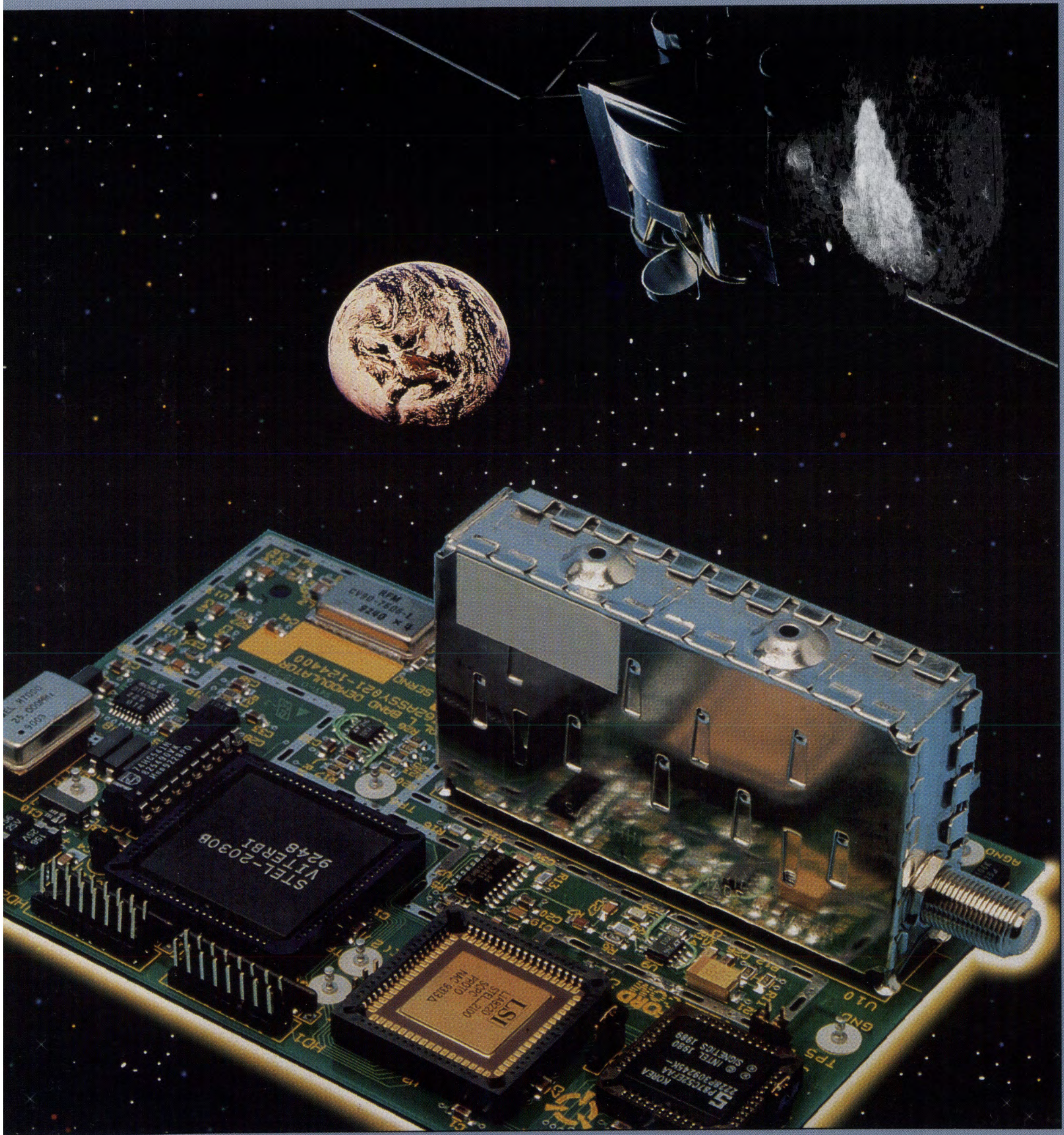
# RF design™

*engineering principles and practices*

October 1993

*Cover Story*

**Designing for RF Manufacturing**

*Plus —*

**New Products at RF Expo East**

**Official Show Issue**
**RF Expo East '9**
**preview on page**

# The Most Space Efficient Video Buffer In The Galaxy.

## At $5.75 It's Also The Most Cost-Efficient.

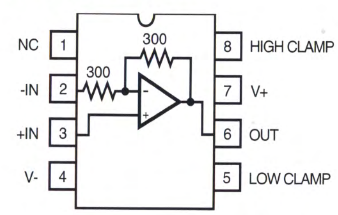
Conquer space and time with the new HFA1113 video buffer, the fastest (850MHz -3dB bandwidth) and most space-efficient buffer you can get. Free up board space with built-in output clamps and gain programming. Warp into production sooner with a standard op amp pinout that lets you drop it into an existing socket. And since the HFA1113 is priced at $5.75, you can go on to conquer another dimension: your budget. You just can't afford not to have one of these amazing devices in your video rack.

**The HFA1113 Makes The Most Out Of The Space Available.**

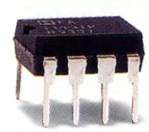
| | Harris HFA1113 | Harris HFA1110 | Analog Devices AD9630 | Comlinear CLC110 | Elantec EL2072 |
|---|---|---|---|---|---|
| Wide Bandwidth | X | X | X | X | X |
| Fast Slew Rate | X | X | X | X | X |
| Video Specs | X | X | X | | |
| Programmable Gain | X | | | | |
| Output Clamping | X | | | | |
| Pinout | *OpAmp* | *Buffer* | *Buffer* | *Buffer* | *Buffer* |
| Price (100's) | *$5.75* | *$5.95* | *$6.25* | *$5.80* | *$5.95* |

## HFA1113 Features

- Excellent gain flatness and differential phase and gain
- 850MHz -3dB bandwidth
- Programmable gain (+2, ±1) eliminates external resistors




- Programmable output clamps eliminate external diodes
- <1 ns recovery time
- 60mA output current
- 8-pin PDIP, SOIC and CERDIP packages
- Just $5.75 (100s)



For a more space-efficient design, you'll have to go back forty years. The Captain Video Survival Ring packed a compass, magnifying glass, rubber raft, and two weeks' supply of food into a module that would fit your pinkie. Today, the HFA1113 will get you out of just as many tight spots. And you won't have to send in a single box top.

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In addition to being a full-featured 22 GHz spectrum analyzer, the AN930 is a powerful measurement tool that can perform frequency or time domain analysis on the complex signals found in many modern communications systems.

A 30 MHz resolution bandwidth filter combined with a powerful set of time domain measurement functions provides the AN930 with a unique ability to capture and analyze pulsed and digital transmissions that previously has been unavailable in any other spectrum analyzer.

A built-in, sensitive FM/AM receiver with modulation measurement scales enables direct measurements of FM deviation or AM percent modulation. The accompanying high-quality audio output aids in identification of unknown or interfering signals.

Other standard AN930 features include an automatic limits test function for unattended detection of random or intermittent signals, a built-in frequency counter for precise frequency measurements, RS-232 and IEEE-488 interfaces, and operation from DC power sources.

Optional features including a built-in 2.9 GHz tracking generator, quasi-peak detector, high stability time base, and a rechargeable battery pack add even greater flexibility for solving your unique measurement requirements.

Contact IFR or your local IFR representative for more information or to arrange for a demonstration of the AN930.

