1	1.5:1
VSWR Input typ.	6.4:1	2.8:1	2.0:1	1.4:1

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impedance will start out high at low frequency and then go to its terminated impedance of 50 ohms during resonance at 1575 MHz then continue to rotate clockwise as frequency goes up. We plotted just past 2 GHz since our initial stability study showed that the problem areas were below 2 GHz. Since these filters are coaxial resonators, one would expect that the locus would continue to rotate around with a radius of near 1 until a weak second resonance occurs at $3f_0$. This is well above the device's unstable region in our configuration so we can assume an ideal filter situation which in our case will be worst case.

Next we considered the NE685's stability regions for the input. A stability plot of the NE685 was made to determine what impedances would potentially cause instability. As you can see from Figure 2, the circles enter the Smith Chart counter clockwise from right to left as frequency increases. At first it appears that the unstable region starts out at high impedances for low frequencies. Careful examination shows that the unstable region follows a relative low resistance contour. These circles show that we want to avoid an inductive reactance as frequency goes down.

This is fortunate since the filter's input impedance maps in the opposite direction. We must be careful with the input matching topology since we do not want to transform either the 50 ohm input or the filter into this unstable region. With this in mind we can choose a topology using basic filter concepts that will provide the desired matching as well as avoid transforming out of band areas in the unstable region.

If we were not so fortunate and were forced to accept the filter's undesired out of band impedance response, we could approach the problem in several ways. The most common method uses a resistor to create a lossy network on the opposite port to absorb the negative resistance caused by the input match. However this method will not only stabilize the design, but can degrade the overall performance of the amplifier.

A more selective approach uses a selective lossy filter: a simple LCR that resonates at the conditionally unstable region. This technique will only work if the Q of the resonator relative to the passband of the amplifier is appropriate.

Choosing a topology can be affected by the bias network limitations. We know already a shunt L could cause problems since it becomes a low impedance at low frequencies. Since the base bias current is small, we can pretty much ignore the reactive effects of the bias circuit by using fairly large resistor values.

We can now consider the match for the input. In this case we desire to obtain a noise match and not S11*. We also want to choose reactive elements that transform out of band impedances away from the stability circles. Many powerful tools are useful for this process. For example, using the "Tune mode" in popular software packages such as EEsof's Touchstone, one can watch what happens to the match by trying out various element topologies and their values.

A bipolar transistor is naturally a low pass element and therefore requires a DC block at the input. To offset the high gain at lower frequencies a high pass element is needed. A capacitor can provide this and at the same time will provide the necessary DC blocking. Fortunately the value of the required capacitor is in a practical range. When using a capacitor under 10 pF, consider that its tolerance will be around ±0.5 pF. As a result, too small of a capacitor might make this element too sensitive relative to its tolerance. However, a smaller value of capacitance has the advantage of avoiding low impedance transformations at lower frequencies. On the other hand, too large of a capacitor will cause undesired low frequency gain. A compromise value of 5.6 pF was chosen for this series element, which results in a normalized impedance of 0.36j.

In order to show the result of our first matching element, both on a Smith Chart and mathematically, we also simplified the network as a one port circuit. This one port network was created by absorbing the 50 ohm input load with a 5.6 pF capacitor. The math of this process was easily done on Mathcad using ideal elements and is given as follows:

i) Starting out with a 50 ohm load and a 5.6 pF capacitor in a 46.06 ohm system: Z_0=46.06 Ω, f=1575 MHz, R=50 Ω

$$Z_1 = \frac{1}{j2\pi f \bullet 5.6x10^{-12}} + R \qquad (2)$$

$$= 50 - j18.056\ \Omega$$

$$Z_{1-46} = \frac{Z_1}{Z_0} = 1.086 - j0.392\ \Omega$$

$$\Gamma_{1-46} = \frac{Z_1 - Z_0}{Z_1 + Z_0} = 0.189\angle -67.045^\circ$$

The next element we chose was another series element that would be charted to provide a starting point for the next transformation as well as lowering the impedance of the network for minimizing pad parasitics for resistive bias elements.

By first having a small, low impedance transformation, one can minimize the effects of component placement as well as reduce the effects of relative high impedance shunt element values which supply bias at this point. Using CAD software, it is easy to flip from one transmission line impedance to another.

The model we used for this method was a one port network. By using a one port network, one can attempt to choose element values which result in an output match that follows the desired noise match of the transistor. After trying vari-

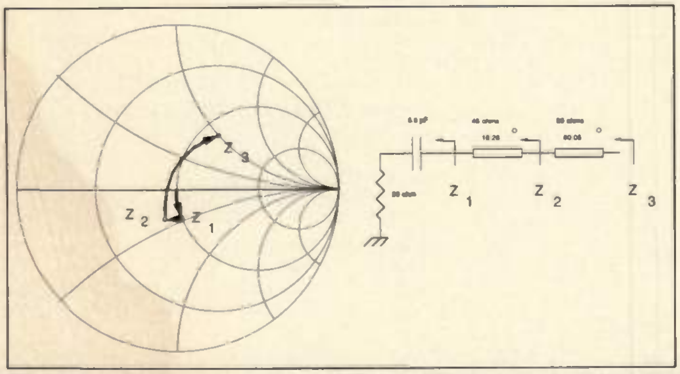


Figure 3. Final mapping of Z1 to Z3 normalized to 50 ohms.

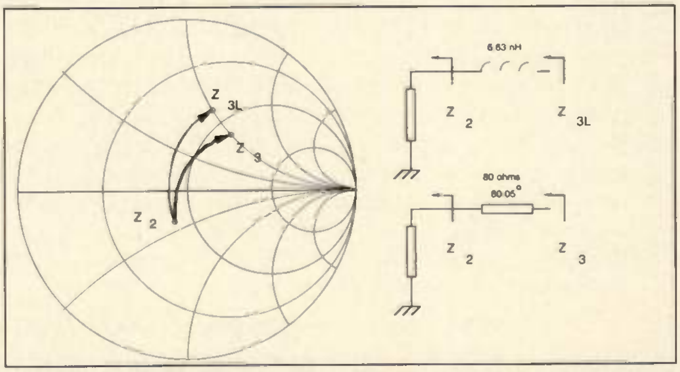
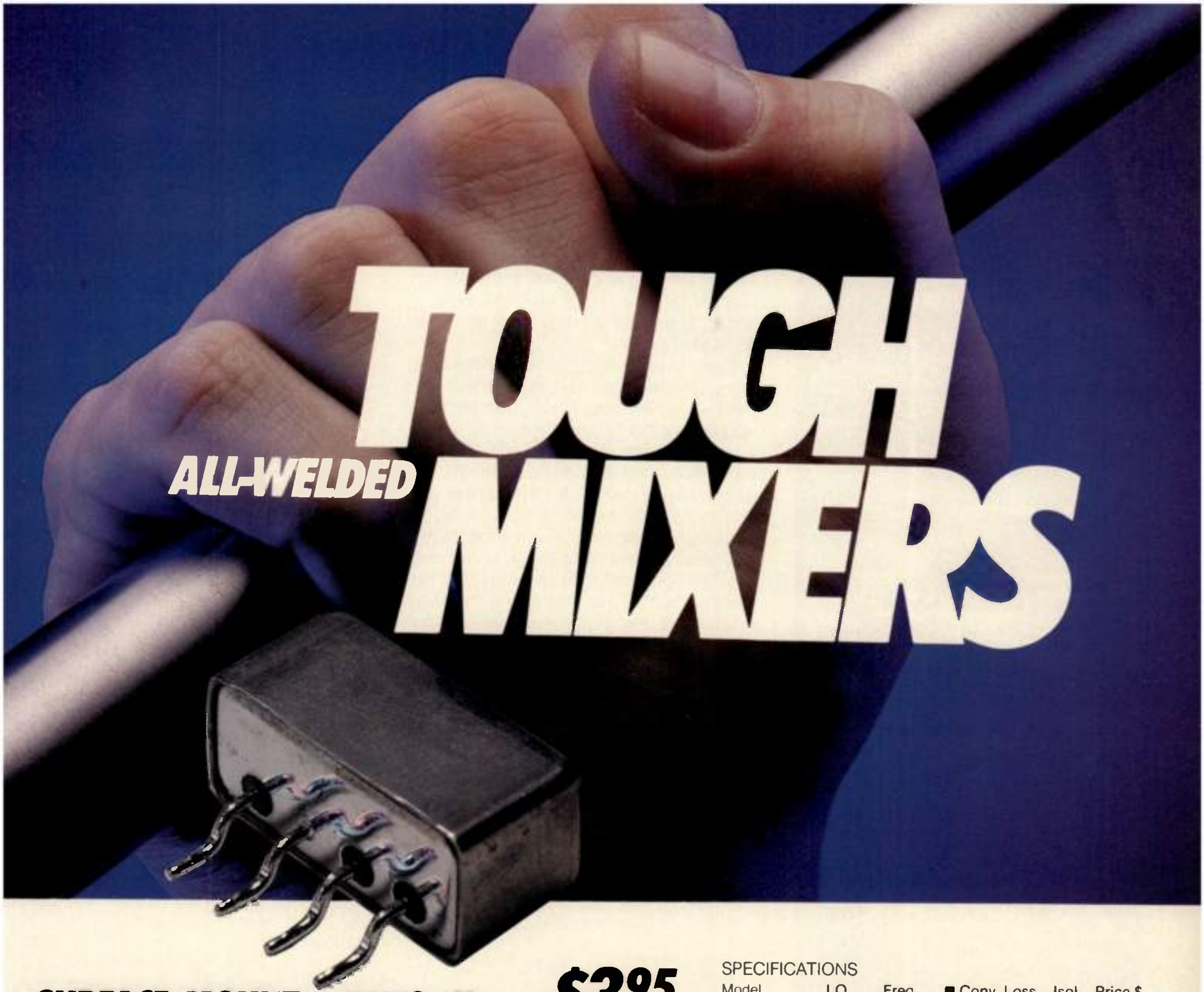


Figure 4. Z trajectory on a Smith Chart for L and an 80 ohm transmission line.



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TUF-3MH	13		5.0	0.33	46	8.95
TUF-3H	17		5.0	0.33	50	10.95
TUF-1	7	2-600	5.82	0.19	42	3.95
TUF-1LH	10		6.0	0.17	50	5.95
TUF-1MH	13		6.3	0.12	50	6.95
TUF-1H	17		5.9	0.18	50	8.95
TUF-2	7	50-1000	5.73	0.30	47	4.95
TUF-2LH	10		5.2	0.3	44	6.95
TUF-2MH	13		6.0	0.25	47	7.95
TUF-2H	17		6.2	0.22	47	9.95
TUF-5	7	20-1500	6.58	0.40	42	8.95
TUF-5LH	10		6.9	0.27	42	10.95
TUF-5MH	13		7.0	0.25	41	11.95
TUF-5H	17		7.5	0.17	50	13.95
TUF-860	7	860-1050	6.2	0.37	35	8.95
TUF-860LH	10		6.3	0.27	35	10.95
TUF-860MH	13		6.8	0.32	35	11.95
TUF-860H	17		6.8	0.31	38	13.95
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ous values for transmission line widths and lengths, an impedance of 46 ohms (about 29 mils wide) turned out to be ideal. The length of the line resulted in about 16 degrees of electrical length (approx. 175 mils long) which is plenty of length to provide enough room for three 0603 SMT pads.

There is another subtle reason for having a slightly longer transmission line element here. Needless to say, neither CAD, resistors nor capacitors are ideal, nor is tuning the length of a short series element easy. As a result, the 5.6 pF capacitor was originally configured to slide up and down this short length of line in a trombone fashion allowing for some tunability.

To see the effects of the 46 ohm transmission line one must re-normalize the Smith Chart's impedance to 46 Ω then move 16 X 2 degrees clockwise. To find the resulting impedance in a 50 Ω system one must then re-normalize back to 50 Ω.

ii) After the 5.6 pF capacitor, a 46 ohm transmission line is added which is electrically 16.26 degrees long.

$$Z_2 = Z_0 \frac{Z_1 + jZ_0 \tan\theta}{Z_0 + jZ_1 \tan\theta} \quad (3)$$

$$= 40.413 - j15.626\ \Omega$$

$$Z_{2-46} = \frac{Z_2}{Z_0} = 0.877 - j0.339\ \Omega$$

iii) Mapping this impedance in 46 ohms.

$$\Gamma_{2-46} = \frac{Z_2 - 46}{Z_2 + 46} = 0.189\angle -99.42^\circ \quad (4)$$

The next element is yet another series element to achieve our final noise match impedance. Conceptually, an inductor was originally chosen to achieve the final value of Z_3. But inductors have drawbacks, so a high impedance line was chosen instead. The limitation of the line's impedance was the minimum line width of 10 mils which we felt could be easily achieved in production. As a result, this works out to be about 80 ohms. The same procedure as used previously resulted in a length of 60 electrical degrees.

iv) The following element is a 80 ohm transmission line which is electrically 60.05 degrees long. Z_2 will be transformed through this line to Z_3.

$$Z_3 = Z_0 \frac{Z_2 + jZ_0 \tan\theta}{Z_0 + jZ_2 \tan\theta} \quad (5)$$

$$= 63.01 + j50.015\ \Omega$$

$$Z_{3-80} = \frac{Z_3}{Z_0} = 0.791 - j0.628\ \Omega$$

vi) To locate the impedances Z2 and Z3 on the Smith Chart from the previous derivations, one must find the corresponding reflection coefficients in an 80 Ω system.

$$\Gamma_{2-80} = \frac{Z_2 - 80}{Z_2 + 80} = 0.35\angle -151.06^\circ \quad (6)$$

$$\Gamma_{3-80} = \frac{Z_3 - 80}{Z_2 + 80} = 0.35\angle 89.486^\circ$$

Note that the differences in the angle of G2-80 and G3-80 is twice the electrical length of the 60 degree transmission line. Using the Smith Chart, Z_3 can be easily located by a 120 degree clockwise rotation from Z_2 using a constant radius of 0.189.

To finally see where Z_3 winds up in 50 ohms, we must re-map to a 50 ohm system. This final result should equal Γ_{opt}.

vi) Re-mapping Z_3 in 50 ohms.

$$Z_{3-50} = \frac{Z_3}{50} = 1.26 + j1.0\ \Omega \quad (7)$$

$$\Gamma_{3-50} = \Gamma_{opt} = \frac{Z_3 - 50}{Z_3 + 50}$$

$$= 0.418\angle 51.546^\circ$$

As you can see there is a difference between our mathematical Γ_{opt} and the CAD value/data sheet value of 0.49 ∠ 49.2°. This difference comes from the variations between an ideal theoretical transmission line model verses a non ideal model which we choose using CAD. Since the math derivation was done after the design for illustration, one also gets an idea how much error can be induced from assuming ideal conditions.

Had we chosen an inductor, it would had a value of around 6 to 7 nH. From this value the parasitics must be subtracted to achieve the final value of a component itself such as a SMT inductor. Considering an estimated 2 nH per pad would result in a part of 3 nH. Though this part is available, its tolerance, ±10% plus the changes in parasitics perhaps ±1 nH for solder volume and part placement variations per pad would most likely result in an unacceptable tolerance for this matching element.

Printed inductors or transmission lines are free as compared to SMT inductors which typically cost 10 to 25 times as much as resistors or capacitors in volume. Printing an inductor is easy and results in highly repeatable results, however it may not be practical. Printed inductors usually exhibit poor Q due to the lossy dielectric, and, if a ground plane exists, they are no more than a high impedance transmission

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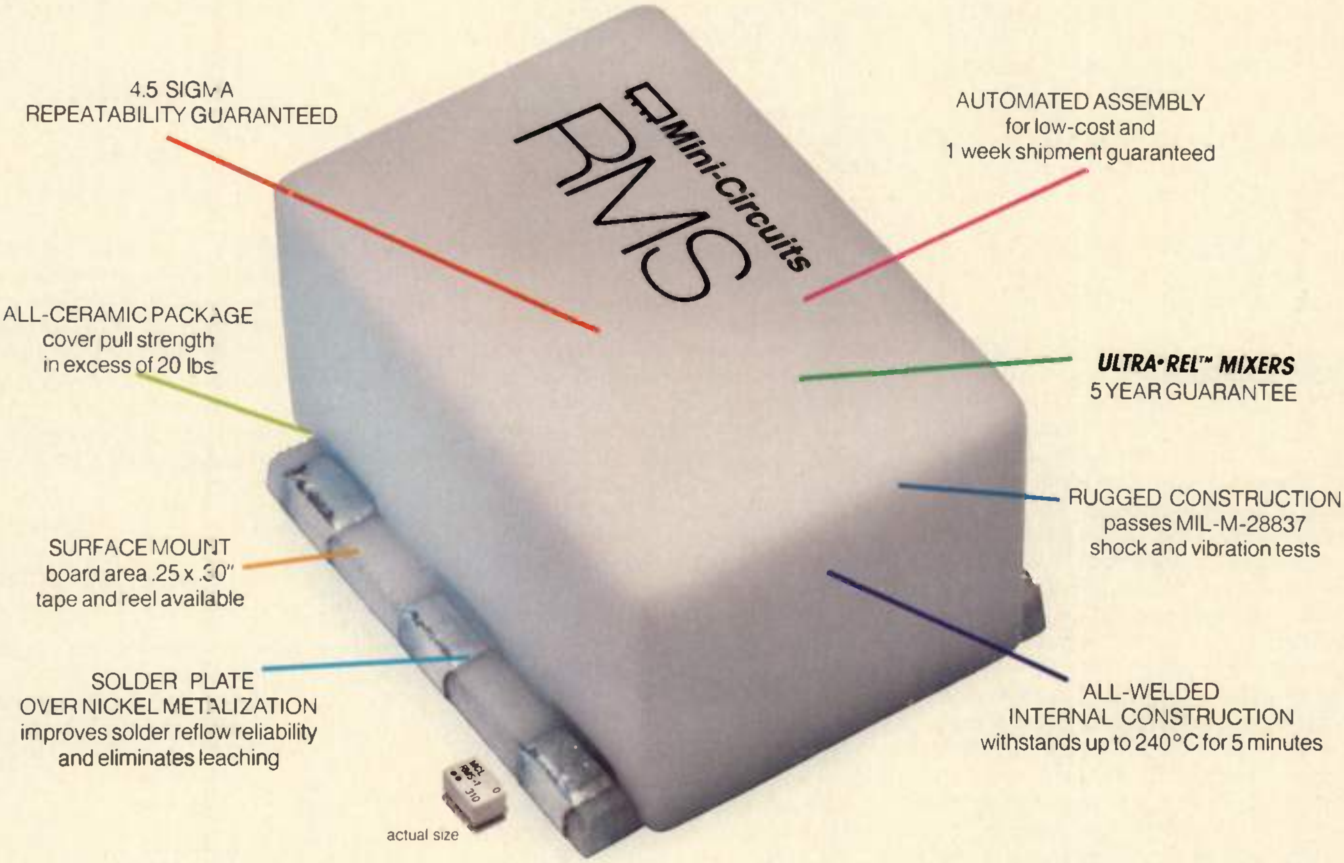

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line. As shown, a high impedance transmission line can replace an inductor to some degree, but inductors and high impedance transmission lines have a different trajectory on the Smith Chart as shown in Figure 4. High impedance transmission lines can be made to look more like printed inductors in cases where the backside of the PCB is suspended away from a grounded chassis. This is accomplished by removing the backside ground plane of the PCB directly under the printed inductor. In our case, even though the ground plane could be removed around this element on a multi-laminate board, we could not afford the risk of digital noise coupling into the input of the LNA from circuitry on the opposite side.