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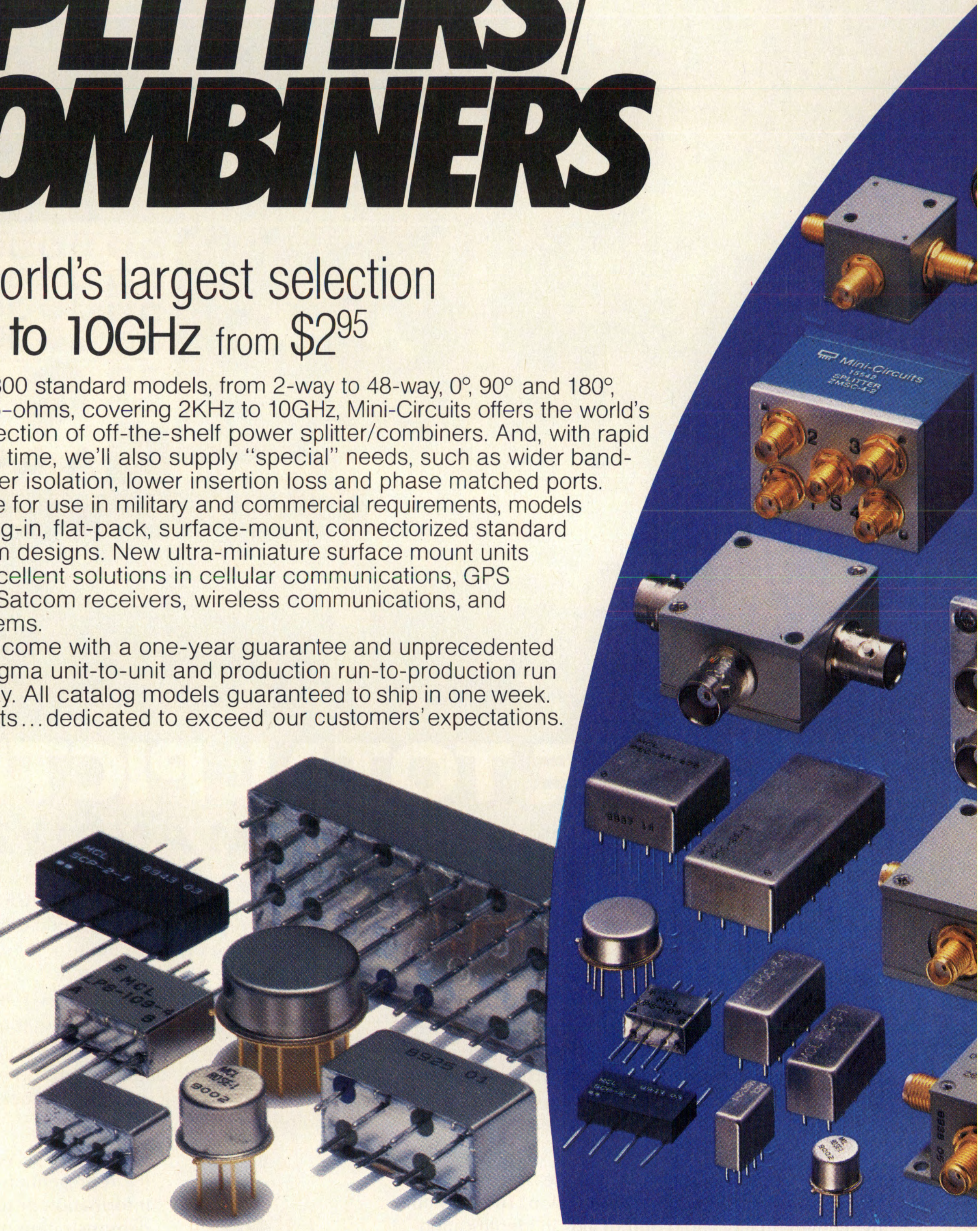

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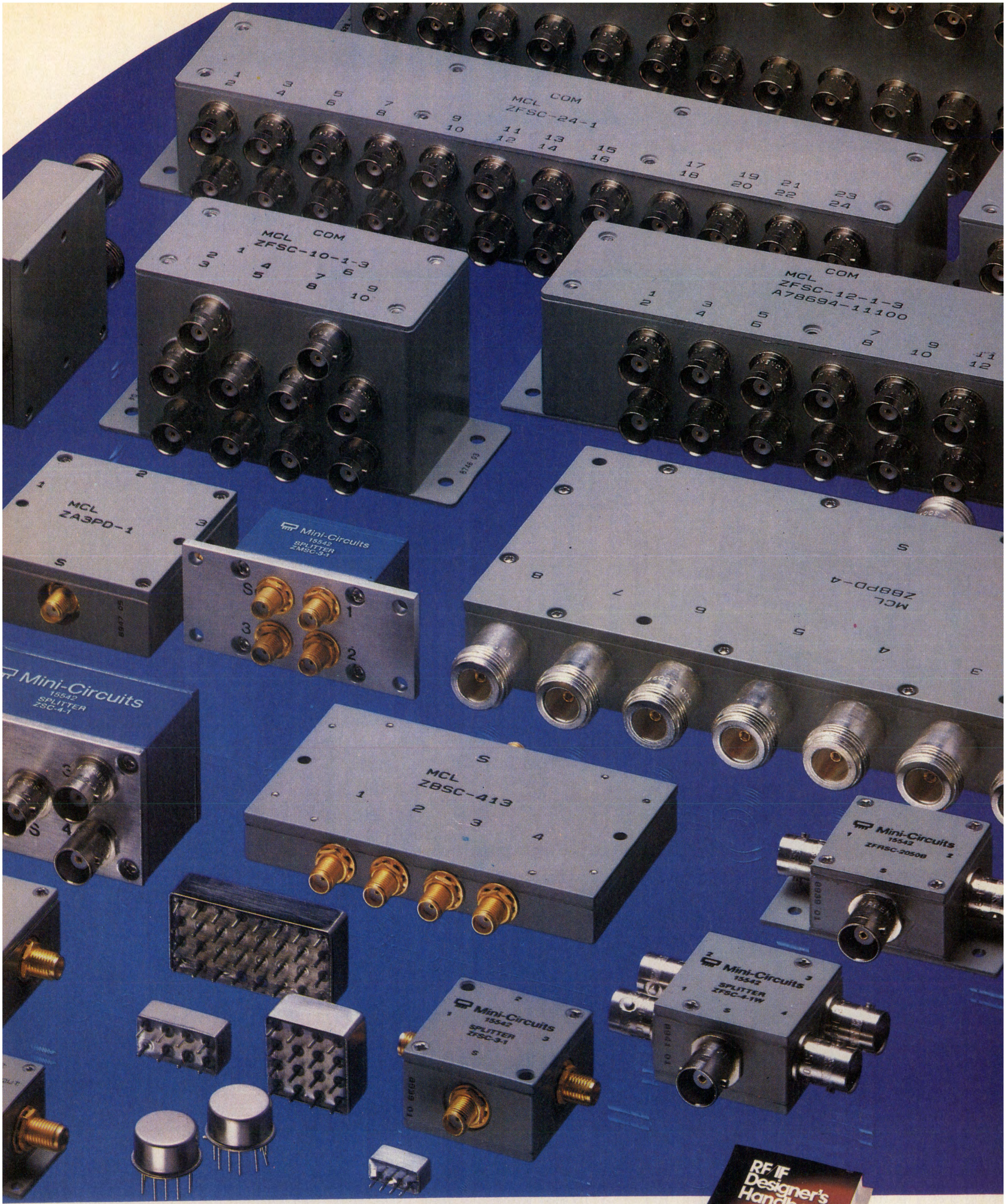
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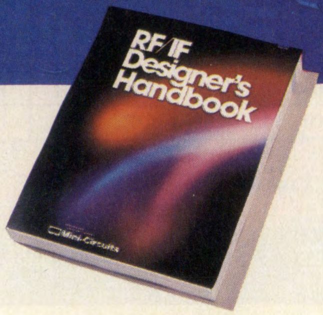
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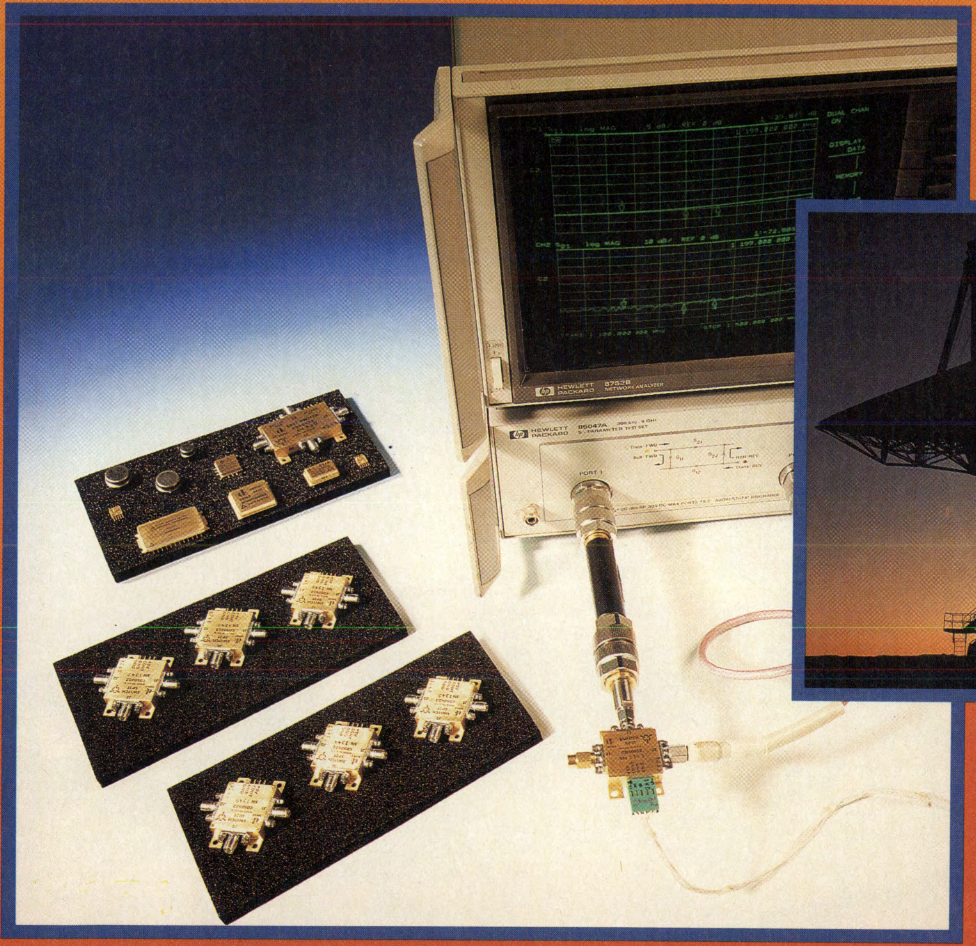