Had we used an inductor instead, the final result of Z_3 would be:

$$X_L = j\mathrm{Im}(Z_3 - Z_2) = j65.641\ \Omega \tag{8}$$

$$L = \frac{X_L}{2\pi f} = 6.637\ \mathrm{nH}$$

$$Z_{3L} = X_L + Z_2 = 40.413 + j50.015\ \Omega$$

$$\Gamma_{3L-50} = \frac{Z_{3L} - 50}{Z_{3L} + 50} = 0.493\angle 71.9^\circ$$

Output Structure

The first problem in this circuit is the low bias of 3V. Due to the 3 mA current, we must use a resistor that is less than 200 Ω. Unless the shunt resistor can be placed in a low impedance point of the network, undesirable loss in gain and RF power as well as degraded NF will result. The obvious element to use is either an inductor or transmission line placed in shunt with the collector which can be appropriately terminated for RF yet still allow for collector bias. This element also has the added advantage of being a high pass element to improve gain flatness which was sacrificed at the input.

The next concern is what load impedance to match. Remember matching to the conjugate of S_{22} is only valid if the input is conjugately matched. Since S_{12} is non-zero, whatever load is presented to the input will cause the output load to change. To predict this change, one can use equation 9 and the following S-parameters, interpolated from the data sheet:

S_{11}=0.288 ∠ −133.24°,
S_{21}=3.305 ∠68.19°,
S_{12} = 0.137 ∠ 35.05°,
S_{22} = 0.526 ∠−62.475°, along with
Γ_{opt}= 0.49 ∠ 49.2°

$$S_{22new} = S_{22} + \frac{S_{12}S_{21}\Gamma_0}{1 - S_{11}\Gamma_0} \tag{9}$$

$$= 0.342\angle -79.53^\circ$$

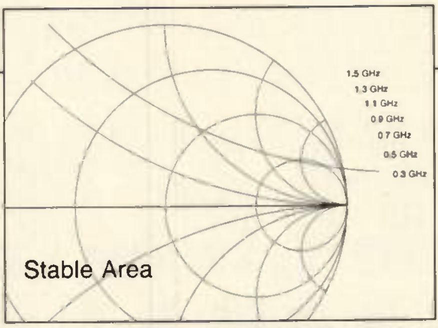

Figure 5. Output stability circles with input match to Γ_{opt}

To match the output of the transistor, we took the conjugate of this impedance. The following is a comparison of the original S_{22}, S_{22} prime, and EEsof's optimized result as well as the conjugate of the previously calculated Γ_4 for an ideal transmission line.

S_{22} of device:	0.526 ∠ −62.475
New S_{22} caused by S_{12}:	0.342 ∠ −79.563
Optimized result::	0.364 ∠ −81.534
Γ_4*:	0.386 ∠ −77.34

The next concern is stability, especially if a filter is going to be used at the input. In our case, the output port can potentially give difficulties since the input is very restricted by its match. Should we enter a conditionally stable region at the input port at any frequency, the output port will see a reflection coefficient greater than 1. Likewise we must consider the conditionally stable region for the output port which may cause the input port to yield negative resistance.

As shown in our stability analysis, we have a little more freedom for the output port. Our only concern is that there will be areas of negative resistance presented at the output port. Whether this poses a problem or not will depend on how the output load is terminated.

Our first element was again a 5.6 pF capacitor. On the output section, bias is not supplied by any SMT components, so no intermediate transformation was used.

i) Starting out with a 50 ohm load and a 5.6 pF capacitor in a 80 ohm system.

Z_0=79.62 Ω, f=1574 MHz, R=50 Ω

$$Z_1 = \frac{1}{j2\pi \bullet 6.5\mathrm{x}10^{-12}} + R \tag{10}$$

$$= 50 - j18.056\ \Omega$$

$$Z_{80} = \frac{Z_1}{Z_0} = 0.628 + j0.227\ \Omega$$

$$\Gamma_{1-80} = \frac{Z_1 - Z_0}{Z_1 + Z_0} = 0.265\angle 39.297^\circ$$

Since DC current is present on the output matching section, a series element was used to match Z_1 to Z_2. This element consists of a high impedance transmission line that will transform Z1 to a point where a shunt transmission line at Z_2 can achieve the final match at Z_4.

ii) After the 5.6 pF capacitor, a 79.62 ohm transmission line is added which is electrically 55.54 degrees long.

$$Z_2 = Z_0 \frac{Z_1 + jZ_0 \tan\theta}{Z_0 + jZ_1 \tan\theta} \tag{11}$$

$$= 59.892 + j32.439\ \Omega$$

$$Z_{2-46} = \frac{Z_2}{Z_0} = 0.752 + j0.407\ \Omega$$

iii) Re-mapping this impedance in 50 ohms.

$$Z_{2-50} = \frac{Z_2}{50} = 1.198 + j0.649\ \Omega \tag{12}$$

$$\Gamma_{2-50} = \frac{Z_2 - 50}{Z_2 + 50} = 0.296\angle 56.595^\circ$$

iv) Since shunt elements are best described as admittances instead of impedances on a Smith Chart we will consider Y_3 instead of Z_3. The next element, Y_3, is a shunt, short circuit, 80 ohm transmission line that is 70.54 electrical degrees long. This admittance will eventually be in parallel with Z2 resulting in a load, Z_4, to the collector of the transistor.

$$Y_3 = (Z_0 j \tan\theta)^{-1} = -j4.438\ \mathrm{mmohs} \tag{13}$$

$$Y_{3-80} = \frac{Y_0}{Y_3} = Y_3 Z_0 = -j0.353\ \mathrm{mmohs}$$

iv) Next we take shunt element Y_3 and connect it in parallel with
Z_2 to find the load the collector will see.

$$Z_4 = \frac{1}{\frac{1}{Z_2} + Y_3} \tag{14}$$

$$= 43.423 + j38.445\ \Omega$$

v) And finally re-mapping Z_4 to 50 ohms.

$$Z_{4-50} = \frac{Z_4}{50} = 0.868 + j0.769\ \Omega \tag{15}$$

$$\Gamma_{4-50} = \frac{Z_4 - 50}{Z_4 + 50} = 0.386\angle 77.34^\circ$$

The one major drawback with using shunt elements is that an additional DC block series capacitor is required. This drawback is associated with stray resonances caused by either the capacitor itself resonating with another elements in the circuit or bias network, or insufficient

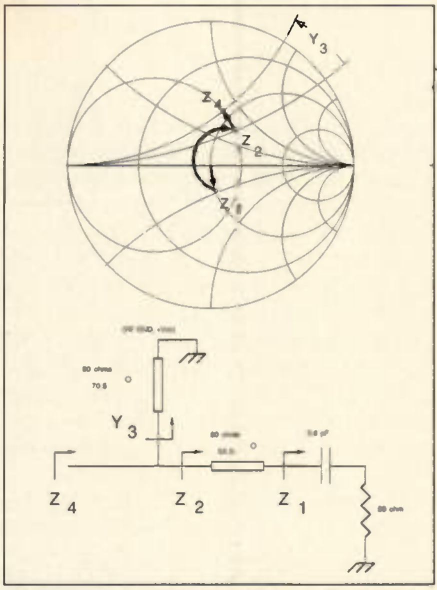

Figure 6. Final mapping of Z_1 to Z_4 normalized to 50 ohms.

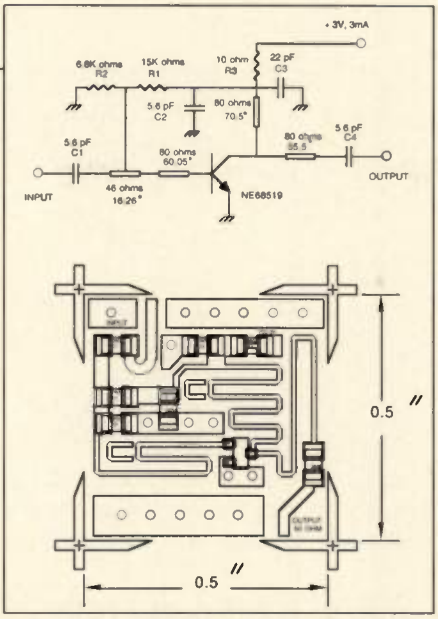

Figure 7. First iteration of LNA circuit.

collector isolation from the supply line at low frequencies. In some cases, designers use two capacitors in parallel, one of a small value, a few pF or so, the other value much larger, around 0.1 µF. While this often works, in some cases, as here, the larger capacitor has sufficient inductance to parallel resonate a 5.6 pF capacitor causing the shorted stub termination to become an open circuit at 1575 MHz. This resulted in a 10 dB notch in the passband of the LNA.

This was resolved by replacing the parallel combination for a pi network and using a different value capacitor from before. The 5.5 pF was substituted by a 22 pF and the larger capacitor was replaced with a 1200 pF. Both capacitors were separated by a 10 Ω resistor which de-Q'ed any resonances which may occur in between the capacitors. The simple pi network worked well and no bias problems were noted after that.

Measured Results

The measured performance of the LNA was very close to what was predicted. The LNA's noise performance could only be improved by about 0.1 dB from our first iteration.

To confirm that the base of the transistor was seeing our intended noise match, the input network was tested by removing the device and connecting a micro coax at the NE68519's base pad. The coax and its connector were carefully de-embedded by a simple port extension on the HP8510. The measurement showed that the network had closely made the intended impedance transformation from 50 ohms to Γ_{opt}.

The total measured noise figure of the LNA, including connectors, is about 1.9 dB, which is not unrealistic considering that the minimum noise figure of the device itself is around 1.5 dB. Had we used higher quality capacitors, some improvement may have been possible. This was of little concern since the LNA was eventually going to be connected to a DC isolated ceramic filter which will absorb the capacitor.

The next measurement is output return loss. Here we can look for ways to improve the high end roll off and to check for potential instabilities. First it was noted that S_{22} was around 11 dB at 1575 MHz, which in theory is a 0.278 dB mismatch loss.

$$\text{Return Loss, RL} = 11\text{ dB}$$

$$\rho = 10^{\frac{-RL}{20}} = 0.278 \tag{16}$$

$$\text{MismatchLoss}_{dB} = -10\log(1-\rho^2) = 0.348\text{ dB}$$

By extending the series element about 50 mils in between the output and the collector, the match was improved to 20

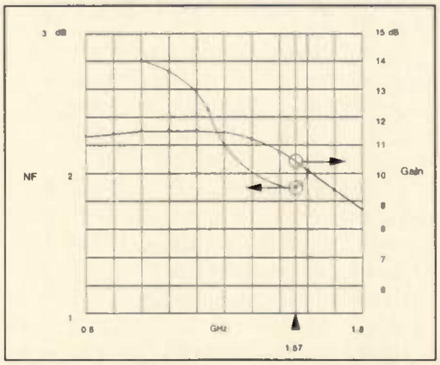

Figure 8. Gain and noise figure performance of LNA.

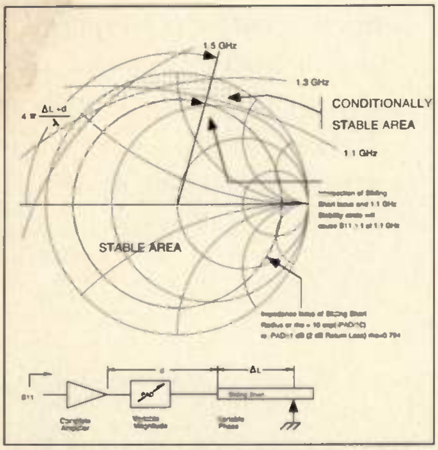

Figure 9. Measuring stability using a sliding short.

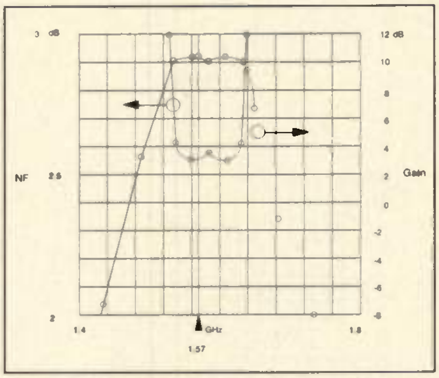

Figure 10. Gain and noise figure response of LNA with ceramic filter.

dB. This extension effected the frequency response between 1.4 to 2 GHz resulting in a high end flatness improvement of 0.25 dB, close to our original estimate of 0.283 dB.

Stability is our next concern. By placing a sliding short at the input of the LNA one can easily see the regions of load and frequency where S_{22} becomes greater than 1. As estimated by our Touchstone model, a potential problem area could lie around 0.8 to 1.2 GHz. This could be a problem with a ceramic filter so our second iteration included a lossy resonator to correct for this potential problem.

Through our sliding short measurements, we found that a load with a return loss greater than 2 dB at the input would not cause S_{22} to become greater than 1. This was accomplished by inserting a 1 dB pad in between the sliding short and the LNA and watching what happens to the magnitude of S_{22}. With out a pad, a return loss of near 0 dB will result in S_{22} having a return gain of around 2 dB. With the pad in place, the return loss drops slightly below 0 dB. We can estimate that a return loss of around 1 to 1.5 dB at some phase angles will cause S_{22} to become greater than 1 around 0.8 to 1.2 GHz, agreeing with our CAD results.

Our next concern was the linearity of the amplifier. Our measured P_{1db} was 2.45 dBm which closely matched our following estimated result. We estimated that the P_{1db} and input intercept point from the following rules of thumb:

Estimated P1dB:
V_{CE}=2.5V, I_C=3mA, Typ. Efficiency=25%

$$P_{1dB} = 10\log\frac{\eta V_{CE} I_C}{1000} \qquad (17)$$
$$= 2.73 \text{ dBm}$$

Estimated input intercept point (IIP3):
P_{1dB}=2.73 dBm Gain=10.5 dB

$$IIP_3 = P_{1dB} + 10 - \text{Gain} \qquad (18)$$
$$= 2.73 \text{ dBm}$$

Second Iteration

Our next iteration involved the incorporation of a ceramic filter and the ability to incorporate a series output LCR resonator for stability. The results of these modifications without the LCR resonator are shown in Figure 11.

The first step was to try to remove the input capacitor, C1, and absorb this capacitance by the parasitic capacitance of the resonator. Even though these capacitances are not equal, the degradation of NF due to the match vs. the loss sustained by the capacitor will make this a worthwhile compromise, even without retuning the input. The filter (Toko: 4DFA-1575B-12) in this case is expected to have a typical loss of 0.7 dB and a maximum of 1.2 dB which will directly add to the noise figure.

As mentioned before, there was a concern of S_{22} becoming greater than one below 1575 MHz with the input ceramic filter. As one can see from the measured data, this is in fact true around 800 MHz and 1.4 GHz. Even though the reflection coefficient is greater than 1 at these frequencies, the device was stable in 50 ohms. Due to the application of this LNA, our biggest concern was at 800 MHz since the following stage will most likely have a reasonable match at 1.4 GHz but an unknown match at 800 MHz.

The simplest way to solve this problem is with a resistive loss at the output which will negate the effects of –R caused by the active device. Resistive loss will also reduce gain at 1575 MHz, which is undesired. The obvious approach is to apply a simple lossy filter which will only add loss at 800 MHz. The problem with this approach is the Q of the resonator. Adding too much loss will deteriorate the Q to the point were the pass band suffers. Insufficient loss will not remove the effects of negative resistance. This simple approach worked well at 800 MHz, but would be difficult if not impossible at 1.4 GHz since it is hard to get enough loss at high enough selectivity to avoid severely degrading the passband performance.

This selective resistor approach can be implemented in two ways -either in shunt or series. In our case, for simplicity and minimum loss to the desired passband, we chose a shunt LCR topology at the output. A unique implementation of this solution is a simple chip resistor that is only connected on one end. An unconnected chip resistor to a UHF amplifier is usually just that; an unconnected resistor, but at X band it makes a very nice lossy shunt resonant transmission line.

The inductor was replaced with a variable transmission line. The L/C ratio is determine from the desired selectivity of the resonator by trying different values on CAD. The entire process can be initially tried on CAD, but the final adjustment, L & R, must be done empirically. The resonance is adjusted by changing the transmission line length while measuring the dip in S_{22}. The resistance is chosen by compromising the loss at 1575 MHz verses the required resistance to make S_{22} equal to zero at 800 MHz.

Our final values were R=20 Ω, C=1 pF and a 10 mil wide transmission line around 800 mils long. The goal was to keep S_{22} below 0 dB at 800 MHz as well as improve S_{22} at 1.4 GHz. The cost to our performance was 0.5 dB loss in gain at 1575 MHz. Though this is not very much loss in general, 0.5 dB was not insignificant at 10 dB gain. As it turned out, the device was stable without our stability resonator in 50 ohms as well as in our desired load which had a S11 as high as +3 dB at some frequencies.

Summary

We have discussed in detail one example of how to design a simple narrow band LNA as well as some of the concerns one must have during the design process. Effort was made to make the design approach practical by using common CAD design tools in balance with standard bench techniques as well as considering real production concerns. Each circuit element was analyzed using

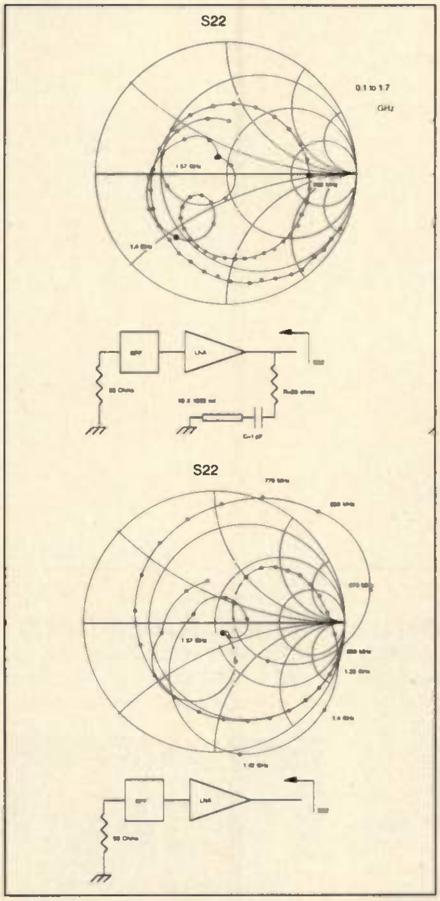

Figure 11. S_{22} with and without stability network.

basic transmission line theory. The values for these models were derived from the final optimized CAD results in order to compare how the non-ideal models from CAD differ from the ideal transmission line elements.

In practice, 10 dB of gain is not sufficient to mask the high noise figure of today's low cost MMIC frequency converters. A typical GPS system NF is around 3 dB or less, requiring the LNA to have a typical gain of about 20 dB at the achieved NF or better. The LNA presented is easily cascaded to achieve this gain. Even though the NE685 is a silicon bipolar transistor, it is an outstanding device at L band under very low bias levels. Considering cost, stability, biasing or matching impedance at UHF, bipolars are typically much easier to use than GaAs FETs. Though GaAs FETs can have a significant performance advantage over silicon, the NE685 is still a viable solution for GPS as well as many other applications through S band. *RF*

About the Author

Terry Cummings has a BSEE from San Jose State University in California. He currently is an applications engineer at California Eastern Laboratories and has over 10 years experience in design engineering, production, and product marketing. He can be reached at CEL, 4590 Patrick Henry Dr., Santa Clara, CA 95056-0964, or by phone at (408) 988-3500.

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Paragon Technology, Inc.
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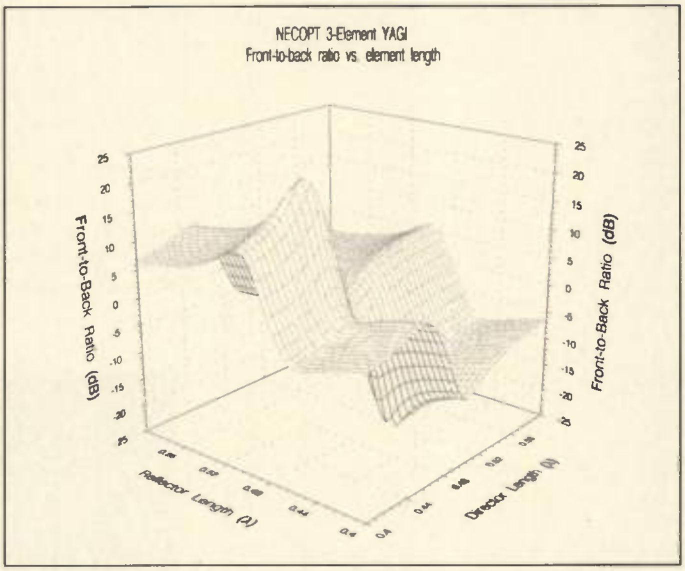


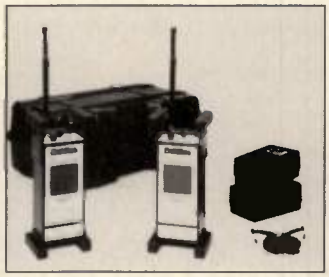
PHEMT LNA MMIC

Hewlett-Packard is releasing what it believes is the first commercially available GaAs psuedomorphic high-electron mobility transistor (PHEMT) low-noise amplifier integrated circuit. The MGA-86576 extends the availability of a simple, self-biased gain stage to applications through 8 GHz. It provides 23 dB gain and 1.6 dB noise figure at 4 GHz, while operating from a single 5 V power source, requiring only 16 mA. The MGA-86576 requires only an RF choke, output blocking capacitor, and a series input inductor to replace a typical discrete design. Applications include GPS receivers, PCS, 2.4 - 5.7 GHz spread-spectrum, and satellite systems operating at 1.5 to 4 GHz. The device is in stock and is priced at $8.00 each in quantities between 1 and 499.

Hewlett-Packard Co.
INFO/CARD #206

Path Loss Measurement

Chesapeake Microwave Technologies has introduced the PathTraxx TM system, designed for measuring the relative signal strength and path loss between elements within wireless communications networks. The PathTraxx system incorporates a self-calibrating, microprocessor-based architecture that makes the system easy to use. It provides coverage in the 864 to 936 MHz band, with other frequency bands available on request. The system can be tuned in 100 kHz steps to avoid interferers. The transmitter and receiver have back-lit LCD displays showing path loss measurements, battery status and other operating parameters. An RS-232 port also allows connection to a computer for data capture. Both transmitter and receiver measure 12 x 4 x 5 inches and weigh 5.5 lbs. The system includes antennas, Nicad battery packs and chargers, and a durable carrying case.



Chesapeake Microwave Technologies, Inc.
INFO/CARD #207

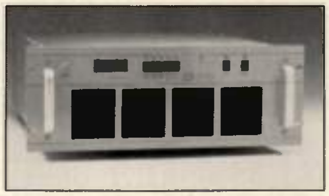
Microstrip Mixers

Magnum Microwave is offering a series of communications-band mixers built on "drop-in" microstrip carriers. The MC33, 34, 36 and 37MS-1 and the MC33, 34, 36 and 37MS-3, are targeted at upconverter applications for the 5.8 to 6.5 GHz band. The IF frequency range covered by these units is DC to 1.5 GHz, with typical conversion loss of 5 dB. LO drive required is +7, +10, +13 or +17 dBm nominal, depending on the model selected, with corresponding third order intercepts of +11, +14, +18 and +21 dBm respectively. The -1 units measure 0.8 x 0.59 x 0.15 inches, while the -3 units are 0.6 x 0.32 x 0.115 inches. Similar mixers are available for downconversion applications (3.6 to 4.3 GHz) and for other satcom bands.

Magnum Microwave Corp.
INFO/CARD #208

Power Amplifiers

The B-Series amplifier chassis provides a perfect environment for Silicon Valley Power Amplifiers' modules. The seven inch-high by 19 inch rack mounted chassis provides power and control to single and combined broadband amplifier modules which cover 5 MHz to 500 MHz. Linear, class AB power outputs from 4 W to 1000 W can be supplied by single modules or combinations of modules. By use of a directional coupler and AGC circuitry, leveling to ±0.25 dB can be maintained over the rated frequency range. Additionally, protection is provided by gently reducing output load as VSWR increases. All combinations utilize Lambda switched-mode power supplies, giving wide AC supply voltage and frequency options. The amplifier units are designed for conduction cooling, but can be supplied with an integral convection-cooled heat sink. Remote control is standard, with all metered functions available as a separate analog line together with ON/STANDBY control and indication. RS-232 and IEEE interfaces are available as an option.



Silicon Valley Power Amplifiers
INFO/CARD #209

Product Spotlight: Filters

Switchable Filters

KW Microwave introduces a new series of integrated switchable filters covering frequencies from 100 MHz to 18 GHz. The switches are either PIN diode or GaAs FET switches. The filters are lumped, combline and suspended stripline designs. The switch filters are available in plug-in, drop-in and connectorized versions.
KW Microwave Corp.
INFO/CARD #210

Switched-Capacitor Filters

Maxim Integrated Products announces the MAX291/MAX295 (Butterworth) and MAX292/MAX296 (Bessel) 8th-order, lowpass, switched-capacitor filters. The filter's corner frequency is set by the frequency of a clock signal, with clock to corner frequency ratio being 100:1 for the MAX291/292 and 50:1 for the MAX295/296. The MAX291/292 has a 0.1Hz to 25 kHz corner frequency range, while the MAX295/296 has a 0.1Hz to 50 kHz range. Prices start at $2.95.
Maxim Integrated Products
INFO/CARD #211

70 MHz LC Filter

Piezo Technology has expanded its standard 70.0 MHz LC filter product line. Model 7997 offers exceptional passband flatness with a 1 dB bandwidth of ±20.5 MHz. The stopband extends out to 2.0 GHz with 70 dB of attenuation. The unit is housed in a 1.5 x 0.4 x 0.4 inch package with 50 ohm input and output impedance and a 1.5:1 VSWR.
Piezo Technology, Inc.
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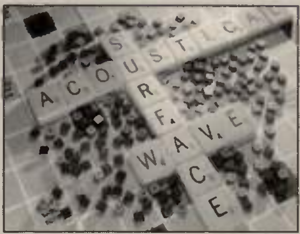
SAW Bandpass Filters

OKI Semiconductor has announced the availablility of surface acoustic wave (SAW) filters in the 750 MHz to 2 GHz range. Packaged in 12-pin, sur-



face mount packages, the filters' footprint is 4.8 x 4.8 mm, with a thickness ≤ 2 mm. Insertion loss is less than 4.5 dB at most frequencies. Production quantities are currently available. Unit prices range from $3.94 to $5.44 in 1000-piece lots.
OKI Semiconductor
INFO/CARD #213

AMPLIFIERS

Base Station PA

M/A-COM recently announced a new 2 W, class A power amplifier for PCN and spread spectrum base station transmitters. This 1700 to 2200 MHz amplifier provides output power of 33 dBm (1 dB compression point) and 34 dBm (psat) with a gain of 32-38 dB. IMD is –38 dBc. The amplifier, part number CPA-131-PAA, covers all established PCN channels in a single unit.
M/A-COM, Inc.
INFO/CARD #214

30 kW Amplifier

Microwave Power Devices has developed a 30,000 W continuous wave, solid-state power amplifier, module number 1050-34/14123. This amplifier will be used for susceptibility test and analysis. The amplifier operates class AB linear anywhere in the 100 to 500 MHz band. It supports CW/FM/AM/phase and pulse modulation formats. Harmonics are 35 dBc, control and status is via IEEE-488 GPIB.
Microwave Power Devices, Inc.
INFO/CARD #215

L-Band PA

Power Systems Technology announces the design and availability of its L-band solid state power amplifier model CHCD 178198-500/3350 that is suitable for use in a satellite ground-based telemetry tracking and control system. The amplifier operates class C, has an RF power output of 500 W and covers 1750 to 1850 MHz. The amplifier operates with an RF input of –5 dBm full power output and can be controlled and monitored remotely via

an RS232 interface.
Power Systems Technology
INFO/CARD #216

SIGNAL SOURCES

Fast Warm-Up OCXO

Frequency Electronics has developed a commercial OCXO, model FE101A. Measuring 1.27 x 1.33 x 1.33 inches, the FE101A is a rugged, compact, ultra-stable, subminiature crystal oscillator that warms up in less than two minutes to 1×10^{-7}. Other specifications include temperature stability of 5×10^{-8} (–50° C), and 1.75 W power consumption.
Frequency Electronics, Inc.
INFO/CARD #217

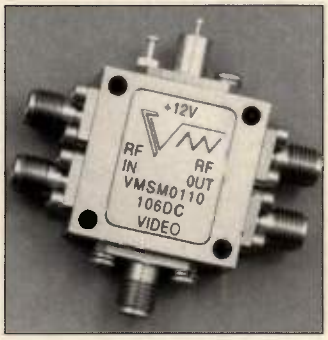
Microcomputer Controlled Oscillator

The MCXO (Microcomputer Controlled Crystal Oscillator) uses a high-stability, third overtone, SC-cut crystal and microcomputer-controlled compensations instead of an oven to acheive $\pm 3 \times 10^{-8}$ ppm stability. Maximum power consumption is 75 mW. Standard output frequencies are 10.000 MHz, 4.000 kHz and 1 pulse per second. The commercial version sells from $559 each in quantities of 500.
Q-Tech Corporation
INFO/CARD #218

SIGNAL PROCESSING COMPONENTS

Fast Switch

The VMSM 0110-106 ultra-broadband nonreflective "active" switch exhibits 6 dB insertion gain across 0.1 to 10 GHz, with rise/fall times of 100/150 ps and video to RF isolation of 18 dB min. This unit will find use as a high speed modulator, switch, or voltage variable attenuator. The VMSM 0110-106 measures 1.0 x 0.99 x 0.40 inches.
Veritech Microwave, Inc.
INFO/CARD #219



Attenuators

A family of power, base-mounted, conductively cooled attenuators, available in 20, 50 and 100 W models, has been introduced by Component General. The units are designed to be bolted directly to a ground plane. In addition, gold plated beryllium copper tabs are provided for circuit connection. Attenuation values from 1 to 20 dB are available. Power ratings are based on maximum base temperature of 85° C. The attenuators are available form stock.
Component General, Inc.
INFO/CARD #220

High Power Circulators and Isolators

Multi-junction 800 to 960 MHz circulators and isolators from UTE Microwave handle 500 W CW when properly heat sinked. Isolators with internal terminations absorb up to 150 W reverse power. The units are available for broadband operation or for optimized bands, where typical loss is 0.15 dB and isolation of 25 dB per junction can be achieved. Size is 5-7/8 x 2-1/4 x 1-1/4 inches for the triple junction unit.
UTE Microwave, Inc.
INFO/CARD #221

Cable Equalizer

Inmet's miniature cable equalizer compensates for cable loss in coaxial runs or delay lines. The units are custom-built for broadband applications. The typical negative slope response can be held within a 1 dB window for 3 dB to 30 dB attenuation values. Typical VSWR is 1.50:1. SMA, Type N, 2.9mm or solderable connectors can be selected.
Inmet Corp.
INFO/CARD #222

Circulator

McManus Microwave's model 106 circulator operates from 140 to 175 MHz and has maximum insertion loss of 0.8 dB. Minimum isolation is 18 dB and maximum VSWR is 1.30:1. The model 106 can handle 10 W CW and uses three SMA female connectors. The unit measures 1.0 inch

INFO/CARD 41

square by 0.4 inches high and weighs 1 oz.
McManus Microwave
INFO/CARD #223

SUBSYSTEMS

Patch Antenna

Antenna Research Associates is introducing a wideband, high gain microstrip patch antenna for applications from 3.0 to 3.5 GHz. The antenna is designed for 16% bandwidth with a 2.0:1 VSWR

and offers gain of 8.0 dB over the band. Typical half-power beamwidth for this antenna is 60° x 40°. The hermetically sealed assembly can be used as a feed for reflector antennas. The antenna is 3.5 inches in diameter, 0.3 inches high, and weighs less than 3 oz.
Antenna Research Associates
INFO/CARD #224

Base Station Modules

Frequency Products has introduced a series of modules, the CR-910 dual conversion receiver, CR-920 transmitter driver, and CR-930 frequency synthesizer, that perform major RF functions in a cellular base station. The CR-910 has –45 dBc IMD3 at –20 dBm, with IF output at 21.4 MHz at –10 dBm. The CR-920 transmitter driver has IMD3 of –45 dBc at –10 dBm. The CR-930 synthesizer can be tuned in 25 kHz steps from 700 to 1030 MHz, with spurious at –50 dBc.
Frequency Products, Inc.
INFO/CARD #225

Antenna Signal Processor

The Marconi CMA-2032 antenna signal processor (ASP) protects radio base stations from co-channel and adjacent channel interference. The ASP uses an independent input to sample interference. Internal signal processing then dynamically cancels the interference received by the base station antenna. Installation requires no modification to the existing base station antenna system or receiver. No switching paths are are introduced into the antenna system and the ASP will contribute less than 1 dB insertion loss.
Marconi Instruments
INFO/CARD #226

SEMI-CONDUCTORS

2.4 GHz PA MMIC

The TAE-1020 from Teledyne Electronic Technologies is a GaAs power amplifier MMIC operating in the 2.4 GHz ISM band. The amplifier operates from ±3 to ±5 V and has 20 dB small signal gain. Minimum linear output power is 24 dBm, and minimum saturated output power is 25.5 dBm. The TAE-1020 comes in a 16-lead surface-mount plastic package.
Teledyne Electronic Technologies
INFO/CARD #227

GaAsFET Bias IC

Maxim introduces the MAX850-MAX853, a new family that combines a charge pump voltage inverter and a low-noise linear regulator in a single space-saving, 8-pin surface mount package. Needing only miniature capacitors to generate a –4.1V output, they supply 5 mA with less than 2 mVpp ripple. The output is also adjustable from –0.5 to –9 V. Prices start at $1.65 for quantities of 1000 and up.
Maxim Integrated Products
INFO/CARD #228

Baseband Receive Port

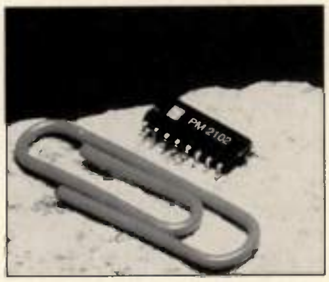
Analog Devices' AD7013 is a complete digital baseband receive port for digital cellular handsets or base-stations. It captures, digitizes, and filters I and Q signals, using two 15-bit sigma-delta ADCs. The device can be used for either TIA IS-54 or AMPS standards by selecting the appropriate digital filter response. The AD7013 also includes three DACs (two 8-bit and one 10-bit) to perform RF control functions such as AGC and frequency control. The device costs $6.30 in 1000s.
Analog Devices, Inc.
INFO/CARD #229

Gilbert Cell Transistor Array

A 5 GHz (power gain BW product) silicon Gilbert cell transistor array IC for radio frequency mixer and amplifier applications up to 2.5 GHz in wireless communications systems is now available from Harris Semiconductor for $2.66 (in quantities of 1000). The IC, called the HFA3101 is a direct pin replacement for the industry's only Si Gilbert cell transistor array, NEC's UPA101.
Harris Semiconductor
INFO/CARD #230

1900 MHz PA MMIC

Developed for 1900 MHz DECT handheld applications, the PM2102, 16-pin plastic SOIC



GaAs MMIC provides +27.5 dBm P_{out} at 3.0 V bias with 50% efficiency. The Pacific Monolithics technology utilized in this device was recently granted a U.S. Patent. Production quantities are available now, as is a demo board and complete application note (App Note 2294A). Pricing is under $4.00.
Pacific Monolithics, Inc.
INFO/CARD #231

LNA MMIC

DAICO Industries introduces a GaAs MMIC low noise amplifier designated model #P35-4170. Operating frequency is 2 to 3 GHz with DC power consumption of +9 mA from a +3 V supply. They provide a low 2.9 dB noise figure and high 17 dB power gain. They require no external components, only a 3 V power supply. Input return loss is 10 dB, with output return loss of 6 dB. The part is configured in a small plastic package.
DAICO Industries, Inc.
INFO/CARD #232

DC Restoration

Burr-Brown's SHC615 is a unique DC restoration device designed to stabilize the performance of RF signals, but is also useful for other applications such as a fast phase comparator, a synchron demodulator in PLL circuits, or an RF pulse code mod/demod. Its key specifications include: 280 MHz bandwidth, 2.2 ns propagation delay, 1700 V/MUs slew rate, 4 nV/√HZ noise, and 600 mW maximum power dissipation. The SHC615 is available in a single 14-pin plastic DIP or SOIC package and in die form. It is priced from $7.95 in 100s.
Burr-Brown Corp.
INFO/CARD #233

TEST EQUIPMENT

Multipath Channel Simulator

The SOFI 03 AR is a wideband multipath mobile radio channel simulator for use in the bands of the GSM mobile radio system. Its frequency range is 890 to 915 MHz and 935 to 960 MHz, with maximum VSWR of 2.0. Several GSM profiles are already programmed into the controller software, with CDMA profiles currently in development. The three path version sells for $37,000.
Sofimation OY
AxTrade Inc.
INFO/CARD #234

Communications Service Monitor

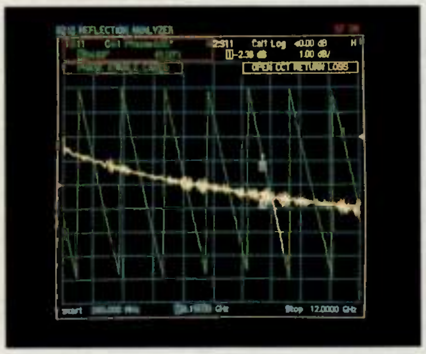
Tektronix has introduced the Rhode & Schwarz CMS 54, a comprehensive communications service monitor for mobile radio transceivers and base station equipment operating from 0.4 to 1000 MHz. The CMS 54 combines a spectrum analyzer, adjacent-channel power meter, harmonics meter, and the ability to measure frequency and power transients in a single unit. The CMS 54 is available to order now for delivery in eight weeks and is priced at $13,9000.
Rohde & Schwarz
Tektronix
INFO/CARD #235

Cellular Test System

Wavetek has announced the release of a digital/analog cellular test system, the model 3600D. The 3600D's programming flexibility allows as little or as much auto-

Marconi's 6200 Family of Microwave Test Sets have been making microwave scalar measurements to 46 GHz and providing high-speed, high-resolution fault location for coax lines & waveguide runs.