# RF design

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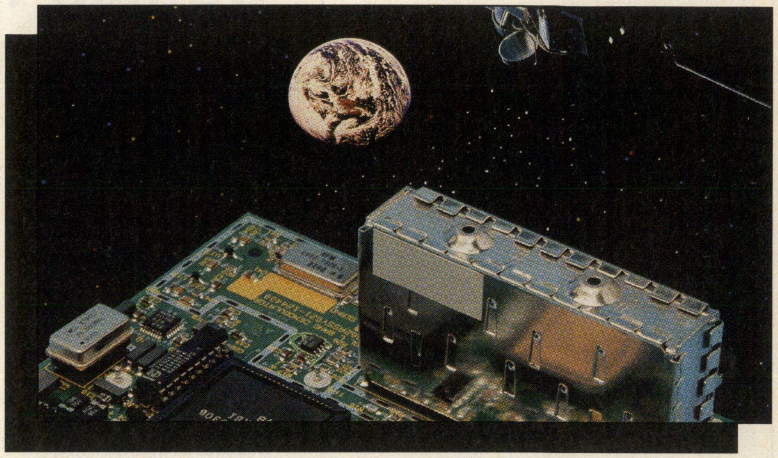

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### tutorial



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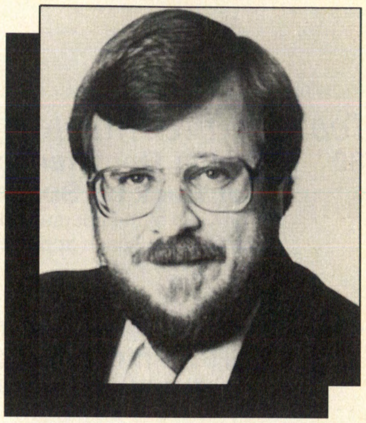
# Quality Only Comes With Commitment



*By Gary A. Breed*
*Editor*

This issue features the subject of manufacturing. Although *RF Design* is primarily known for its attention to circuit design, we chose this topic for several reasons:

Increasingly, circuit design is being formally integrated with manufacturing and test engineering. In an effort to improve quality, reduce time-to-market, and gain a pricing advantage by cutting costs, companies are instituting organized cooperative development practices.

Next, American business is giving its manufacturing methods renewed attention. Not only is the design process changing, but the whole approach to making products is under review. The driving force is fear of losing even more business to foreign companies. A reduced military market is pushing companies to re-learn consumer and commercial electronic manufacturing.

Finally, it just makes sense to talk about manufacturing with design engineers. They will create better finished products if they understand how their design choices will affect the manufacturing process. We are fortunate to have three feature articles covering different aspects of the changing manufacturing environment. The first addresses ultra-miniature packaging considerations, the next discusses ASIC design for manufacturing on silicon instead of circuit boards, and our cover story is a contract manufacturer's "wish list" of things that designers should know. Read these articles carefully; they will help you do a better job!

## The Issue of Quality

Another area of manufacturing that is getting a huge amount of attention is the implementation of quality programs. In our "Letters" column, we have a commentary that takes issue with recent developments in the ISO-9000 series of standards for certification of quality programs. Rather than discuss that letter here, read it yourself and send us your own comments on the subject.

There is one aspect of the quality issue that has not received proper attention — any program for quality control and improvement requires total commitment! It doesn't matter whether your company has ISO-9000,1,2 certification, the Malcolm Baldridge Award, SPC, TQM, zero-defects, six sigma, or any other quality program. If there is no mandate to make such a program work, it won't work.

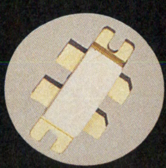
I know of one major semiconductor company that put all the pieces into place for a major quality program, then saw virtually nothing happen. After much finger-pointing and arguing about the poor performance of the program, the company president finally applied the Nike shoes philosophy — Just Do It!

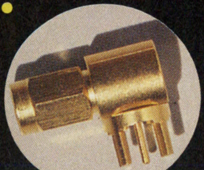
With such a clear mandate, the program finally began to work. Hard choices could finally be made on proper specifications (instead of over-stated specs), on changing well-established procedures, and on demanding that employees work harder than they have previously been required to work. Without the president's direct involvement, managers just couldn't get out of their "comfort zone" and make the necessary changes.

None of us can get too comfortable in this time of rapid technological development and uncertain economic direction. But we should remember that times of discomfort and uncertainty are often the times when we do our re-thinking and moving ahead. That's what seems to be happening today in manufacturing.

# Richardson Electronics is Your Global Partner in RF Technology

**Offering solutions to customer problems for more than 40 years, Richardson Electronics, Ltd. has grown into one of the world's premier suppliers of RF devices.**



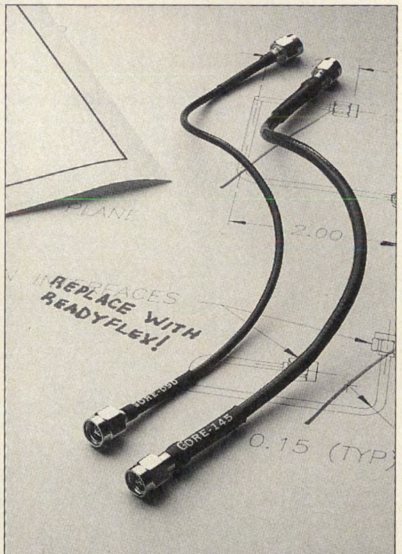


## *This Month's Feature . . .*

# Flexible and Reliable

GORE

Introducing ReadyFlex™ microwave assemblies from Gore




### A cost-effective, high performance alternative to RG & semi-rigid assemblies

A truly flexible, easy to install microwave cable assembly for use to 18 GHz is now available directly from stock.

New ReadyFlex microwave assemblies, in a choice of .090" and .145" O.D.'s, replace hard to work with semi-rigid tubing with an easy to handle high performance package. Designed with a pre-determined feature set, these highly reliable assemblies are offered at an extremely cost effective price.

**ReadyFlex assemblies offer short & long term savings**
ReadyFlex assemblies not only reduce costly installation time and failures typical of stiffer semi-rigid and less durable RG assemblies, they also eliminate expensive and time consuming 3-D drawings. This results in immediate cost savings. A single ReadyFlex assembly also replaces multiple semi-rigid configurations and lengths, further saving on inventory and paperwork.