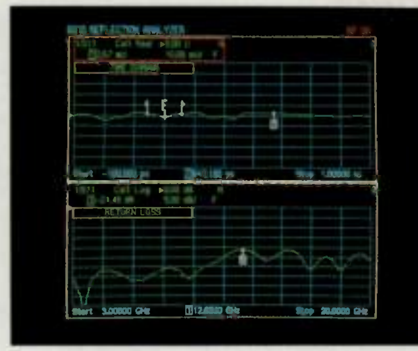
Now a powerful & economical new addition to the family, the **6210 Reflection Analyzer** is here. It uses the highly accurate "6-port coupler" technique to measure phase and amplitude characteristics of network inputs (S_{11}). The Test Set thus provides accurate return-loss measurements, vector measurements, and time domain measurements. The frequency range of the Model 6210 starts at 250 MHz and extends to 26.5 GHz (or as limited by the host 6200 Series Microwave Test Set).



The Reflection Analyzer is housed in an add-on adaptor that fits below the 6200 Series Microwave Test Set thereby retaining its compact profile for portable and field use. This adaptor technique also provides an easy upgrade route for users of the 6200 Test Set now and at any time in the future. All existing features of the 6200 Series Test Set are retained.

Key features of the 6210 include:

- Higher accuracy and wider range reflection measurements as compared to the RF bridge technique.
- Both vector and time domain analysis so that the causes of reflections can be diagnosed. Especially useful with fault location.
- Smith Chart presentation for easier impedance matching adjustments.



- Simultaneous time domain and frequency domain measurements for full characterization of an input port.

For more information or to arrange a demonstration contact:

Marconi Instruments, Inc.
3 Pearl Court
Allendale, NJ 07401
1-800-233-2955
201-934-9050

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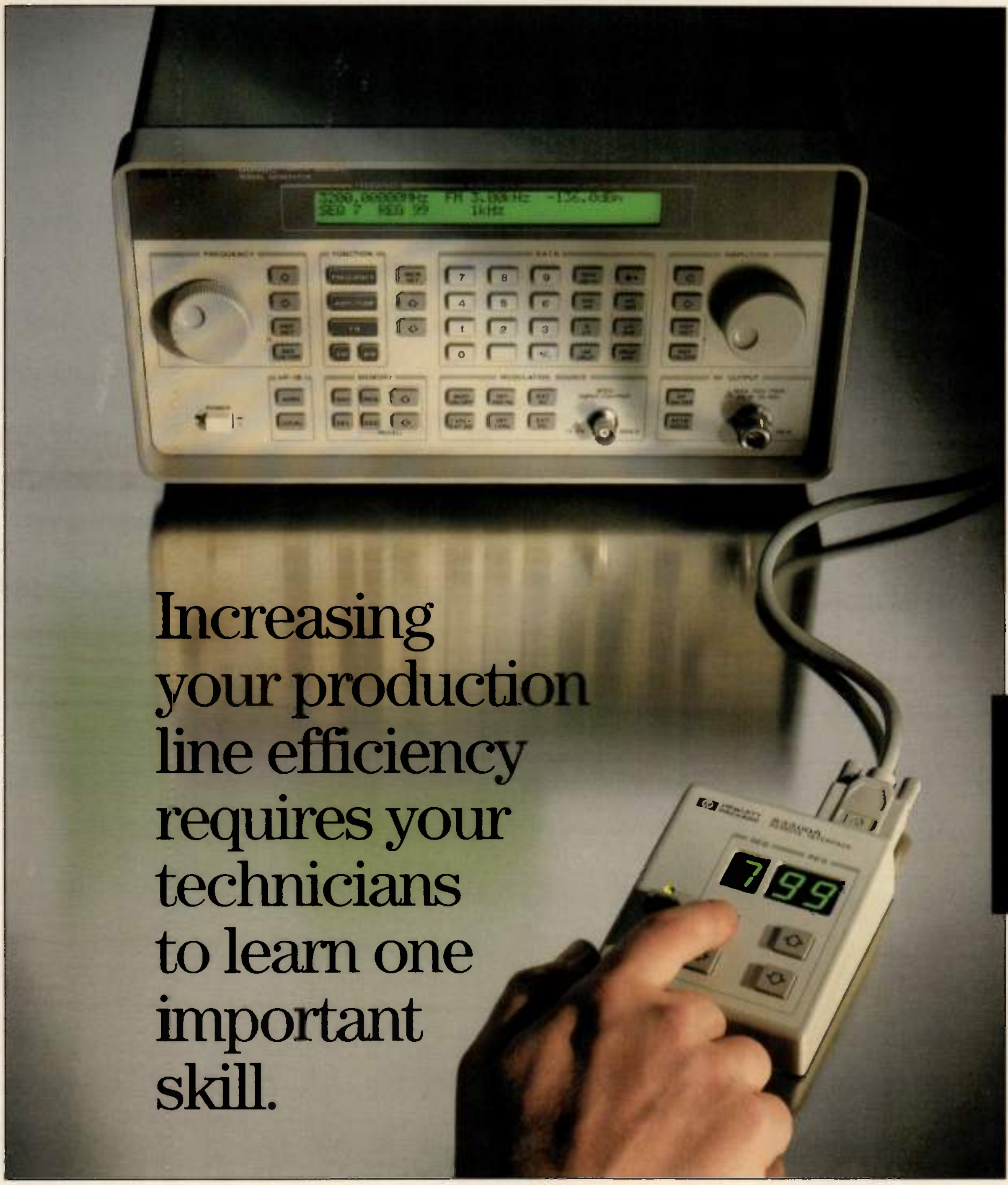
mated testing as desired. Nonstandard modulation tests can be implemented using the two internal 10 Hz to 100 kHz generators. A parallel printer port, RS-232 serial port and 1.44 MB PC-compatible disk drive allow for data transfer and instrument control. An IEEE-488 interface is optional.
Wavetek Communications Div.
INFO/CARD #236

Arbitrary Function Generator

The LeCroy WaveStation 400 series has 100 ps waveform feature placement, a fast 400 MHz maximum clock rate per channel, and up to 1 MB of waveform playback memory. In both the single channel LW410 and dual channel LW420 combine state-of-the-art DSP technology and precision reconstruction filtering techniques. A smoothing function minimizes discontinuities caused by editing. The LW410 costs $13,945, while the LW 420 costs $18,950.
LeCroy
INFO/CARD #237

Improved Power Meter

Boonton Electronics has enhanced its 4400 peak power meter by incorporating new software that allows measurement of long term average power between markers, peak power hold between markers, and peak to average power ratio between markers. Other benefits include disk support for storing and recalling up to 99 instrument setups on a DOS compatible disk. The 4400 performs power domain analysis from 30 MHz to 40 GHz.
Boonton Electronics Corp.
INFO/CARD #238



New! HP 8648C provides 3.2 GHz for less than $9,500.

The HP 8647A Signal Generator: a new level of productivity at the touch of a button.

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Yet the HP 8647A brings more performance to the production line. The 250 kHz–1 GHz range and clean +10 dBm output lets you tackle more applications. Separate manual frequency and amplitude control knobs make dialing in signals as easy as tuning a car radio. And the reliability of an electronic attenuator virtually eliminates the most common cause of failures.

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Application Circuits for MMIC Amplifiers

By Gary A. Breed
Editor

Here is small collection of circuits using MMIC amplifiers. In addition to amplification, these simple "building blocks" can be also be used to make simple, inexpensive oscillators, mixers and multipliers.

An amplifier gain block impedance-matched to 50 ohms at the input and output is about as simple as an RF component can get. Fabricated in silicon or GaAs, these products are available from several major manufacturers. The circuits illustrated in this article show biasing and internal descriptions for the common Darlington amplifiers implemented using silicon bipolar technology. Most of these applications (but not all) can be accomplished with GaAs devices, as well. The information given here is provided as a general reference only — consult the manufacturer's data for power supply and biasing information for each model of MMIC amplifier, whether silicon or GaAs.

Figure 1 shows a simplified internal circuit diagram of a "garden variety" silicon MMIC amplifier. This Darlington configuration is used in most of the available versions. Linearity and impedance matching are achieved using feedback. R_p provides parallel feedback, while R_s controls series feedback. R_{b1} and R_{b2} set the DC bias level for the two transistor bases.

Variable specifications include frequency range, power handling capability, noise figure, and power consumption. Some models are available with feedback optimized for highest gain or lowest noise figure, and give up a well-matched input in exchange. These devices require input matching to obtain specified performance, but usually the network required is simpler than would be needed for a discrete transistor.

Application Circuits

The simplest application is as an amplifier in a 50 ohm system. Figure 2 shows how easily these components can be applied. The MMIC amplifier, input and output DC blocking capacitors, and a decoupling choke to the power supply represent the minimum number of components.

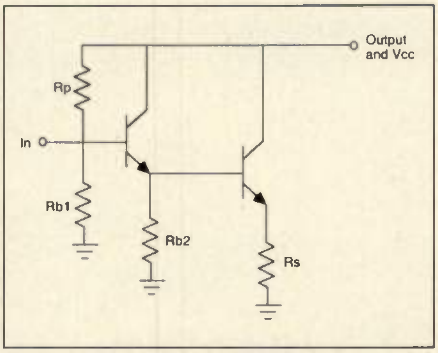


Figure 1. basic diagram of a Darlington amplifier, as used in most silicon MMIC amplifiers.

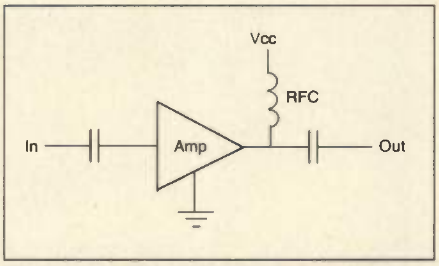


Figure 2. The simplest implementation of a MMIC amplifier.

If the available power supply voltage is higher than the ratings of the MMIC (typically in the 5 volt range), a voltage dropping resistor may be added (R_{cb}) to establish the collector bias at the proper voltage, as shown in Figure 3. In some cases, where R_{cb} is much larger than 50 ohms or where some mismatch is acceptable, the RF choke may be eliminated as a cost saving measure.

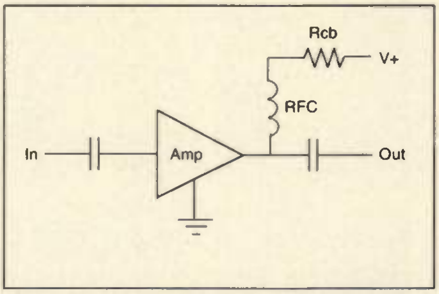


Figure 3. Addition of a series dropping resistor allows operation at higher supply voltages.

Nearly all MMIC amplifiers are designed for direct cascading (Figure 4). For example, two amplifiers with 13 dB gain can be cascaded to obtain 26 dB gain, with little additional cost and complexity. However, when cascading amplifiers, layout becomes important, as the system gain will create an oscillator if adequate input to output isolation is not maintained. Broadband gains to 60 dB can readily be achieved in this configuration.

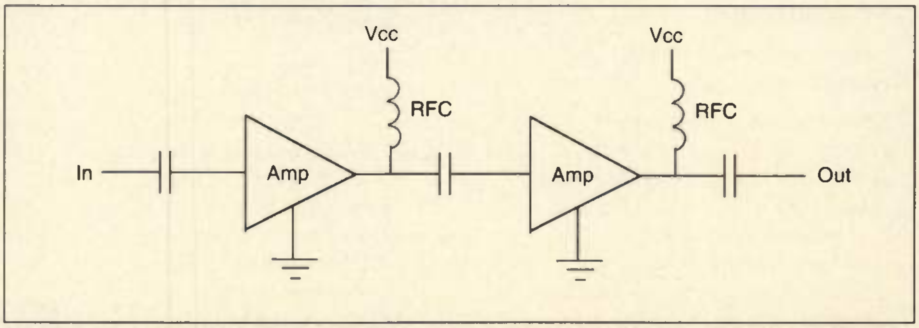
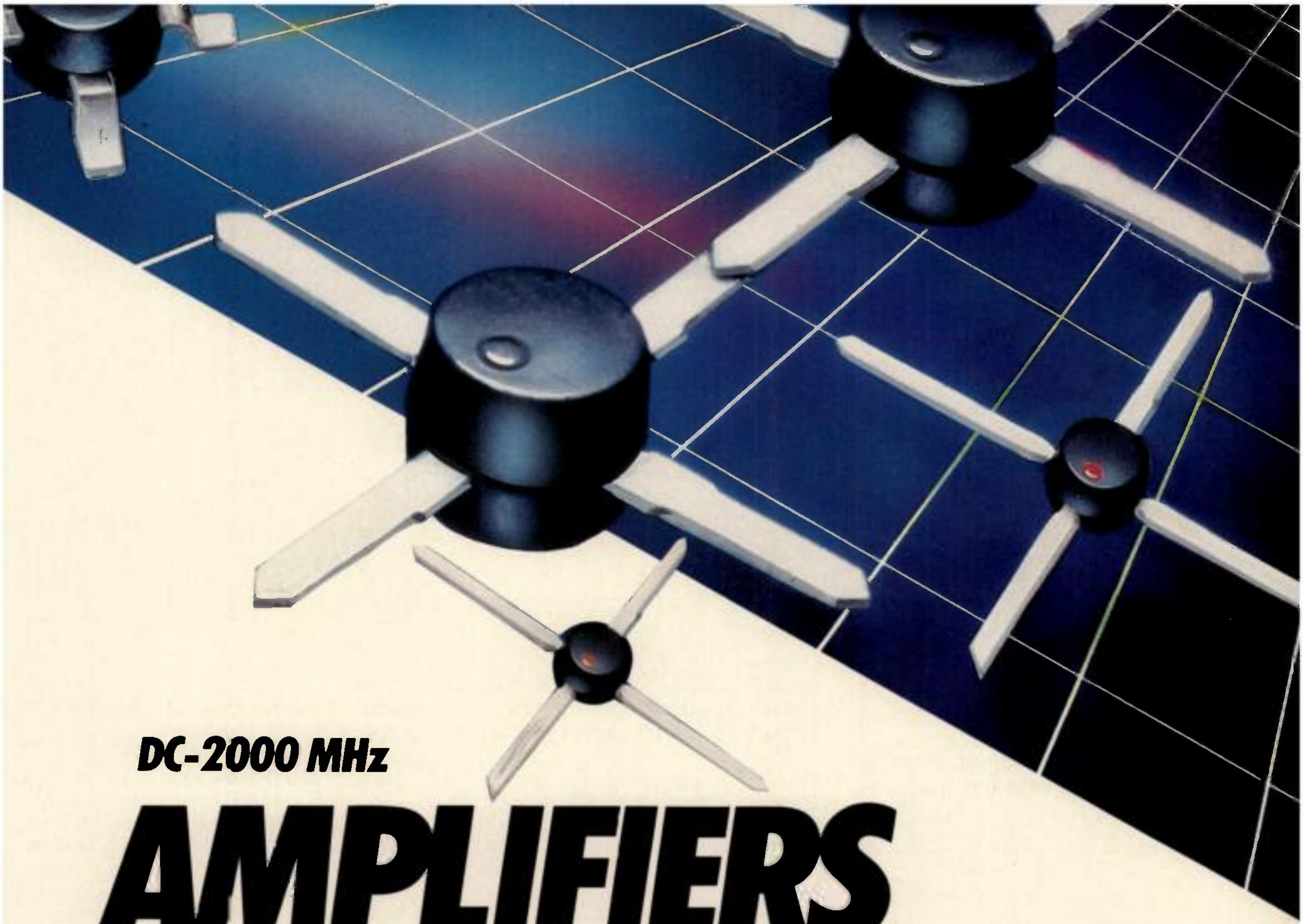


Figure 4. Most MMIC amplifiers can be cascaded, with DC blocking as the only interstage connection requirement.



DC-2000 MHz

AMPLIFIERS

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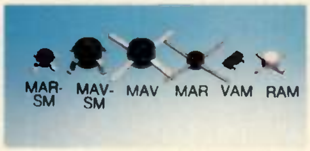


Models above shown actual size

from **99¢** Unit price $ (25 qty)

PLASTIC SURFACE-MOUNT			++VAM-3 1.45		+VAM-6 1.29	++VAM-7 1.75		
add suffix SM to model no. (ex. MAR-ISM)	MAR-1 1.04	MAR-2 1.40	MAR-3 1.50	MAR-4 1.60	MAR-6 1.34	MAR-7 1.80	MAR-8 1.75	
	MAV-1 1.15	+MAV-2 1.45	+MAV-3 1.55	MAV-4 1.65				MAV-11 2.15
CERAMIC SURFACE-MOUNT	RAM-1 4.95	RAM-2 4.95	RAM-3 4.95	RAM-4 4.95	RAM-6 4.95	RAM-7 4.95	RAM-8 4.95	
PLASTIC FLAT-PACK	MAV-1 1.10	+MAV-2 1.40	+MAV-3 1.50	+MAV-4 1.60				MAV-11 2.10
	MAR-1 0.99	MAR-2 1.35	MAR-3 1.45	MAR-4 1.55	MAR-6 1.29	MAR-7 1.75	MAR-8 1.70	
Freq.MHz,DC to	1000	2000	2000	1000	2000	2000	1000	1000
Gain, dB at 100MHz	18.5	12.5	12.5	8.3	20	13.5	32.5	12.7
Output Pwr. +dBm	1.5	4.5	10.0	12.5	2.0	5.5	12.5	17.5
NF, dB	5.5	6.5	6.0	6.5	3.0	5.0	3.3	3.6

Notes: + Frequency range DC-1500MHz ++ Gain 1/2 dB less than shown

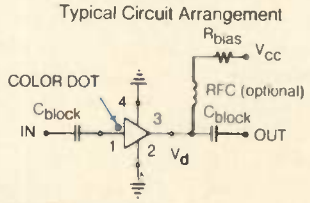
designer's amplifier kits
DAK-2: 5 of each MAR-model (35 pcs), only $59.95
DAK-2SM: 5 of each MAR-SM model (35 pcs) only $61.95
DAK-3: 3 of each MAR, MAR-SM, MAV-11, MAV-11SM (48 pcs) $74.95

designer's chip capacitor kit
KCAP-1: 50 of 17 values, 10pf to 0.1µf (850 pc), $99.95

chip coupling capacitors at .12¢ each (50 min.)

Size (mils)	Value
80 x 50	10, 22, 47, 68, 100, 220, 470, 680 pf
80 x 50	1000, 2200, 4700, 6800, 10,000 pf
120 x 60	.022, .047, .068, .1µf

Typical Circuit Arrangement



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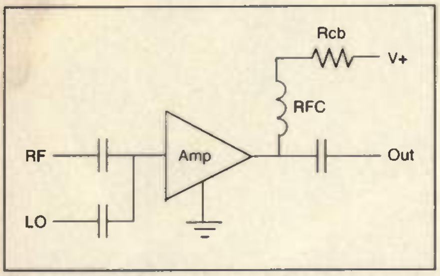

Figure 5. A simple mixer using a MMIC amplifier.

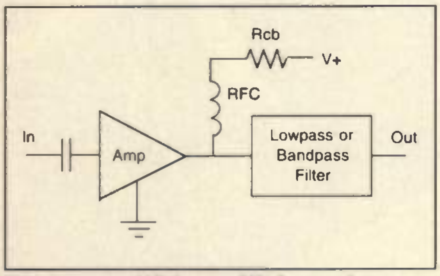

Figure 6. With sufficient drive, a MMIC can be an efficient frequency multiplier.