ReadyFlex assemblies offer SMA male connectors with unique pin captivation. Bend radii as tight as 1/4", reinforced strain reliefs, operating temperatures from -55°C to 125°C, and other features have also been added to ensure long term reliability. And because they're flexible, vibration stresses don't get transferred to the connector/cable interface, further eliminating in-use failure.

**Direct from stock**
ReadyFlex assemblies are stocked in lengths of 4, 6, 9, 12, 15, 18, 24, 30, and 36 inches. Normally, within 24 hours of an easy toll free call you can have these high performing, off-the-shelf, flexible microwave assemblies on the way to your door. For as little as $68! Call now to order.

**VISA & Mastercard Accepted.**

**Guaranteed Insertion Loss (dB)**

| Length (In.) / Freq. (GHz) | 6" | 12" | 24" | 36" |
|---|---|---|---|---|
| .090" Assemblies | | | | |
| 2 | .31 | .48 | .84 | 1.19 |
| 4 | .40 | .66 | 1.16 | 1.66 |
| 8 | .56 | .92 | 1.64 | 2.36 |
| 12 | .69 | 1.14 | 2.03 | 2.92 |
| 18 | .87 | 1.42 | 2.52 | 3.62 |
| .145" Assemblies | | | | |
| 2 | .22 | .32 | .51 | .71 |
| 4 | .29 | .43 | .70 | .98 |
| 8 | .40 | .60 | .99 | 1.39 |
| 12 | .49 | .74 | 1.23 | 1.73 |
| 18 | .62 | .93 | 1.54 | 2.16 |

**Specifications**

| | |
|---|---|
| Cable O.D.: | .090", .145" |
| Frequency Range: | D.C. to 18 GHz |
| VSWR: | 1.35:1 up to 18 GHz |
| Shielding: | >90dB up to 18 GHz |
| Impedance: | 50 Ohm +/- 1 Ohm |
| Velocity of propagation: | 85% |
| Bend Radius (static): | .090" - 0.25"<br>.145" 0.5" |
| Temp. Range: | 55°C to 125°C |
| Cable MIL SPEC: | MIL-C-17 |
| Connectors: | Straight SMA to SMA Males |
| Interface: | Per MIL-C-39012 |

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Creative Technologies Worldwide

**Introducing ReadyFlex™ microwave assemblies from Gore. A cost-effective, high performance alternative to RG & semi-rigid assemblies. High performance GORE-TEX® insulation makes the difference! For off-the-shelf availability of Gore products in the U.S. and Canada, call Richardson today.**

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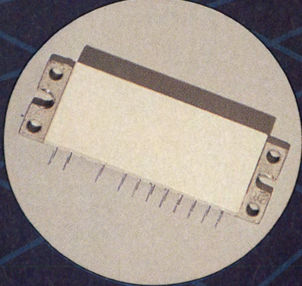

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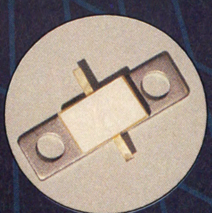

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**GIGA-TRONICS 8540 UNIVERSAL POWER METER**

# Incredibly Fast and Accurate CW *and* Peak Power Measurements At A Truly Incredible Price.

**UNIVERSAL POWER MEASUREMENT**

Incredible is credible when describing the 8540 Series of Universal Power Meters.

From Giga-tronics, the new power in power meters.

For the very first time, you can make CW *and* peak power measurements quickly and accurately with a single meter—a Universal Power Meter.

And all for about the same price you'd pay for the competitor's CW *only* power meter.


*The two-line display also lets you set the desired resolution and select either Lin or Log readout for each line.*

**POWER MEASUREMENTS INSTANTLY**

Imagine seeing display updates instantly: measurement speeds over the GPIB exceeding 200 readings per second and an exclusive Burst Mode capturing more than 2,000 readings in the same tick of a clock.

And because the 8540 Series uses diode sensors, you can measure all the way from −70 to +20 dBm with the same sensor, and without range changing delays.

If you're worried about having to write new code for your computer controlled testing, don't be: The 8540 Series uses the same GPIB command set as HP's 436A, 437B and 438A.

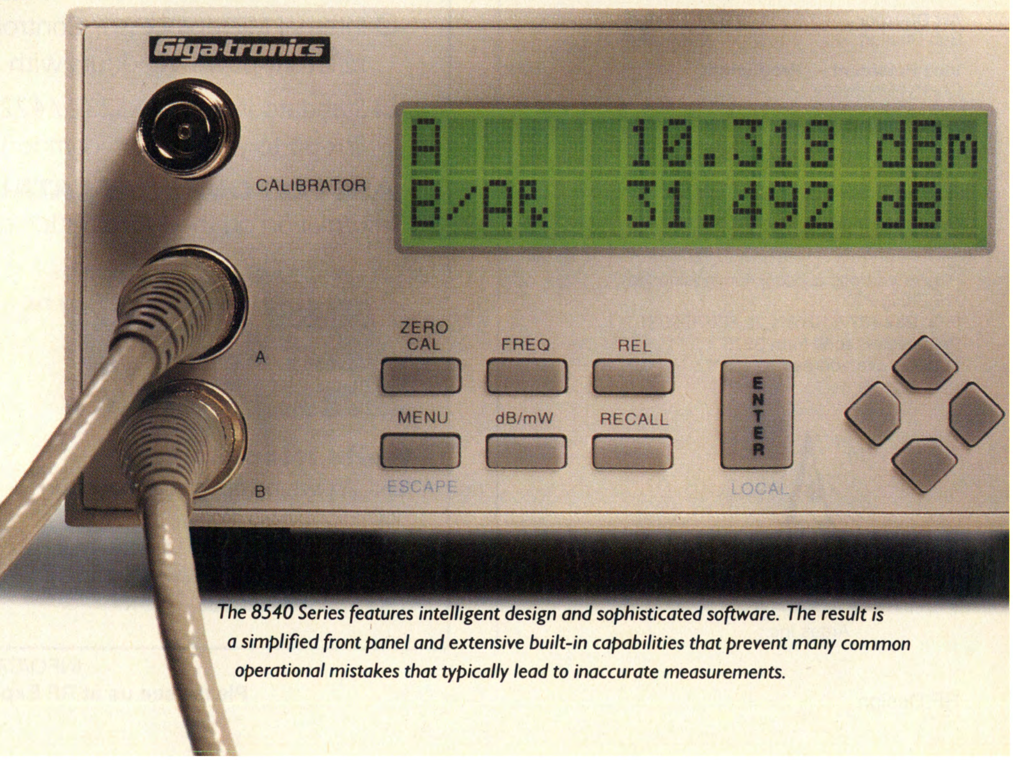
Think about what all this will do for your ATE productivity as well as for your company's bottom line.

**FAST, EASY PEAK POWER MEASUREMENT**

Now, an easy-to-use CW power meter can also measure pulsed RF signals with the simple addition of a peak power sensor.

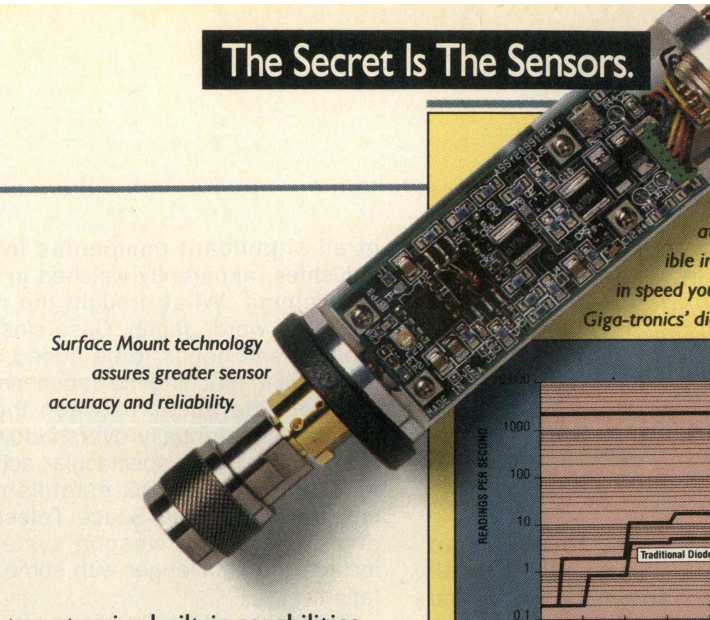
There are no time-consuming, unreliable duty cycle corrections, and you'll get the same accuracy and speed you'd get with a much-more-expensive dedicated peak power meter.