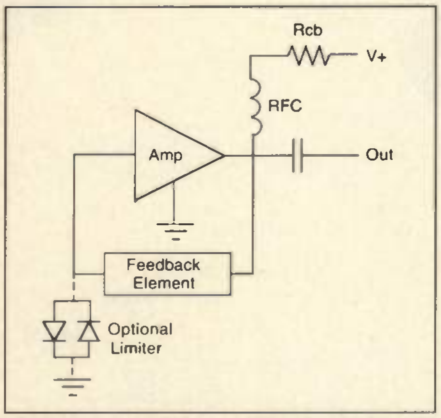

Figure 7. Basic oscillator circuit using a MMIC amplifier.

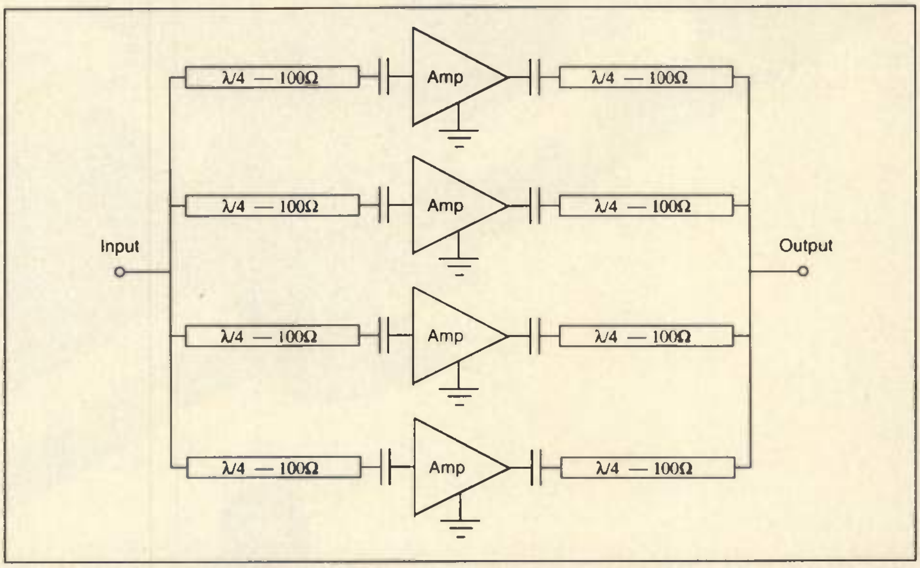

Figure 8. Quadrature combining, practical in microstrip, can be used to obtain a medium power output from several MMIC amplifiers.

Nonlinear Circuits

Mixers, multipliers and oscillators can be constructed using MMIC amplifiers. Figure 5 shows a mixer circuit, with both the RF and local oscillator applied to the MMIC input terminal. This configuration provides no RF/LO isolation, of course, relying on the preselector filter for RF port isolation. Operation of this mixer is similar to a simple single-transistor circuit often used (in the past) in inexpensive consumer radios. Performance is modest at best, but with unarguable simplicity.

Figure 6 shows the simplicity of a MMIC frequency multiplier. The drive level must reach saturation of the input transistor, or sufficient harmonic energy may not be generated. The gain of the amplifier makes it possible to build low loss 3× and 5× multipliers.

MMIC oscillators provide an interesting design exercise. They provide the necessary gain to maintain a gain of greater than one, including circuit losses and energy extracted at the output. Amplitude control can be achieved in either of two ways: saturation of the amplifier, or an external limiter.

The simplest oscillator uses a transmission line between the input and output terminals, with oscillation occurring at the frequency where the phase shift of the amplifier and transmission line combined is 360 degrees. At lower frequencies, the transmission line can be replaced with a LC network that provides the necessary phase shift.

At low frequencies (below 100 MHz or so), MMICs maintain close to 180 degree phase shift, but at higher frequencies, the phase shift changes. Consult the data sheets or obtain experimental results to confirm the phase shift at the desired frequency.

One more application not illustrated is a squaring circuit to make digital-compatible signals from a sine wave input. If the input is driven into hard saturation, a pretty good square wave will be created. The output can be DC-coupled with selection of the proper dropping (pull-up) resistor. Alternatively, the output can be AC-coupled, with high-impedance bias to hold the digital input at the mid-point of the logic family's voltage swing.

All nonlinear applications of MMICs require attention to collector (or drain) biasing. Since the circuit is driven into saturation in most of these applications, the specified current rating can be exceeded. R_{cb} must be carefully chosen to provide current limiting, or the power supply must provide constant current. Note that GaAs MMICs are more difficult to use in nonlinear applications since they have higher linearity and do not saturate as easily as silicon devices.

Combined Amplifiers

MMIC amplifiers rarely provide more than +13 dBm (20 mW) power output. A few models offer higher power, but at higher cost and sometimes at the expense of bandwidth. Combining amplifiers to obtain medium power (+20 dBm or more, combining amplifiers may be an inexpensive design choice.

At low frequencies, MMICs may simply be connected in parallel to obtain higher power. Broadband transformers can restore the system to 50 ohms. However, linearity usually suffers, since the amplifiers are not precisely matched.

The quadrature combining scheme of Figure 8 is another option, where the low cost of microstrip can be utilized, and relatively narrow bandwidth is required. on G10 or FR-4 type p.c. board material, the 1/4λ lines can be very short. the characteristic impedance of the lines can be adjusted to accommodate nearly any number of amplifiers, with 2 to 4 being the most common.

This short note illustrates some of the uses that "universal building block" MMIC amplifiers can provide with simplicity and economy. *RF*

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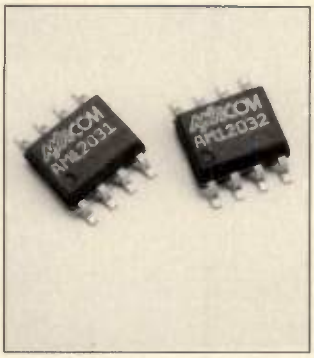

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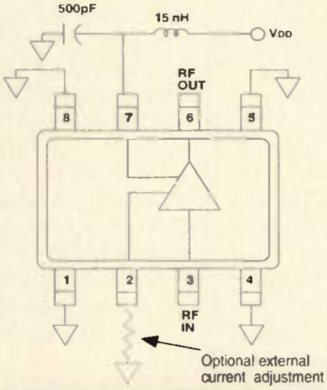

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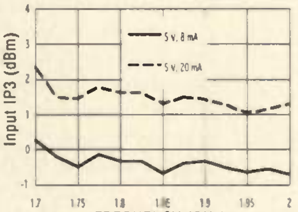

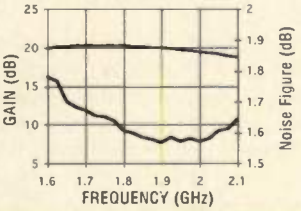

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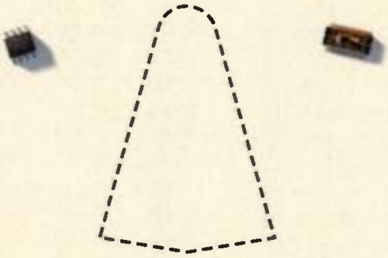

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Using Design of Experiments to Optimize Filter Tuning Steps

by Dan Pleasant
Hewlett-Packard Company

Computer-aided engineering (CAE) software continues to reach new levels of sophistication, enabling designers to tackle new classes of design problems. This article describes how two features of CAE software can be used in a novel fashion to solve an old problem — the design of tunable high-Q filters for high manufacturing yield — using Design of Experiments and post-production tuning yield analysis.

Design of Experiments (DOE) is a capability that has only recently been added to CAE software. Since most high-frequency design engineers are not familiar with DOE or its capabilities, a brief introduction is in order.

DOE has been used for many years in manufacturing to enhance the reliability of processes in fields as diverse as automotives and pharmaceuticals. For any manufacturing process, there are parameters that can be controlled and others that cannot. Any of these parameters can affect the manufacturing yield. DOE is a systematic technique for finding the parameters that have the greatest effect on the process (and therefore the manufacturing yield), and finding ways to enhance the process and the reliability of the product. For high-frequency circuit designers, the parameters of the manufacturing process are usually circuit component values, although other parameters such as temperature are often also important.

A complete description of the DOE technique would be quite lengthy and well beyond the scope of a short article. For high-frequency designers, though, here is a brief description of DOE:

Imagine a simple circuit such as the one shown in Figure 1. There are three parameters in this circuit: two resistor values, and one capacitor value.

Suppose the values of the resistors can vary in manufacturing from 24 to 26 ohms, and the value of the capacitor can vary from 0.9 to 1.1 pF. Using the DOE technique, this circuit can be simulated by the computer using the high and low values for each parameter for all possible combinations of parameter values. This circuit would then would be simulated a total of eight times, with parameter values as shown in the following table:

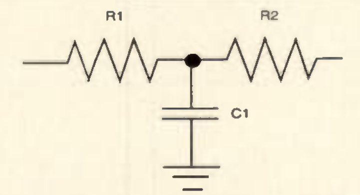

Figure 1. Simple lowpass filter example.

	R1	R2	C1
Simulation #1	24	24	0.9 pF
Simulation #2	24	24	1.1 pF
Simulation #3	24	26	0.9 pF
Simulation #4	24	26	1.1 pF
Simulation #5	26	24	0.9 pF
Simulation #6	26	24	1.1 pF
Simulation #7	26	26	0.9 pF
Simulation #8	26	26	1.1 pF

Table 1. Parameter values.

This table defines an "experiment" in the terminology of DOE. Using the results of these eight simulations, the circuit's sensitivity to each component value can be seen. Furthermore, interactions among the three components can be estimated. All of this information is useful to circuit designers, and can be used to enhance manufacturing yields, as will be demonstrated later.

Because it uses all possible combinations of parameter values, the experiment defined in this example is called a "full-factorial" experiment. Often, the designer knows that some interactions are not important and do not significantly affect the circuit's performance. Then, some parts of the experiment (corresponding to rows in the table above) can be eliminated. This is important because full-factorial experiments can become computationally expensive for circuits with many parameters.

Post-Production Tuning Yield

Post-production tuning yield is, quite simply, a combination of circuit optimization and Monte Carlo analysis for the design of circuits that include tunable components.

For circuits that do not use tunable components, Monte Carlo analysis calculates manufacturing yields by simulating the circuit many times (hundreds, or even thousands of times), statistically varying each parameter for each simulation. Eventually, the computer will gather enough information about the circuit to estimate its manufacturing yield (within a user-specified uncertainty interval).

Post-production tuning yield extends the concept of Monte Carlo analysis to cover the case of circuits that contain tunable components. At each Monte Carlo iteration, the computer statistically varies the circuit parameters. Then it performs an optimization on the circuit. In fact, the computer may perform several optimizations on the circuit, with each optimization tuning different components to mimic the actions of manufacturing personnel. The Monte Carlo iteration is not finished until the optimization is finished, resulting in a yield-after-tuning simulation.

Tunable Filter Design for High Yield

Used together, DOE and post-production tuning yield analysis can be used to design high-Q, manually tuned filters. Filter tuning has always been considered something of a black art, because of two difficult-to-solve problems: determining which filter components to tune, and in what order to tune them. The following example illustrates one way to solve both problems.

Figure 2 shows a simple lumped-element design for a Chebyshev lowpass filter of order 15. The component values for this circuit were calculated in the classic fashion, which means that they were found in a filter cookbook and simply typed into the computer.

The specifications for this filter are straightforward. The filter must have less than 3.6 dB of loss up to 4.75 MHz, and must have at least 36 dB of loss at 5 MHz and above. To allow for some mar-

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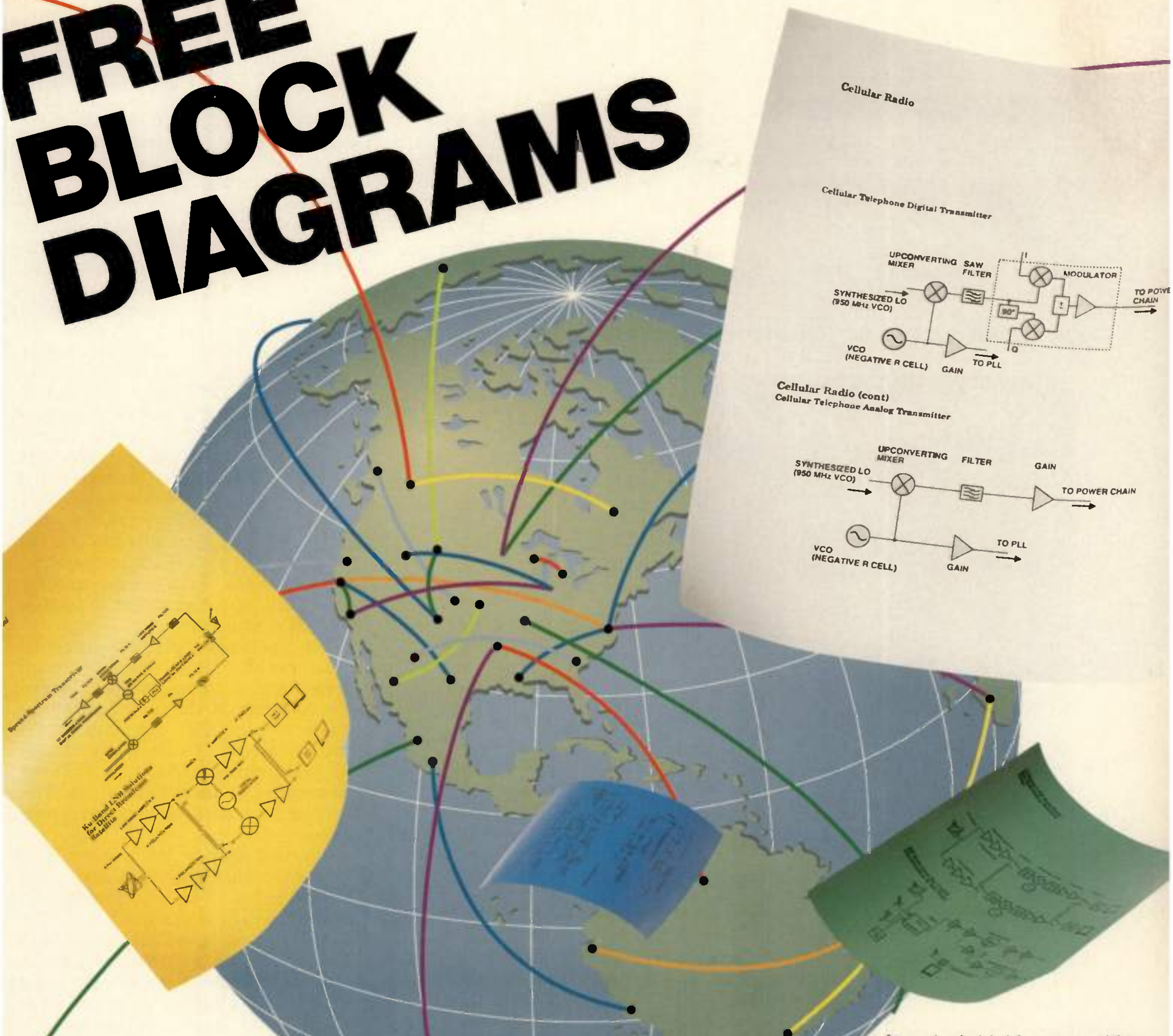

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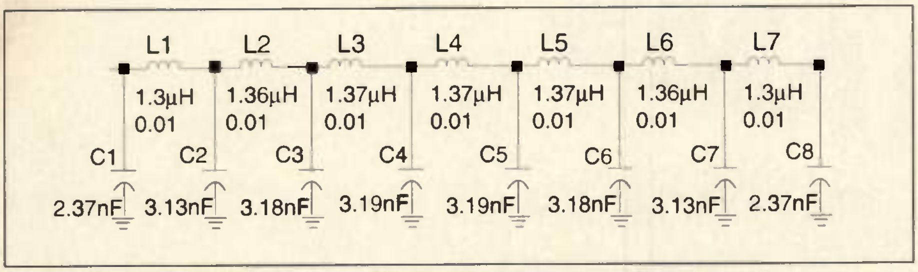

Figure 2. Perfect 15th order, Chebychev lowpass filter.

gin, the filter is designed for 3 dB of ripple. The 15-pole filter design is dictated by the need for a sharp dropoff in the stopband response.

When simulated, this filter exhibits excellent 3-dB Chebyshev response, as shown in Figure 3. This plot shows the filter's response assuming that all components are perfect.

Unfortunately, the designer is under pressure to use low cost components and to make the manufacturing process as simple and as inexpensive as possible. The designer is certain that real filters will not exhibit a perfect Chebyshev response, and some of the components must be tuned on the assembly line. This circuit is to be built using surface-mount technology (SMT) components wherever possible, and there are really very few tunable SMT inductors available at a low price. Some of the capacitors, therefore, must be tunable. The problem is to determine which capacitors should be tunable, and how to tune them.

First, a yield calculation is made on the "ideal" circuit. This is a straightforward Monte Carlo simulation, without any tunable components. Using published tolerances for good SMT components, the simulator predicts a yield of only 42% for this circuit. Roughly half of the filters will fail because of the passband specification, half because of the stopband specification.

The next step in the design is to choose tunable components and, using post-production tuning yield analysis, calculate the yield of the tunable filter. If designers do not have CAE software with DOE, and if they are not proficient with mathematics, a possible next step would be to choose a capacitor or two at random, find the yield after tuning those capacitors, and repeat until a good solution was found. Figure 4 shows a modified filter circuit with the last two capacitors defined to be variable.

The circuit in Figure 4 was analyzed

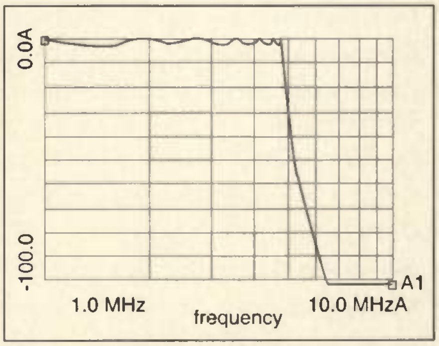

Figure 3. Calculated response of filter of Figure 2.

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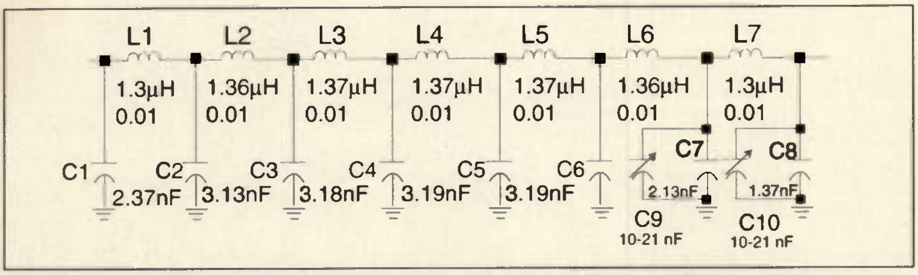

Figure 4. Lowpass filter with randomly selected tunable capacitors.

under the following manufacturing rules: First, the capacitor on the far right was adjusted to obtain acceptable passband performance (if possible). Then, the other capacitor was adjusted for best stopband performance. It was assumed that this second adjustment would derate the passband performance, so the first adjustment was repeated, followed by a repeat of the second adjustment. All other components in the filter were varied at random by the Monte Carlo analysis.