View the pulsed signal's amplitude profile on a scope and see the exact power measurement point on the pulse. Measure the overshoot. Measure the droop.



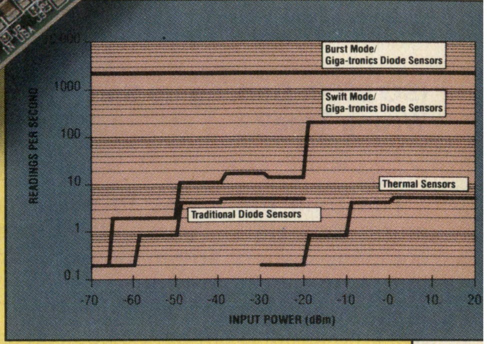

*The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.*

# The Secret Is The Sensors.



*Surface Mount technology assures greater sensor accuracy and reliability.*

*Just look at the incredible improvement in speed you'll get with Giga-tronics' diode sensors.*




The Giga-tronics 8540 Series delivers incredible performance by taking full advantage of the speed and dynamic range of diode sensors.

What's more, Giga-tronics has solved the challenge that previously limited diode sensors to the "square law" region—below −20 dBm—by utilizing a built-in power sweep calibration system. So you get speed and a full 90 dB dynamic range without sacrificing accuracy.

You'll be confident of your peak power readings, and still have all the benefits of an incredibly fast CW power meter.

**ONE OR TRUE TWO CHANNEL OPERATION**

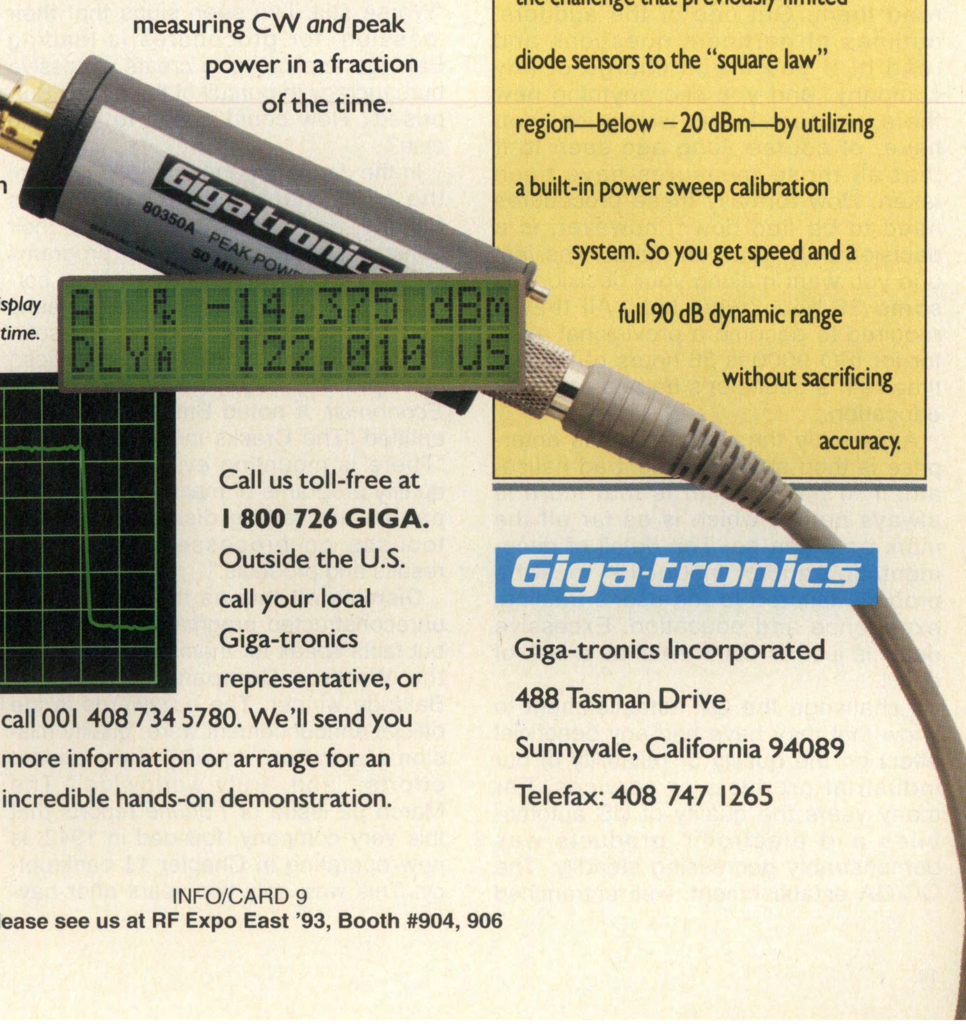
If a single-channel meter is what you need, the Model 8541 is the meter for you. But if you need two-channel capability, the Model 8542 lets you see readings from both channels *simultaneously.*

**SIMPLE, INTELLIGENT OPERATION**

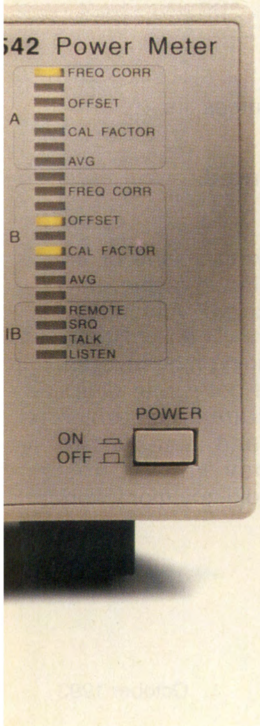
The 8540 Series has only half as many controls as other power meters, but don't let that fool you. Intelligent design and sophisticated software give you easy access to extensive built-in capabilities.

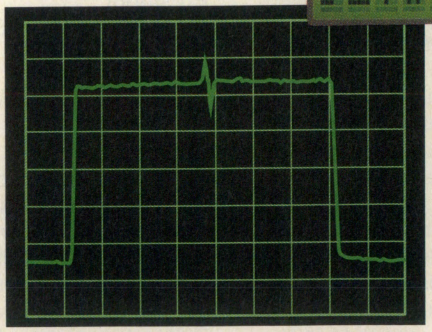
For example, you use the same key to zero and calibrate the power sensors. The meter automatically determines the function you want by detecting whether a sensor is connected to the calibrator.

Imagine all this power and performance. But why just imagine? Get the truly incredible Giga-tronics 8540 Series Universal Power Meter, and start measuring CW *and* peak power in a fraction of the time.




*A two-line back lit LCD display provides you more data in less time.*



*The peak sensor adds a marker on a monitor output for setting an exact measurement point on pulsed signals.*

Call us toll-free at **1 800 726 GIGA.** Outside the U.S. call your local Giga-tronics representative, or call 001 408 734 5780. We'll send you more information or arrange for an incredible hands-on demonstration.

**Giga-tronics**

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

## ISO 9000: Barking up the Wrong Tree?

Editor:

The time has come to sound an alarm that alerts thousands of small company owners and managers to the latest boondoggle-regulation that is about to be imposed on them. There is still time to fight, but not much. You can not open a trade magazine without reading yet another story about the blessings of ISO 9000. Don't fool yourselves, this will not just be for those who export. The QC/QA establishment will see to it that you will essentially be forced to accept this costly, ineffective system of regimentation which does absolutely nothing to improve quality, but adds new costs to your product or service.

If you do not believe what I stated, I urge you to get the ISO standards and read them. Get one of the auditors' outlines of pertinent questions and read it. If you are in charge of any company, and you see anything new there, you are in the wrong job. You have, of course, long ago seen to it that all these measures have been taken. How formally these procedures need to be tied down, however, is a decision only you can make. The last one you want making your decisions is some 36-hour "whiz kid." All that is required to become a provisional auditor for ISO 9000 is 36 hours of training time with a bachelor's degree or lesser education.