Given the tunable circuit in Figure 4, the calculated manufacturing yield was 62%, an increase of 20% over the untuned original circuit. In fact, this is not a bad result if one considers that the tunable capacitors, as well as the tuning procedure, were chosen entirely at random. Now, two other capacitors can be chosen, the experiment repeated, and a new yield calculated. This can be repeated until the designer is either satisfied with the yield, or gets tired of the whole exercise and decides to go home.

Selecting Tuning Elements Using DOE

Using the DOE capability, the designer can find the most logical tuning elements, and the best order in which to tune them, in a single step. For this filter, a full-factorial DOE analysis was run. Since the capacitors are the only elements that potentially can be tuned, every capacitor was allowed to vary in the DOE analysis, and every inductor was held constant at the nominal value. Since there are eight capacitors, this results in 28, or 256, total simulations of the circuit. (The computer automatically executes all 256 simulations with no user intervention.) This is a simple circuit, and 256 simulations is well within the range of possibility using even a modestly equipped computer workstation, and takes only a few minutes on a more powerful workstation.

After the DOE analysis is completed, the software presents several different data displays for interpreting the data. Perhaps the most useful display is the Pareto chart. The Pareto chart for this analysis is shown in Figure 5.

Although the Pareto chart looks something like a histogram, it is actually nothing of the sort. Each "bar" in the Pareto chart illustrates the relative sensitivity of a circuit response to a parameter, or to the interdependence of two parameters. The Pareto chart in Figure 5 shows the sensitivities of two different circuit response values: the passband loss (ripple), and the stopband loss (rejection).

In Figure 5, note that the bars labeled

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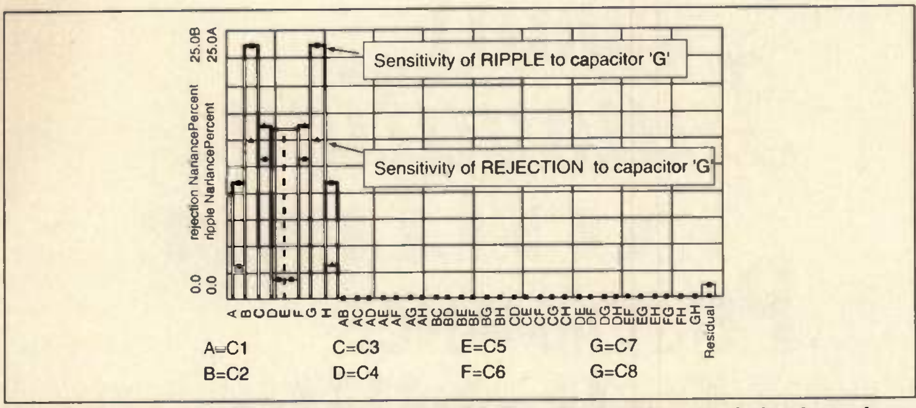


Figure 5. Pareto chart showing sensitivity of rejection and ripple to lowpass filter elements.

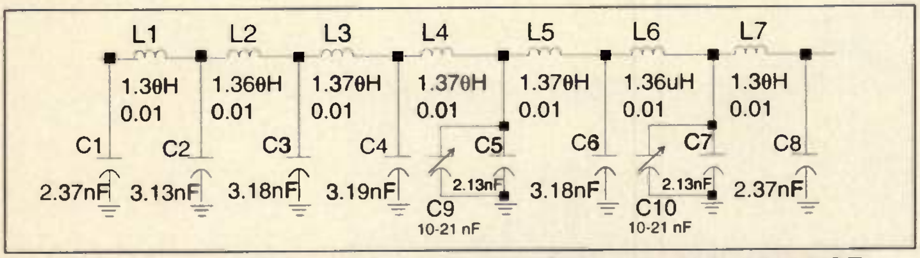


Figure 6. Lowpass filter with tunable capacitors selected using DOE.

A through H are taller than the others; AB through GH are nearly zero. Bars A through H represent the sensitivity of the filter to the capacitor values. There are two bars for each letter. One bar represents the sensitivity of the ripple to each capacitor, and the other represents the sensitivity of the rejection.

The bars marked AB through GH represent the sensitivity of the circuit to interdependencies of the capacitors. For instance, the AB bar represents the circuit's sensitivity to the product of the values of capacitors A and B. These interdependencies are all small, making the analysis of this circuit rather easy. Note that these interdependencies become much more significant if the capacitor values are allowed to change over a wider range, so high-quality capacitors are important.

An examination of the plot shows that the bars marked "E" have a large effect on the filter's rejection, but a small effect on ripple. The bars marked "G" have the largest effect on ripple, but they also have a large effect on rejection. Therefore, it is reasonable to tune this filter in the following way:

1. Tune the "G" capacitor (which is the next-to-last capacitor on the right in the circuit diagram) for best ripple. This will have an effect on rejection as well, but that can be ignored at this step.

2. Tune the "E" capacitor for best rejection. This tuning step will probably bring the filter into compliance, as it will have only a small effect on the ripple.

3. To ensure the best filter performance, repeat steps 1 and 2.

Figure 6 displays the new filter circuit with the appropriate capacitors set up for tuning.

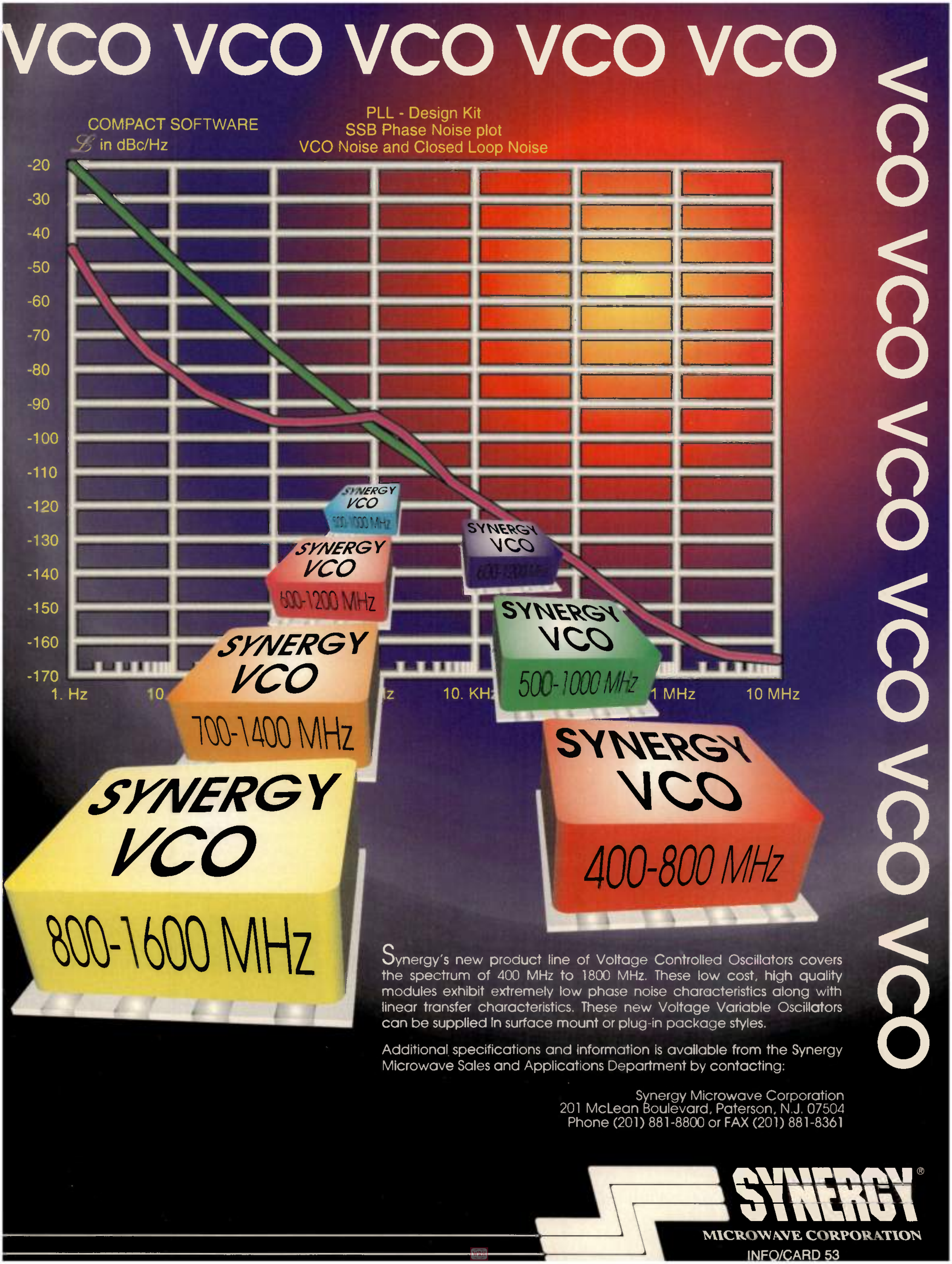
Using post-production tuning yield analysis, this circuit can be analyzed according to the tuning rules suggested by the DOE analysis. For this example, the circuit's yield becomes 72%, up from 42% with no tuning and 62% with randomly chosen tuning with no increase in manufacturing cost or complexity. Obviously, higher yields can be obtained by either tightening tolerances or making more components tunable.

Conclusion

Using two new features of high-frequency CAE software, Design of Experiments and post-production tuning yield analysis, designers can approach difficult problems in new ways. The new features, used singly and in combination, are powerful tools for designing high-yield circuits down to the last detail of tuning on the manufacturing line. This example has shown how to apply these new tools to the design of high-Q filters. Designers will doubtless find other creative uses for them as well. *RF*

About the Author

Dan Pleasant is an application engineer with Hewlett-Packard's HP-EEsof operation in Santa Rosa, California. He received his BSEE from the University of Colorado in 1979 and his MSEE from Stanford University in 1981. He had been with the MIT Lincoln Laboratory, Lexington, Massachusetts, prior to joining Hewlett-Packard in 1989. He can be reached at Hewlett-Packard Co., HP-EEsof, 1400 Fountaingrove Parkway, Santa Rosa, CA 95403.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Phase-Locked Loop Parameters and Filters

By Jack Porter

Phase-locked loops are often required to have specified values for closed-loop parameters such as damping factor and undamped natural frequency or noise bandwidth. However, the parameters used in loop filter design are open-loop gain and corner frequencies; calculating values for these is thus the first step in designing such filters.

Table 1 defines symbols used in Table 2, which contains formulas for open- and closed-loop gain and equations showing the relationships among the parameters in these formulas. Figure 1 is a typical open-loop gain plot with some of these parameters noted on it.

A loop with a single corner frequency is completely specified by choosing values for any two of the parameters appearing in the equations in Table 2; values for all of the remaining ones can then be calculated.

An additional corner frequency, designated F_3 in Figure 1, is often used to attenuate noise outside the loop bandwidth. This may represent a single-pole filter used to attenuate phase detector switching noise or an anti-aliasing filter when a digital loop filter is used. With the additional corner frequency the loop is no longer second-order, thus parameters such as ζ and ω_n are no longer defined. However, closed-loop bandwidth can still be calculated using the equations in the last part of Table 2. Since the loop phase margin has been reduced, the bandwidths calculated using these equations are always greater than those of the basic loop.

The equations in Table 2 can be used in designing any type of loop filter, digital, active analog or, with some approximations, passive. The single-op amp active filter, various forms of which are shown in Figure 2, is simple and widely used.

Figure 2 also contains the equations used in designing these filters. Definitions of the symbols used in these equations can be found in Table 1. Figure 2-1 is the basic loop filter; the remaining three filters represent various ways of implementing the corner at F_3. When these filters are used, a value must be selected for F_3 or T_3 in addition to the two parame-

Open loop gain:

$$A_0 = \frac{K_a}{s^2}(T_2 s + 1)$$

$$= \frac{2\zeta\omega_n s + \omega_n^2}{s^2}$$

Closed loop gain

$$A_0 = \frac{K_a T_2 s + K_a}{s^2 + K_a T_2 s + K_a}$$

$$= \frac{2\zeta\omega_n s + \omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}$$

$$\omega_0 = 2\pi F_0$$

$$\omega_2 = 2\pi F_2 = 1/T_2$$

$$\omega_H = 2\pi B_H$$

$$B_L = B_N/2$$

$$2\zeta\omega_n = K_a T_2$$

$$\omega_0 \equiv K_a T_2$$

$$K_a = \omega_n^2 = \omega_0\omega_2$$

$$\zeta = \frac{\omega_n T_2}{2} = \frac{1}{2}\sqrt{\omega_0 T_2}$$

$$B_N = \omega_n\left[\zeta + \frac{1}{4\zeta}\right] = \frac{1 + K_a T_2^2}{2T_2}$$

$$= \pi(F_0 + F_2)$$

$$\omega_H = \omega_n\sqrt{(2\zeta^2+1) + \sqrt{(2\zeta^2+1) + 1}}$$

$$= \omega_n \exp\left(\frac{1}{2}\sinh^{-1}(2\zeta^2+1)\right)$$

$$\omega_H^4 - (2 + K_a T_2^2)K_a\omega_H^2 - K_a^2 = 0$$

For loop with additional corner frequency F_3:

$$A_0 = \frac{\frac{K_a}{s^2}(T_2 s + 1)}{(T_3 s + 1)}$$

$$\omega_3 = 2\pi F_3 = 1/T_3$$

$$A_C = \frac{K_a T_2 s + K_a}{T_3 s^3 + s^2 + K_a T_2 s + K_a}$$

$$B_N = \frac{1 + K_a T_2^2}{2(T_2 - T_3)}$$

$$T_3^2\omega_H^6 + (1 - 2K_a T_2 T_3)\omega_H^4$$
$$-(2 + K_a T_2^2)K_a\omega_H^2 - K_a^2 = 0$$

Table 2. Open- and closed-loop gain formulas and related parameters .

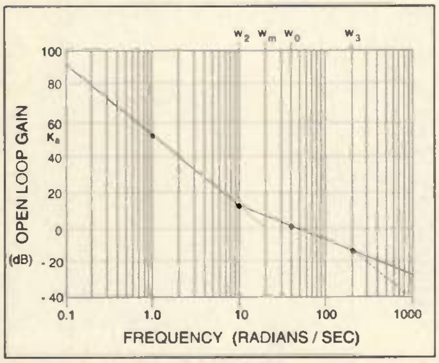

Figure 1. Typical PLL open-loop gain plot.

$A_c(s)$	:	Closed-loop gain
$A_o(s)$	:	Open-loop gain
B_H	:	Closed-loop half-power (–3 dB) BW (Hz)
B_L	:	Closed-loop noise BW (Hz)
B_N	:	Closed-loop 2-sided (equivalent i-f) noise BW (Hz)
F_0	:	Nominal open-loop unity-gain BW (Hz)
F_2	:	2nd-order loop filter corner frequency (Hz)
F_3	:	Optional additional corner frequency (Hz)
K_a	:	Open-loop gain constant (sec^{-2}), numerically equal to the open loop gain at ω=1
K_{pd}	:	Phase detector sensitivity (volts/rad)
K_{vco}	:	VCO sensitivity (Hz/volt)
Nm	:	Frequency multiplier ratio
N_d	:	Frequency divider ratio
T_f	:	Loop filter gain constant (sec), numerically equal to the reciprocal of the loop filter gain at ω=1
ζ	:	Closed-loop damping factor
ω_n	:	Closed-loop undamped natural frequency (rad/sec)

Table 1. Definitions of symbols used in Table 2.

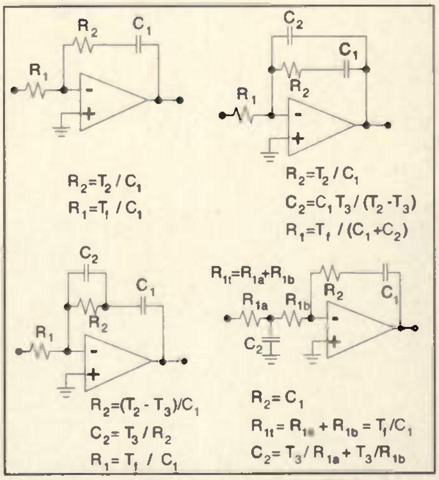

Figure 2. Various active PLL loop filters.

ters required for the basic filter.

It isn't difficult to solve most of the equations in Table 2 and Figure 2 using a scientific calculator. (The last equation in Table 2-2 may present a problem.) In the distant past they were solved quite easily with a slide rule. Nevertheless, it's still easier to use the True Basic program PLLS, which performs all of these calculations.

Any one of six different combinations of the two required input parameter values may be chosen: ζ & ω_n, ζ & B_L, ζ & B_H, F_0 & F_2, B_H & T_2 or K_a & T_2. The last two combinations are used mainly for analyzing existing loops when T_2 can be obtained from the schematic and K_a can be calculated or B_H measured. When the two values have been chosen all of the others are calculated.

The loop filter component values are calculated next, based on specified values for K_{pd} and K_{vco}. If the loop includes frequency multipliers or dividers, this ratio must also be specified. (The divider ratio is the reciprocal of the multiplier ratio). In loops which include both, e.g. phase-locked receivers, the multiplier ratio is not necessarily an integer. The filter type and values for C_1 and F_3, if applicable, are then chosen. After the component values have been determined the loop bandwidths are automatically recalculated.

Finally, the open and closed loop gain can be tabulated and plotted. Table 3 shows the printed program output corresponding to the open-loop gain plot in Figure 1.

PLLS is available from the RF Design Software Service. See page 78 for ordering information *RF*

Ka= 400 = 52.0412 dB F0= 6.3662 Hz
F2= 1.59155 Hz T2=1.00000e-01 sec
F3= 31.831 Hz T3=5.00000e-03 sec
BH= 9.48276 Hz BL= 13 1579 Hz BN= 26.3158 Hz

	Frequency (Hz)	Open Loop Gain (dB)	Open Loop Phase (deg)	Closed Loop Gain (dB)	Closed Loop Phase (deg)	Loop Error Gain (dB)	Loop Error Phase (deg)
0	.10	+60.13	-176.6	+0.01	-0.0	-0.12	+176.6
1	.13	+56.26	-175.7	+0.01	-0.0	-56.25	+175.7
2	.16	+51.99	-174.5	+0.02	-0.0	-51.97	+174.5
3	.20	+48.14	-173.2	+0.03	-0.0	-48.11	+173.2
4	.25	+44.30	-171.5	+0.05	-0.1	-44.25	+171.5
5	.32	+40.08	-169.2	+0.08	-0.1	-39.99	+169.1
6	.40	+36.30	-166.6	+0.13	-0.2	-36.17	+166.4
7	.50	+32.56	-163.5	+0.20	-0.4	-32.36	+163.1
8	.64	+28.52	-159.2	+0.31	-0.8	-28.21	+158.5
9	.80	+24.97	-154.8	+0.45	-1.5	-24.51	+153.3
10	1.00	+21.55	-149.7	+0.64	-2.6	-20.91	+147.1
11	1.25	+18.32	-144.1	+0.87	-4.5	-17.45	+139.6
12	1.60	+14.97	-137.7	+1.15	-7.9	-13.82	+129.9
13	2.00	+12.17	-132.1	+1.36	-12.3	-10.81	+119.8
14	2.50	+9.57	-127.0	+1.48	-18.4	-8.09	+108.6
15	3.20	+6.89	-122.2	+1.41	-26.8	-5.48	+95.4
16	4.00	+4.61	-118.9	+1.09	-35.7	-3.52	+83.1
17	5.00	+2.41	-116.6	+0.48	-45.7	-1.93	+70.9
18	6.40	+0.04	-115.3	-0.56	-57.4	-0.61	+57.9
19	8.00	-2.08	-115.4	-1.84	-68.4	+0.24	+47.0
20	10.00	-4.22	-116.5	-3.41	-79.3	+0.81	+37.2
21	12.50	-6.41	-118.7	-5.28	-90.1	+1.14	+28.5
22	16.00	-8.94	-122.4	-7.66	-101.9	+1.28	+20.5
23	20.00	-11.36	-126.7	-10.11	-112.2	+1.25	+14.5
24	25.00	-13.95	-131.8	-12.83	-122.0	+1.12	+9.8
25	32.00	-17.05	-138.0	-16.14	-132.0	+0.91	+6.0
26	40.00	-20.07	-143.8	-19.36	-140.1	+0.71	+3.6
27	50.00	-23.30	-149.3	-22.78	-147.2	+0.52	+2.1
28	64.00	-27.07	-155.0	-26.72	-153.9	+0.35	+1.1
29	80.00	-30.63	-159.4	-30.38	-158.8	+0.24	+0.6
30	100.00	-34.28	-163.3	-34.12	-162.9	+0.16	+0.3

Table 3. Example PLL frequency response.