Apparently the dictum for this enterprise is thou shalt document ad nauseam. The naive belief is that more is always better, which is as far off the mark as it can be. The detail of documentation and/or instruction must be properly tailored to the user's training, experience and education. Excessive detail is just as detrimental as a lack of detail.

I challenge the QA establishment to show that they have had any beneficial effect on the quality or reliability of our industrial products or services. For many years the quality of US automobiles and electronic products was demonstrably decreasing steadily. The QC/QA establishment, well entrenched in all significant companies in both industries, apparently watched in silent amusement. What brought the reversal? One word, Japan. Case closed. A grain of salt needs to be added about Mr. Deming teaching the Japanese. The cultural differences between the two peoples substantially overshadow any such claim. Other spectacular achievements of the QA/QC establishment include the Hubble Space Telescope, and some of our weapon systems. I include the Challenger with some hesitation.

ISO 9000 and the Baldrige contest are first cousins. They share their predilection for formalization; nothing left to chance; a patent recipe designed by generalists without any specified scientific or engineering knowledge to ensure the quality of anything you can think of.

By no means is this writer alone in his assessment of ISO 9000. Management expert and author Tom Peters says, "I do know a bit about the ISO 9000 series European quality standards, which I see as substantially misguided." This is a direct quote from his periodical "On Achieving Excellence," December 1992. As to the Baldrige Award, Peters states, "Worse still, I've seen signs that their 'passion' for procedures is leading Baldrige applicants to create excessive bureaucracy in pursuit of the prize." Surprised? How could it lead to anything else?

In the January/February 1992 issue of the *HBR*, management consultants Schaffer and Thomson summarize their objection to activity-centered programs such as TQM, etc. this way. "Most corporate change programs mistake means for ends, process for outcome. The solution is to focus on results, not activities." On April 18, 1992, an article in *The Economist*, a noted British publication, entitled "The Cracks in Quality" states, "There is mounting evidence that the quality programs of many Western companies are failing dismally. . . TQM focuses on processes rather than results and products."

Dismiss all this as the polemics of unreconstructed pragmatists if you will, but facts speak for themselves. In 1990, the Wallace Company, Inc. was a Baldrige winner. The buzzwords in the official announcement were "quality mission statement," "quality improvement efforts," and "truly worldwide." The March 92 issue of *Fortune* reports that this very company, founded in 1942, is now operating in Chapter 11 bankruptcy. This was only two years after having demonstrated to the Baldrige auditors ". . .the company's strategic planning process for the short term (1-2 years) and longer term (3 years or more) for customer satisfaction, leadership and overall operational performance improvement."

What is my motivation for this severe censure of both Baldrige and ISO 9000, when my own company could most likely successfully participate or implement these "rain dance" schemes? It is the solid conviction that our nation's need to get back its traditional competitive edge is severely jeopardized by this empty symbolism, which totally lacks the time-honored problem-targeted pragmatic approach.

It is a fair question to ask, what credentials the writer possesses regarding quality products. I have a Dipl. Ing. degree in E.E. and a half century of international experience in the electronic industry. As founder and president of PTS, I believe we have something to say about quality products and service. We export 30% of our products to Europe, and all of these carry UL, CSA and VDE certifications. We have won blue-chip companies' quality awards several years running and long ago polled all our major customers in our own formal quality survey, again earning only the highest marks. Our products have demonstrably met their calculated MTBF (MIL 217) of 20-40,000 hours for many years. We are an employee-owned and employee-motivated company, profitable managed for each year of its 18-year history.

George H. Lohrer
President
Programmed Test Sources, inc.

## "Almond" Info Wanted

Editor:

I was recently looking at one of the software programs offered in the early days of *RF Design* called "Almond" by Chris Trask. This is really an outstanding program for designing high frequency circuits, but it will not run in Quick Basic because of statements which must be changed, and it has a few bugs. Does anybody know Chris' address or if he ever did any more work with it?

Jon GrosJean
(203) 974-0677

# *RF calendar*

## October

**19-21** **RF Expo East**
Tampa, FL
**Information: Renae Fierros, Cardiff Publishing, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 770-0253.**

**19-21** **Eighth Annual Minnesota EMC Event**
Bloomington, MN
Information: Kim Valleen, AMADOR Product Service, 1775 Old Highway 8, New Brighton, MN 55112-1891. Tel: (612) 631-2487. Fax: (612) 631-3515.

**19-21** **Scan-Tech '93 International Show and Seminar**
Philadelphia, PA
Information: AIM USA, 634 Alpha Drive, Pittsburgh, PA 15238. Tel: (800) 227-0206, (412) 963-8588. Fax: (412) 963-8753.

**26-28** **Nepcon Southeast '93**
Orlando, FL
Information: Nepcon Southeast '93, P.O. Box 465, Brookfield, IL 60513-0465. Tel: (708) 390-2420. Fax: (708) 345-6278.

**26-30** **Second International Conference on Signal Processing**
Beijing, P.R. China
Information: Prof. Yan BAOZONG, Institute of Information Science, Northern Jiaotong University, Beijing 100044, China.

## November

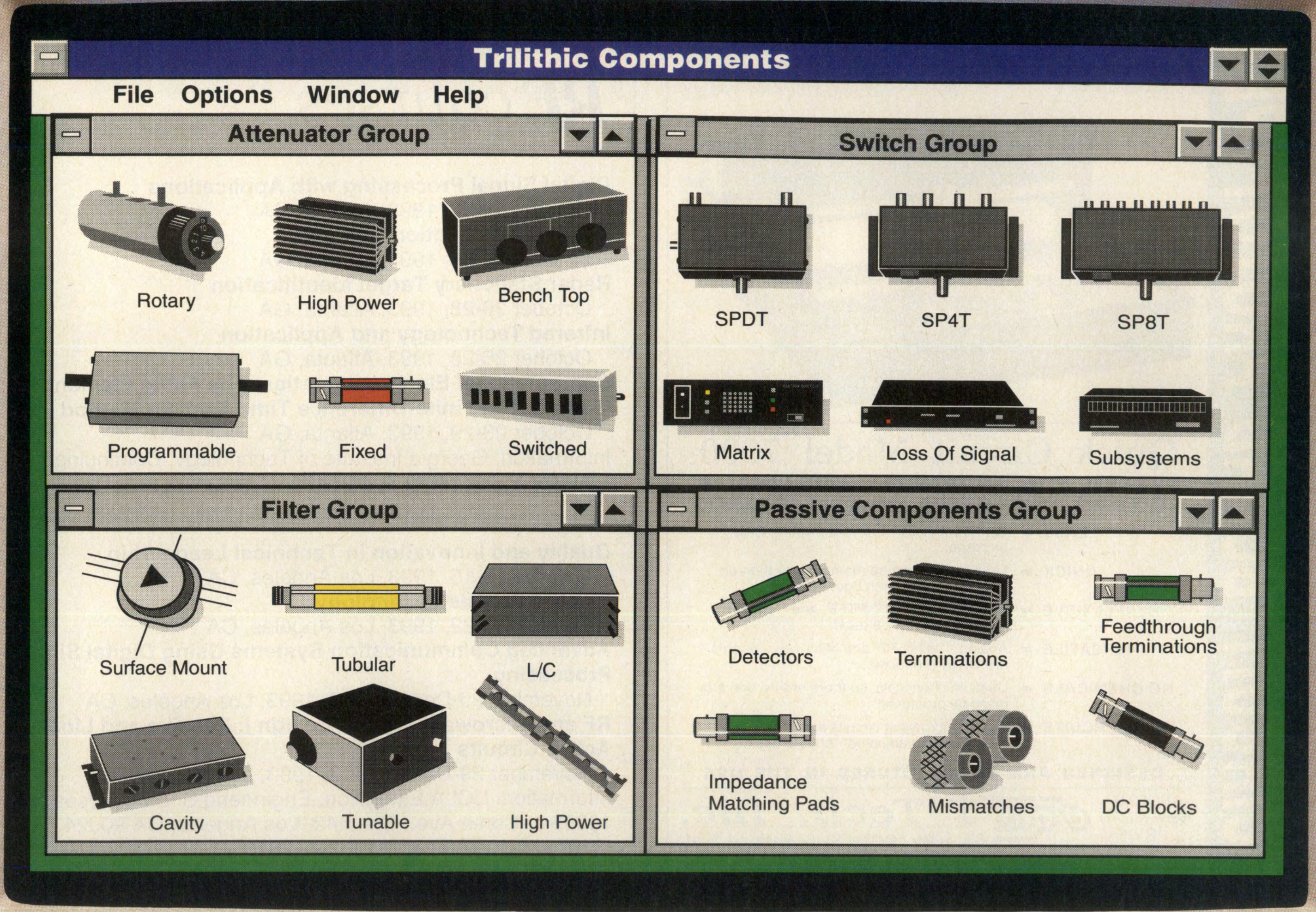
**1-4** **EuroComNet/Amsterdam 1993**
Amsterdam, Netherlands
Information: TWI, International Exhibition Logistics, 3190 Clearview Way, San Mateo, CA 94402. Tel: (415) 573-6900. Fax: (415) 573-1727.

**2-4** **Third European Conference on Satellite Communications**
Manchester, UK
Information: Miss J.A. Gordon, ECSC 3, Conference Services, IEE, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071 344 5477. Fax: 071 497 3633.

**3-5** **Second International Conference on Broadband Services, Systems and Networks**
Brighton, UK
Information: Conference Services, IEE, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071 344 5477. Fax: 071 497 3633.

**9-11** **2nd International Conference on Multichip Modules, ISHM '93**
Dallas, TX
Information: ISHM, 1861 Wiehle Ave., Suite 260, Reston, VA 22090. Tel: (703) 471-0066. Fax: (703) 471-1937.

**18-21** **Expotelecom 1993 - Exhibition for Telecommunications and Electronic Services**
Lisbon Protugal
Information: TWI, International Exhibition Logistics, 3190 Clearview Way, San Mateo, CA 94402. Tel: (415) 573-6900. Fax: (415) 573-1727.

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**Please see us at RF Expo East '93, Booth #704**

# RF courses

**Digital Signal Processing with Applications**
October 18-22, 1993, Atlanta, GA
**Radar Cross Section Reduction**
October 26-29, 1993, Atlanta, GA
**Radar Stationary Target Identification**
October 26-28, 1993, Atlanta, GA
**Infrared Technology and Application**
October 26-28, 1993, Atlanta, GA
**Computational Electromagnetics: The Finite Element Method, The Finite-Difference Time-Domain Method**
October 26-29, 1993, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894-2547.

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**Quality and Innovation in Technical Leadership**
October 13-15, 1993, Los Angeles, CA
**Electric Vehicle Technology**
October 18-22, 1993, Los Angeles, CA
**Advanced Communication Systems Using Digital Signal Processing**
November 29-December 3, 1993, Los Angeles, CA
**RF and Microwave Circuit Design I: Passive and Linear Active Circuits**
November 29-December 3, 1993, Los Angeles, CA
Information: UCLA Extension, Engineerig Short Courses, 10995 LeConte Ave., Ste. 542, Los Angeles, CA 90024. Tel: (310) 825-1047. Fax: (310) 206-2815.

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**Radar Fundamentals with Emphasis on Airborne Applications**
October 25-19, 1993, Washington, DC
**Low Earth Orbit Satellite Systems**
November 1-3, 1993, Washington, DC
**Electromagnetic Interference and Control**
November 1-5, 1993, Washington, DC
**Advanced Signal Processing**
November 8-12, 1993, Washington, DC
**Satellite Microwave Remote Sensing and Applications**
November 8-10, 1993, Washington, DC
**Mobile Cellular Telecommunication Systems**
November 15-17, 1993, San Diego, CA
**Microwave System Engineering**
November 15-19, 1993, San Diego, CA
**Synthetic Aperture Radar with Remote-Sensing Applications**
November 15-19, 1993, Washington, DC
**Modern Radar System Analysis**
November 15-19, 1993, San Diego, CA
**Digital Data Communication Over a Multipath Fading Link**
November 16-18, 1993, San Diego, CA
**Grounding, Bonding, Shielding, and Transient Protection**
November 16-19, 1993, San Diego, CA
**Electronic Intelligence: Analyzing Radar Signals**
November 30-December 2, 1993, Washington, DC
**Digital Transmission Systems**
December 6-9, 1993, Washington, DC
**New HF Communication Technology: Advanced Techniques**
December 6-10, 1993, Washington, DC
**Optical-Fiber Communications**
December 6-10, 1993, Washington, DC
**Analyzing Communication System Performance**
December 13-15, 1993, Orlando, FL

**Modern Digital Modulation Techniques**
December 13-17, 1993, Orlando, FL
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

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**Modern RF & Microwave Techniques**
October 26-29, 1993, Monterey, CA
Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995-6335. Fax: (818) 995-2932.

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**Management of Electromagnetic Energy Hazards**
October 12-14, 1993, New Brunswick, NJ
Information: Cook College Office of Continuing Professional Education. Tel: (908) 932-9271.

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**RF and Microwave Circuit Design: Linear and Non-Linear (Theory and Applications)**
November 15-19, 1993, Cambridge, UK
**Modern Digital Modulation Techniques**
November 17-19, 1993, Cambridge, UK
**Far-Field, Compact & Near-Field Antenna Measurement Techniques**
March 21-24, 1994, Switzerland
**Aspects of Modern Military and Commercial Radar**
March 21-25, 1994, Switzerland
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122-175-70. Fax: (46) 122-143-47.

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**Workshop on Information Technology Equipment**
November 16-17, 1993, San Diego, CA
**Globalability: The Key to International Compliance**
November 18-19, 1993, San Diego, CA
Information: Underwriters Laboratories Inc., P.O. Box 1385, Northbrook, IL 60065. Tel: (708) 272-8800.

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**RF and Microwave Device Test for the '90s**
October 19, 1993, Portland, OR
November 9, 1993, Pittsburgh, PA
November 11, 1993, Cincinnati, OH
November 16, 1993, Cleveland, OH
November 18, 1993, Rochester, NY
November 19, 1993, Syracuse, NY
November 30, 1993, St. Paul, MN
December 2, 1993, Rolling Meadows, IL
December 3, 1993, Ft. Wayne, IN
Information: Hewlett-Packard Company, Microwave Instruments Division (MID), 1400 Fountaingrove Parkway, Santa Rosa, CA 95403-1799. Tel: (800) 765-9200.

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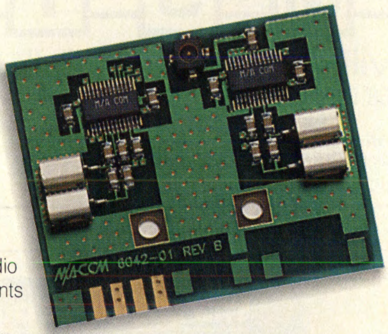
**RF/MW Small Signal/Low Noise Amplifier Design**
November 8-9, 1993, Indianapolis, IN
**RF/MW Circuit Design I**
November 29-December 3, 1993, Los Angeles, CA
Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel: (415) 949-3300. Fax: (415) 949-4400.

# We're bridging the gap between

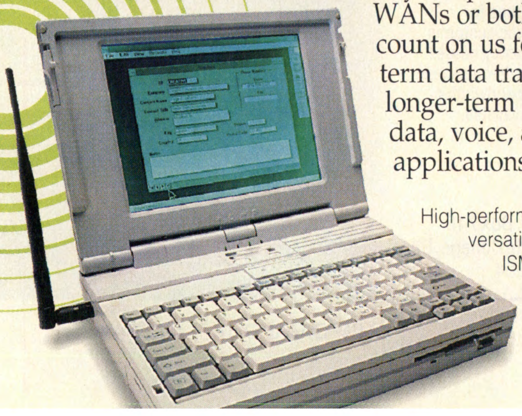


2.4 GHz surface mount radio components and connectors.

Relax. Your wireless data transmission worries are over.