About the Author

Jack Porter is retired from Cubic Corporation in San Diego and previously worked at Genral Dynamics. He has a BS degree in physics from Michigan State and an MSEE from San Diego State. He can be reached at 2481 Root St., San Diego, CA 92123 or by e-mail at Compuserve ID no. 75279,1265.

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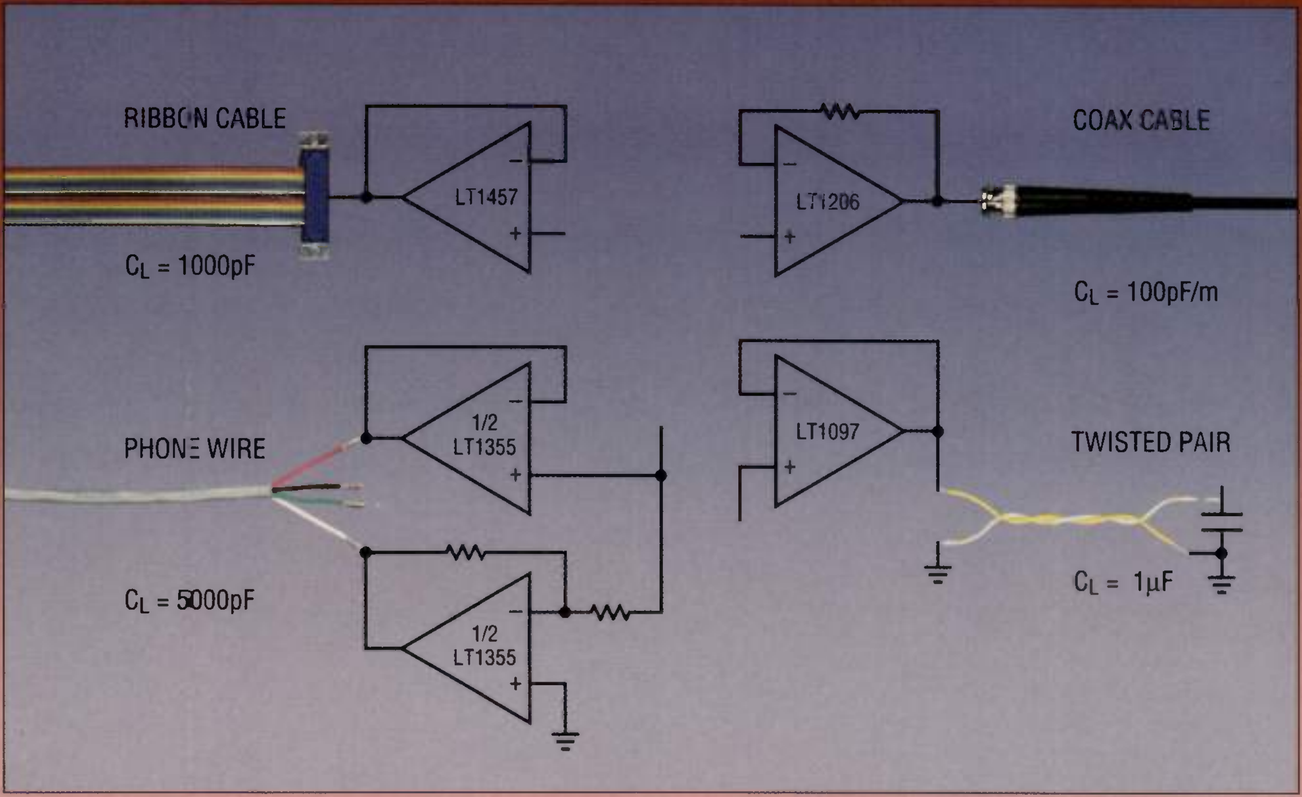

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LT1224*	2.0mV	8µA	24mA	45MHz	400V/µs
LT1354*	800µV	300nA	30mA	12MHz	400V/µs
LT1357*	600µV	500nA	30mA	25MHz	600V/µs
LT1360*	1.0mV	1µA	40mA	50MHz	800V/µs
LT1363*	1.5mV	2µA	70mA	75MHz	1000V/µs
LT1457	800µV	75pA	10mA	1.7MHz	4V/µs

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INFO/CARD 69

Power Amplifier Design Using Quadrature Hybrids

By Povl Raskmark
Aalborg University

Cancer treatment using heat (hyperthermia) has been investigated for several years; deep seated tumors may be heated using RF power and a phased array system. In the Danish Hyperthermia Foundation system, a four applicator array (1) is used. The system is phase locked near 70 MHz, with a four channel power amplifier controlled by a signal generator.

The most important design goals for the RF power amplifiers were linear operation at 70 MHz, output power of 350 watts, 50 ohm output impedance, and good reliability and low cost.

The frequency and power design goals are based primarily on theoretical considerations and practical experience with the applicator array. A low output VSWR is an advantage when a given array excitation voltage is desired since the interaction between the applicators will be less significant (2).

Design Considerations

Linear operation with low output VSWR implies a class A amplifier, but that would certainly not be a low cost solution. A single class B amplifier would not posses a low output VSWR across the entire dynamic operating range. A design which combines two amplifiers using quadrature hybrids may be a good choice because it offers linear operation and good output match at a low cost (3).

RF power field-effect transistors in the 500-watt Gemini package intended for push-pull operation are available from various companies. The TP1940 from Motorola (4) seems to have good performance combined with low price. It is around this transistor that the balanced power amplifier and the block diagram of Figure 1 are designed.

The input signal from the signal generator is amplified by a 10 watt class A-B preamplifier. At the output, a 5th order Cauer- Chebyshev lowpass filter is added to attenuate the 2nd and 3rd harmonics. A directional coupler follows the filter. These three circuits will not be described further.

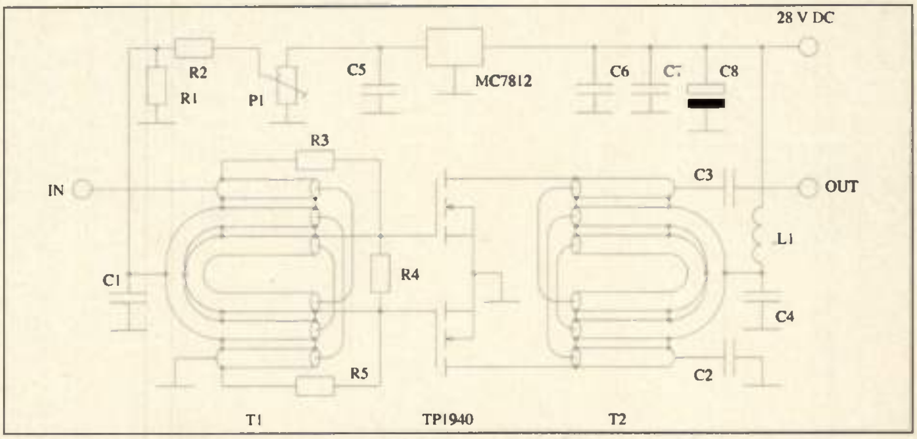

Figure 2. Circuit diagram for a 200 Watt push-pull power amplifier.

The Quadrature Hybrid

A hybrid is a four-port passive symmetric device, which ideally splits an input signal at one port into two signals with 90 degree phase shift when the remaining isolation port is matched. If the two output ports are connected to loads with reflections Γ_a and Γ_b, the input reflection coefficient Γ_i may be found:

$$\Gamma_i = 1/2(\Gamma_a - \Gamma_b) \quad (1)$$

To achieve input matching for the balanced amplifier, it is therefore sufficient if the two amplifiers are identical. Output matching is achieved in exactly the same way.

Quadrature hybrids in this frequency and power range are commercially available; however, branch line couplers are easily fabricated using printed circuit techniques (5). To achieve reasonable dimensions, the coupler is made in stripline using PTFE (Ultralam 2000) for the 50 ohm lines, and epoxy glass (G-10) for the 35.4 ohm lines. The two sets of stripline are stacked and the ground planes soldered together and then mounted on a heat sink to avoid high temperatures in the PCB. The input hybrid could be made much smaller using lumped elements, but the stripline technique is very easy to make in large quantities.

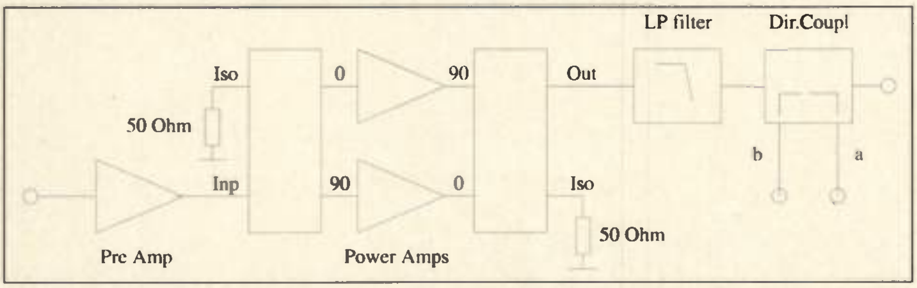

Figure 1. Principle of a 350 Watt quadrature combined power amplifier.

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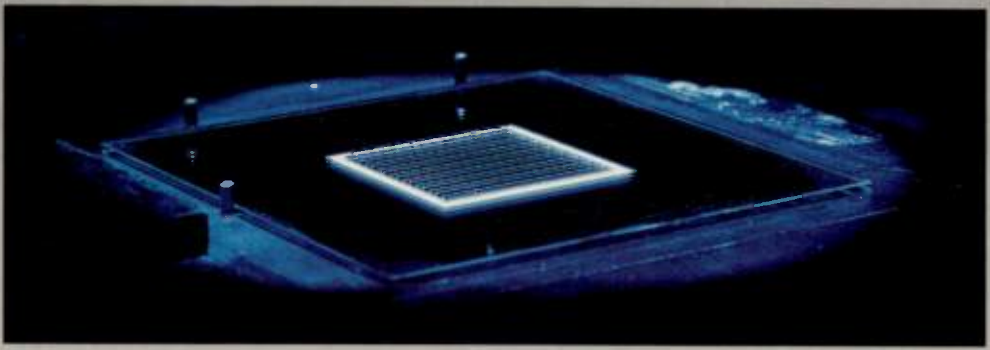

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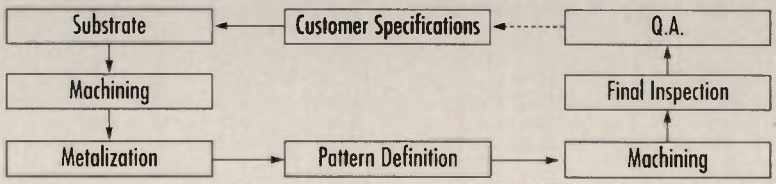

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INFO/CARD 56

R1, R2: 1k, 1/4 W
R3, R4, R5: 15 ohms, 1 W
P1: 1k

C1, C7: 500 pF (ATC)
C2, C3: 82 pF (ATC)
C4: 1 nF (UNELCO) & 3 x 100 pF (ATC) in parallel
C5, C6: 47 nF, 100 V
C8: 100 _F, 63 V

L1: 10 turn, 1.2mm enameled Cu wire 4mm inside diameter

T1: 1:9 input transformer, 25 ohm semirigid coax, 0.9 mm outside diameter (Suhner EZ-34-35 or equivalent). Two pieces 53 mm and two pieces 35 mm total length, approx. 6mm of outer conductor and 3 mm of insulator removed from each end. The four transmission lines are soldered together along the outer conductor (as shown in Figure 3), loaded with 4C6 ferrite, and then the inner conductors connected. T2: 1:9 output transformer, each line in Figure 2 represents two 50 ohm semirigid cables in parallel. Cable is 1.2 mm outside diameter (Suhner type EZ-47-TP-M-17 or equivalent). Four pieces 45 mm and four pieces 65 mm total length, otherwise as T1 but without ferrite.

Table 1. Power amplifier components.

A Single Power Amplifier

Taking filter and coupler losses into account, the necessary output from each of the two single power amplifiers must be 200 watts, well below the maximum capability of the TP1940. The approximate possible power from the push-pull amplifier may be found knowing the supply voltage (V_{ds}), the drain-source on-voltage (V_{dson}), and the load resistance (R_l):

$$P_O = 2\,(V_{ds} - V_{dson})^2 / R_l \qquad (2)$$

The transistor, operating with a 50 V supply and a 1:4 output transformer, delivers approximately 325 watts output power. Reducing the supply voltages to 40 V may give the wanted output power; this is a good operating point with a very high efficiency. Unfortunately, only a 28 V supply was available for the complete system. The output transformer was changed to 1:9, resulting in a lower overall efficiency, but achieving the desired output power. Lowering the supply voltages also improves ruggedness

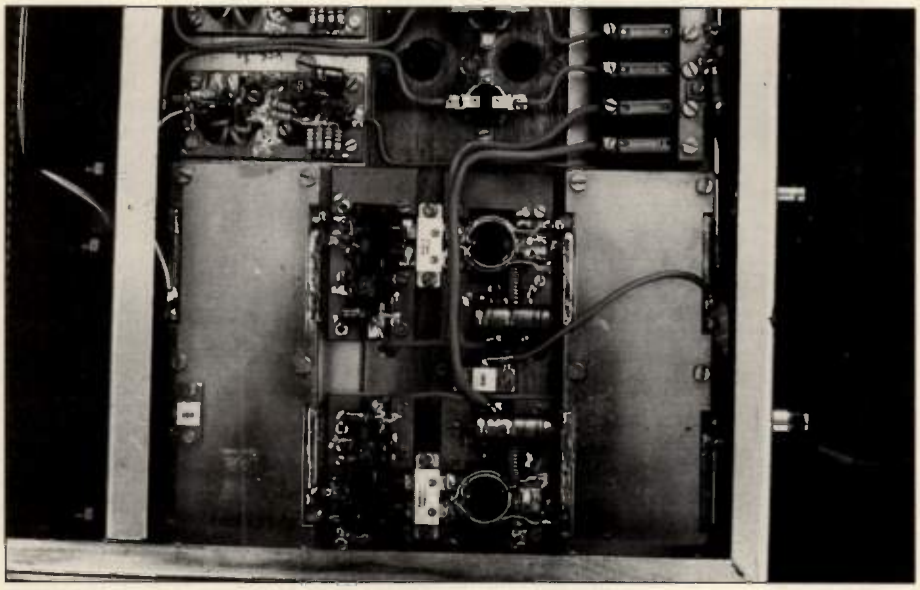

Figure 3. Photo of complete balanced amplifier.

(2), because a large mismatch will not cause a break down.

The reduced stress with lower voltage should improve reliability, although in this case it may be debatable, since the dissipated power in the transistor only decreases slightly compared to full power operation at 50 V.

When combining with quadrature hybrids each amplifier may still see a large mismatch, and so stability must be a consideration when designing the amplifiers. A theoretical approach is not possible because of lack of data, however, experiments with a single amplifier with no stabilization circuits showed good stability. However, this inherent stability can be improved using resistors, either as feedback, or as in this case, as a load directly on the gates.

The input and output matching circuits use a simple and effective transformer circuit (4), combining transformer and balun. To avoid instability, the input transformer is loaded with ferrite cores to achieve a relatively high common mode impedance at low frequencies. There is no need for ferrite cores in the output transformer because of the narrow frequency range. Note that the out-

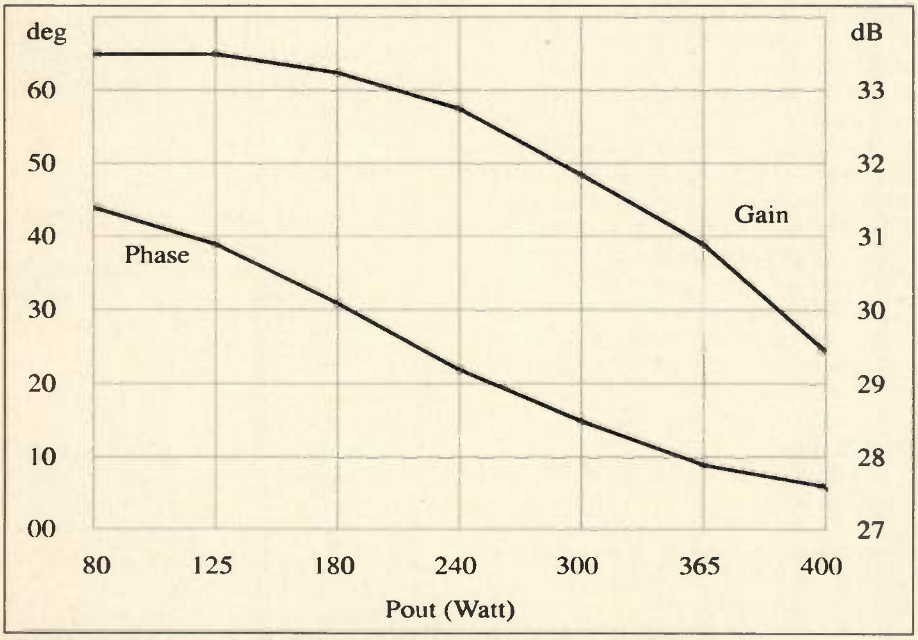

Figure 4. Gain and phase characteristics for the amplifier.

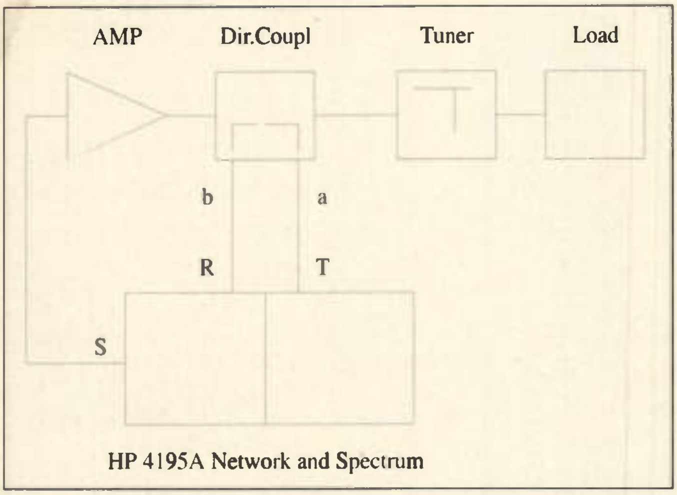

Figure 5. Passive load pull measurements using HP4195A.

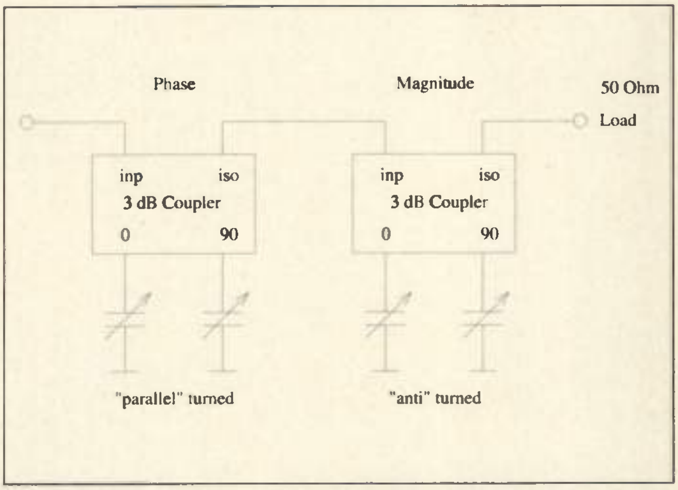

Figure 6. Tuner using two quadrature hybrids.

put tuning capacitor is divided into two capacitors, slightly improving the operating balance.

DC bias is made with a simple resistive circuit without temperature compensation, and the idle current is adjusted to only 0.1 A. The complete push-pull amplifier circuit is shown in Figure 2. Two amplifiers, together with the hybrids, etc., are mounted on a 10mm thick copper bar, distributing the heat to a forced-air-cooled heat sink as shown in Figure 3. The isolation resistor is a 250 watt type PPT975-250-3 50N from KDI, although a lower power type may be used.

Experimental Results

Measurement of gain and phase characteristics from the complete amplifier (including preamplifier) is shown in Figure 4. The gain decreases with higher power levels to approximately 2 dB down at 350 watts. More surprising, however, is the phase shift which is more than 40 degrees over the power dynamic range. At the output, harmonics are more than 60 dB below the carrier.

Output impedance is measured using the passive load pull technique (6) and the test setup shown in Figure 5. The combined network and spectrum analyzer HP4195A is programmed to shift between measuring and displaying the impedance and actual delivered power. For a given power level, the tuner is adjusted to deliver maximum power, and the output impedance is directly given as the complex conjugate of the load impedance in the reference plane.

Power(W)	75	200	350
Γo	$0.30\angle 110°$	$0.17\angle 210°$	$0.16\angle 150°$

Table 2. Measured output reflection coefficient at three power levels.

The directional coupler must have a high directivity in order to avoid errors. When tuning manually, it is convenient to have a tuner where reflection amplitude and phase can be set easily. It is common to use a quadrature hybrid as a phase shifter, but it can also be used as a tuner, with two different lossless loads on two ports (0 and 90 degree ports) and a load on a third (isolation port). Equation 1 shows that the input-reflection- coefficient's magnitude and angle may be adjusted by changing each output-reflection-coefficient's angle alone. This adjustment can be tricky. In Figure 6, one hybrid shifts magnitude only, using two variable capacitors in antiparallel. If one capacitor increases the reflection coefficient angle, the other one must decrease it. The angle range should be symmetric around -90 degrees so only the magnitude is shifted. In the phase shifter, the two variable capacitors are in parallel. The prototype tuner covers VSWR up to 1:3 at all angles. Table 2 shows the prototype balanced amplifier's measured output impedance as a function of power level.

Conclusion

A power amplifier designed to operate at 70 MHz in a phased array hyperthermia system is described, which features ruggedness and low cost. The amplifier is tested for stability, and the output impedance is measured using a passive load pull technique, where an easy-to-adjust tuner based on two quadrature hybrids is used. The overall amplifier efficiency could be improved using 40 volt supply and a 1:4 output transformer in each push-pull amplifier. Four complete amplifiers have been built, and they have almost the same gain and phase characteristics, primarily due to the broadband matching design. ***RF***

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3. Paul F. Combes et. al., *Microwave Components, Devices and Active Circuits*, pg. 124, John Wiley & Sons, 1987. 4. *Motorola RF Device Data* vol. 1 and 2, Motorola Inc., 1990, Sixth Edition, Rev. 4, pg. 2-1146.

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6. R. Tucher and P. D. Bradley, "Computer-aided error correction of large-signal load pull measurements", *IEEE Trans.*, Vol MTT-32, No. 3, 1984.

About the Author

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RF guide to editorial coverage

RF CAD/CAE Companies Try a Variety of Approaches

By Ann Marie Trudeau
Assistant Editor

The RF engineering field has expressed its requirements for CAD/CAE software by requesting software that provides flexibility, speed, accuracy, and easy to use interfacing. It is critical that engineers have the software reliability that allows them to speed to market a product that will actually perform as designed, which can cut back on the time-consuming task of building prototypes. Some CAD/CAE software companies have a narrow RF focus while others feel that they can also address many areas in the RF field.

Two antenna software design companies have seen the need for programs that fit between those used by the amateur radio community and expensive software that costs in the tens of thousands dollars. Their programs aid the engineer who needs to design an antenna, build it and have it work the first time according to design. Brian Beezely has based two of his four antenna programs on the mathematics of MININEC, a public domain program he has refined for easier use. Other programs use specialized models or are derived from the Numerical Electromagnetics Code (NEC). Paragon Technology, Inc,'s Vice President Todd Erdley said that his company has developed an intuitive antenna design and analysis program based on NEC, called NEC-OPT which offers optimization to the designer who is aware of basic concepts but is not well versed in RF antenna design.

Simulation Methods and Models

SPICE is a well known and powerful program that is preferred for some applications by the RF community. Its time-domain basis is very useful, but vendors don't provide designers with enough RF simulation models. "RF is a very expanding market and very few SPICE vendors are going after this market," Charles Hymowitz, Vice President of Product Development for Intusoft, said. Their SPICE program includes a special RF device library.

The inter-relationship of simulator models and designing is paramount to developing high performance products. An engineer designs a product for lower performance level if a model is not accurate. With more accurate models the designer has the option of producing a higher performance product. This also increases the likelihood that it will work correctly in the first prototype, saving time and money in development.

President David Kennedy of Optotek also notes the importance of models, pointing out that they have 172 tested simulation models in their MMICAD program.

Testing and Data

Manufacturers' data sheets don't reflect the correct RF data at higher power rates because most manufacturers don't spend the time to do large-signal modeling testing of their devices. Dr. Ulrich Rohde, President of Compact Software, said. Compact provides a fully tested model library which adds to manufacturers' data.

Another software company that is committed to testing is Eagleware Corporation. President Randy Rhea said that they get involved with their users' activities by building and testing the product using feedback to evaluate their algorithms and models for accuracy.

PC Use Increases

Nearly all the companies noted in this report provide products for the PC platfors, and some it is their primary focus.

Using the PC to address tasks basic to RF, Les Besser of Besser Associates has focused on narrow and broad band low noise RF amplifiers with the EEZ-Match and EZ-Chart programs. His self explanatory programs allow designers one key stroke or mouse operation which eliminates having to read the manuals. "It's a niche market but it's very efficient and effective in that market segment," Besser said.

Using the PC to focus on layout for RF circuits CAD Design Services has developed a program to be used within AutoCAD. The program is called Electronic Packing Design and engineers have access to automated systems which provide a net list check, parts creation, and produces a negative ground plane, according to John Sovinsky, President.

New processors like Intel's Pentium chip gives PCs the power to handle very complex RF programs. Because PCs can now handle the wide range and various types of electronic systems that HP-EEsof analyzes, Jeremy Bunting, Product Manager System, Design Tools, said that they will be releasing a program using Microsoft's upcoming Chicago (Windows 4). "We just haven't released it yet," Bunting said.

Dick Webb of Webb Laboratories said that they are releasing SYSCAD-6 which is for frequency domain based systems or receiver design. SYSCAD supports DOS now and will soon be available for Windows 4. Webb said that they feel they are continuing the right market approach and new releases won't change the company's course.

James Rautio, founder and chief scientist of Sonnet Software said that their software technique is based on FFTs, and has recently been ported from UNIX to the PC. This previously has been best applied to predominately rectangular circuits. Now the technique has been extended to include arbitrary subsection size and limitless circuit geometry. Subsections of circuits can bend and fold to match the outline of the circuit. This can result in extremely fast electromagnetic analysis of circuits with arbitrary angles and curves. They just do analysis because Rautio felt that in order to be successful in the electromagnetic area you have to be a small company and be very focused on electromagnetics.

Summary

It seems that CAD/CAE software providers are trying to put as much as they can into their programs to give engineers as much flexability and accuracy as is available, to design today's higher performance products. *RF*

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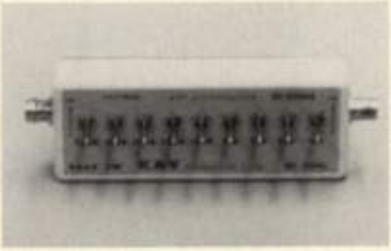
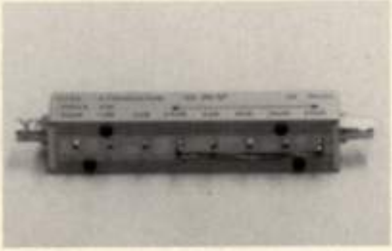


839 Manual Step Attenuator



4550 Programmable Attenuator

Manual Step Attenuators

837	50Ω	DC-1500MHz	0-102.5dB	.5dB Steps
839	50Ω	DC-2000MHz	0-101dB	1dB Steps
1/839	50Ω	DC-1000MHz	0-22.1dB	.1dB Steps
847	75Ω	DC-1000MHz	0-102.5dB	.5dB Steps
849	75Ω	DC-1500MHz	0-101dB	1dB Steps
1/849	75Ω	DC-500MHz	0-22.1dB	.1dB Steps
860	50Ω	DC-1500MHz	0-132dB	1dB Steps
865	600Ω	DC-1MHz	0-132dB	1dB Steps

Programmable Attenuators

4540	50Ω	DC-500MHz	0-130dB	10dB Steps
4550	50Ω	DC-500MHz	0-127dB	1dB Steps
1/4550	50Ω	DC-500MHz	0-16.5dB	.1dB Steps
4560	50Ω	DC-500MHz	0-31dB	1dB Steps
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RF marketplace

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REPS WANTED

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REQUEST FOR PROPOSAL

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CAREER OPPORTUNITIES

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Philips Semiconductors is dedicated to the manufacture of world-class IC products. We stand by our commitment to continuous quality improvement and customer satisfaction. Join us in the following opportunities.

PRODUCT MARKETING MANAGER

In this key position, you will define new board level wireless products and be responsible for their successful promotion. You will be responsible for advertising, tradeshows, field training, customer presentations, customer interface, pricing, and forecasting. You should have a BS/MSEE or equivalent, an MBA is preferred. Your background should include 7+ years' experience in marketing OEM data communications products.

RF SYSTEM ENGINEER/ INTEGRATOR

You will oversee all aspects of this new program from architectural definition through final product integration to ensure full compliance with FCC and industry standards and with customer requirements. You must have a BS/MSEE or equivalent with experience in technical systems integration/program management of complex system oriented products involving integration of hardware and software. Knowledge of RF, LAN and CDPD is important.

BASEBAND/MODEM ENGINEER

Take responsibility for system design and debugging of baseband modulator/demodulator and radio control circuitry on board level wireless data products in this challenging role. This will involve implementation of wireless data modem circuits using Philips IC technology and working with other Engineers in integration of a complete solution. You need 5+ years' experience in modem and digital modulator/demodulator design using state-of-the-art components and design simulation tools. Working knowledge of cellular or wireless LANs a plus.

RF SYSTEM DESIGNER

This challenging position will be responsible for system design and debugging of RF circuitry on board level wireless data products. This will involve implementation of transmitter and receiver circuits using Philips RF IC technology and working with other Engineers in the integration of a complete solution. A minimum of 5 years of experience in UHF/microwave circuit design using state-of-the-art components and design tools is required. A background in cellular or wireless LAN design and FCC type acceptance a plus.

In addition we are looking for:

- ***RF Product Marketing Engineer***
- ***RF Applications Engineer***

Philips Semiconductors offers competitive compensation and a comprehensive benefits package. Please send your resume to: Philips Semiconductors, Professional Staffing Dept. RF, MS 07, 811 East Arques Ave., Sunnyvale, CA 94088-3409, or fax to (408) 991-2656. We believe in maintaining a safe work environment for all employees and require pre-employment drug and alcohol testing. Philips Semiconductors is a division of Philips Electronics North America Corporation, and an Equal Opportunity Employer. Principals only, no phone calls, please.

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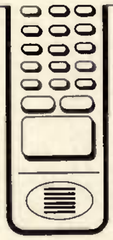

Cellular Engineers: Design/develop Rf and analog circuits for high capacity cellular systems. Requires minimum of 2 years experience in any of the following: DSP, ASIC Design, Cae Development, Digital Modulation, Digital Mobile Communications, Channel Equalizers, Transmitter-Receiver Synthesizer or Audio Design, Digital Signal Processing.

...YOUR CAREER

Cellular Amplifier Design: Design and development of HBT MMIC power amplifiers for portable cellular phone applications. Experience in design and characterization of microwave GaAs integrated circuits using advanced solid-state devices such as heterojunction bipolar transistors (HBTs) and field-effect transistors (MESFETs, HEMTs, HFETs).

Product Marketing Manager: Our client, a well known microwave firm, seeks an experienced professional to perform classical marketing functions for a product line: i.e., develop short and long term business plan, determine marketplace, etc. Excellent compensation package along with executive benefits

MMIC Design Manager: Management of an engineering group responsible for the development of non-linear GaAs MMIC products: Responsible for the management of both R&D projects for standard components as well as writing, meeting with customers, conducting design reviews, motivating employees, budgeting and bidding jobs to fixed prices. The group has successfully completed contracts with products such as mixers, converters, oscillators and frequency multipliers. These products range in operating frequencies from 200 MHz to 18 GHz.

RF/Synthesizer Design: Positions available for candidates experienced in the design and development of communication receiver/transmitter systems and subsystems. Experience in RF synthesizer design, RF circuitry, low power circuit design and GaAs/Silicon MMIC design is desirable. At least a BS degree in Engineering is required.

RFIC Design: MS or PhD in Electrical Engineering with minimum 5 years related experience preferred. The candidate should have a good knowledge and experience in Linear Bipolar High Frequency IC design and measurement techniques in design IC's like Amplifiers, Mixers, Oscillators, VCO's, Prescalers, Synthesizers, Limiting Amplifiers, etc. operating up to 2 GHz in Bipolar or BICMOS technologies.

MMIC Design Engineer: Develop L/S band GaAs MMIC power amplifiers for commercial wireless communications. Requires: M.S. or BSEE, +2 years experience with GaAs MMIC design, simulation, packaging and test.

RF Design Engineer: Responsible for the design of analog and RF systems and circuits for consumer and commercial digital wireless products. Five to ten years experience in RF systems analysis and design. Experience with low-cost techniques for frequency synthesizers, power amplifiers, up/down converters and baseband circuits for digital communications systems. Must be able to derive RF systems and module requirements to meet overall performance and cost goals. Familiarity with time division duplex or CDMA a plus.

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RF literature

Magazine's Premiere Issue

M/A-COM has released a direct marketing magazine on M/A-COM products called *LOUD & Clear* which will come out in March, June and October every year. It will include feature columns with the RF engineer in mind. They also are announcing a new Microelectronics Product Catalog which provides specifications, application notes and outline drawings for their complete line of standard, active and passive devices, components and assemblies.

M/A-COM
INFO/CARD #249

Gilbert Cell Mixer Note

Compact Software has an application note covering the design of bipolar Gilbert cell mixers using Compact's Microwave Harmonica nonlinear analysis program. Gilbert cells modulate or de-modulate RF signals and can be designed to have low current consumption while providing excellent conversion gain and good distortion characteristics, especially useful for battery operated equipment. This note describes nonlinear analysis at all stages of the design process, from the bias circuit through multi-tone analysis.

Compact Software
INFO/CARD #248

Short-Form Catalog

Murata Electronics' short form catalog shows their complete line of passive electronic components that include potentiometers, piezo alarms, resistor networks, posistors, piezoelectric ceramic filters and resonators, oscillators, custom circuit modules and micowave-related products.

Murata Electronics
INFO/CARD #247

Wireless Data Handbook

Philips Semiconductors' new 1994 RF/Wireless Handbook incorporates specific application information as well as design shortcuts for engineers. The 1100 page handbook provides technical information on Philips' RF wireless integrated circuits.

Philips Semiconductors
INFO/CARD #246

Low Power Components

RF Monolithics, Inc. offers a 10-page guide and catalog for low-power UHF components based on quartz surface acoustic wave (SAW) technology operating in the range of 224 to 928 MHz. The guide illustrates applications and worldwide frequency regulations.

RF Monolithics, Inc.
INFO/CARD #245

RF advertising index

RF software

Amplifier Design Tool

EZ-Chart™ is a low cost program to assist amplifier designers, available from Besser Associates. Without reading extensive manuals, the intuitive pull-down menus guide the novice through the design process. The program operates under DOS or as a non-Windows application. (EZ-Chart is a trademark of Aciran Software Systems).

Besser Associates
INFO/CARD #244

Filter CAD Package

PCFILT from ALK Engineering is a passive L-C and transmission line filter design and analysis package. Customers receive the software and security key, a 100-page users manual, 1 year support and updates. A demo version is available.

ALK Engineering
INFO/CARD #243

RF Cell Library

The Semiconductor Products Center of Hughes Aircraft Company has introduced its first applications specific ASIC cell library for RF communications components and systems designers. The library works with Hughes' Crescendo™ design system for integrated design of RF identification tags, electronic lot travelers, security systems and other RF applications. Features include FM modulator, PLL, class AB driver, RF amplifier, Gilbert cell, Manchester encoder and pseudo-random number generator designs.

Hughes Aircraft Company
INFO/CARD #242

DSP Software Adds Real-Time DSP Board Support

Hyperception, Inc. announces an enhancement to their Hypersignal for Windows Block Diagram simulation software. These enhancements allow DSP algorithms to be run and tested in real-time on standard DSP boards. Drivers are available for several industry-standard products.

Hyperception, Inc.
INFO/CARD #241

Software Updates

Cedars Software has updated their STRAN string analysis program to version 5.0, and have released two new programs, CPN and STAT, which provide solutions to phase noise problems. CPN and STAT implement equations presented in the book, *Phase Noise Analysis in RADAR Systems Using Personal Computers.*

Cedars Software
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Math Package

Scientist™ is a new package dedicated to experimental data fitting and simulation, using numeric and symbolic techniques.

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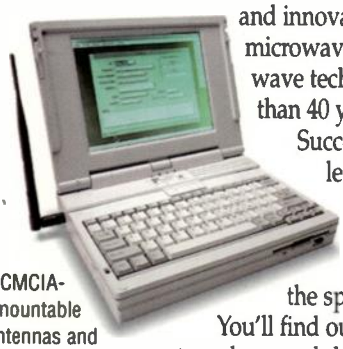

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