At M/A-COM, we have the microwave expertise to transmit your ones and zeros across the spectrum and ensure that they're received as ones and zeros. Easily and effectively. We've been leaders in RF, microwave and millimeter wave technology for more than 40 years, and today we're the largest independent company in the business.

Our digital signal processing knowledge allows us to transparently link microwave technology to the digital world. We supply connectivity solutions for anything from pen-based computers to bar code readers to personal communicators. Whether they're operating on LANs, WANs or both. You can count on us for both near-term data transmission, and longer-term integrated data, voice, and video applications.



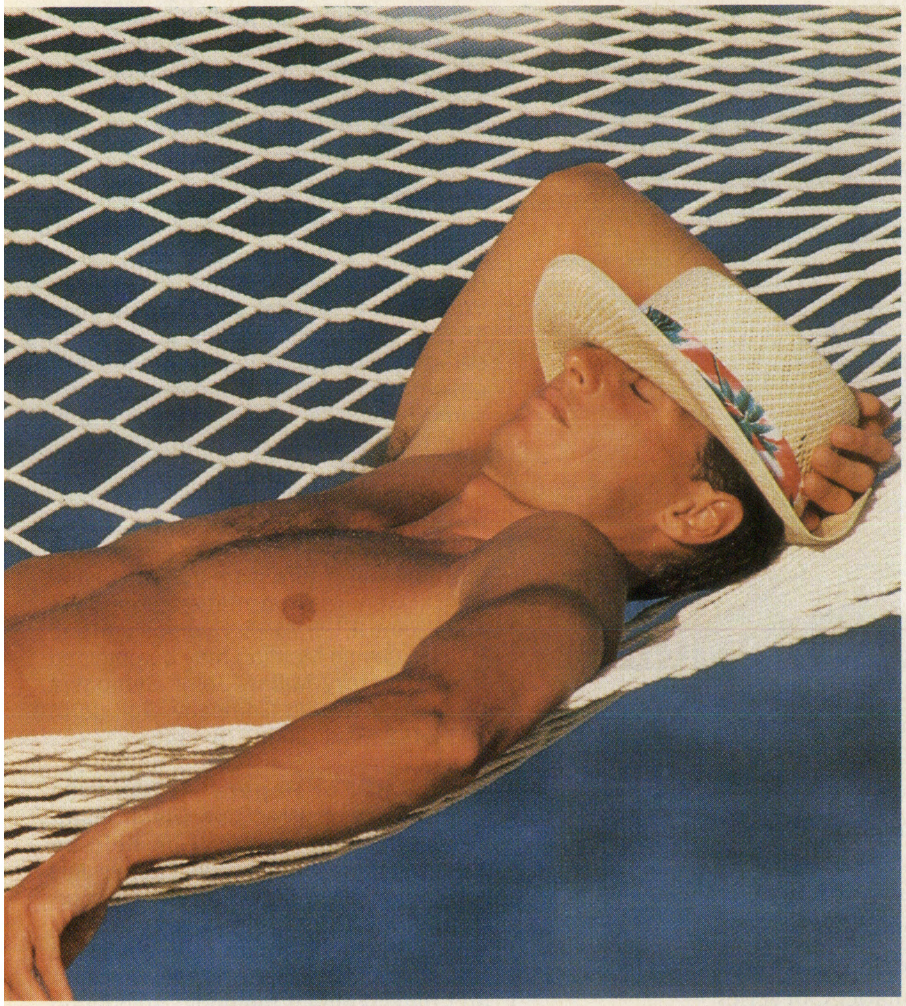
High-performance, low-cost, versatile antennas for ISM at 902 MHz and 2.4 GHz, and PCN at 1.9 GHz.

**We'll help you transmit bits. And sound bites.**

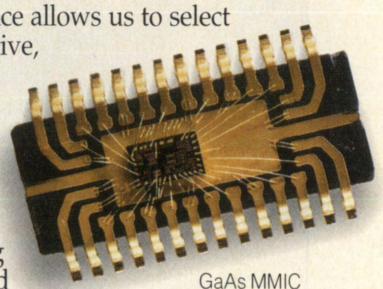
We manufacture wireless antennas, RF connectors, discrete semiconductors, transmitters, and integrated circuits, including the MMICs, HMICs™, and GMICs™ that higher frequencies require. All of which are ever smaller, lighter and more energy efficient.

But we offer the wireless data marketplace something that's more important than discrete devices for PC cards. We offer the ability to work with you to create strategic partnerships that encompass planning, manufacturing and ongoing research and development. Ultimately, these partnerships will make worldwide wireless data communications successful.

# communications RF and digital.



materials science allows us to select the most effective, economical materials for each application. We design for maximum manufacturing advantage, and use techniques like process-oriented and continuous-flow production. We drive costs down and quality up.



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**Wireless is not the wave of the future.**

It's here now. It's helping businesses control inventory, trade commodities, make airline reservations, track sales. It's getting faster, going further, and growing more important every day.

As wireless data applications increase, the interface between microwave and digital becomes increasingly important. By working together to link these two worlds, we'll create the integrated, multifunction solutions which will become the industry standards. Together, we'll bridge the gap. Then we can both relax.

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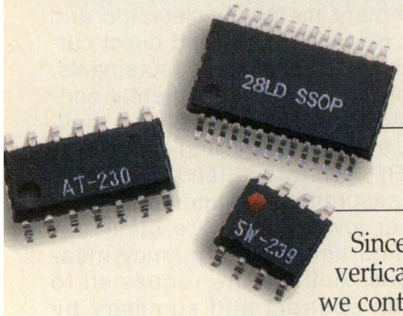


Low-cost, high-performance GaAs and Silicon integrated circuits.

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We use physics-based predictive modeling to make sure designs work as well in your PCs as on our Computer-Aided Design workstations. Our expertise in

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INFO/CARD 17

**Please see us at RF Expo East '93, Booth #604, 606**

## HP to Acquire EEsof

Hewlett-Packard Company and EEsof Incorporated announced that they have signed a definitive agreement for HP to acquire EEsof. Upon completion of this acquisition, EEsof will report to HP's High Frequency Design Software Operation, which is part of the HP Santa Rosa Systems Division in Santa Rosa, California. The new organization will be called HP EEsof. EEsof develops computer-aided engineering (CAE) software used to design high-frequency systems, circuits and devices.Their strengths are in communications system simulation and support of design software on Windows- and DOS-based PC platforms. Hewlett-Packard Company is an international manufacturer of measurement and computation products and systems. HP's strengths are in electromagnetics and instrument automation. HP and EEsof market a range of simulation and analysis tools to customers worldwide, providing design solutions for applications including cellular telephones, wireless LANs, collision-avoidance radar, defense electronics and satellite communications links. Both companies have experience in high-frequency linear and nonlinear circuit simulation, framework technology on UNIX® system-based platforms and device characterization. The proposed acquisition is subject to receipt of the necessary government approval.

**Final Call For Papers** — The 1994 IEEE AP-S International Symposium and URSI Radio Science Meeting will be held on the campus of the University of Washington, Seattle, Washington, June 19-24, 1994. The symposium is sponsored by the IEEE Antennas and Propagation Society and the meeting is sponsored by USNC Commissions A,B,D,E, and K of the International Union of Radio Science. The technical sessions, workshops, and short courses will cover the five-day period of June 20-24. Authors are invited to submit papers on all topics of interest to the AP-S and URSI. General information about the 1994 joint IEEE AP-S Symposium and Radio Science Meeting may be obtained from Dr. Gary Miller, Joint Symposia Chair, The Boeing Company, Tel: (206) 773-3482, Fax: (206) 773-4946.

**First Call For Papers** — The 1994 Conference on Precision Electromagnetic Measurements will be held on Monday, June 27, through Friday, July 1, 1994, in Boulder, Colorado. The purpose of the biennial meetings of CPEM is to exchange information on advanced instrumentation, including new sensors and measurement methods; automated measurement methods; dielectric and antenna measurements; and direct current and low-frequency measurements. Other topics include fundamental constants and special standards; laser, optical fiber, and optical electronic measurements; RF, microwave, and millimeter-wave measurements; superconducting and other low-temperature measurements; and time and frequency measurements. Authors are requested to submit an abstract and summary by January 18, 1994. For further information, contact Gwen E. Bennett, Conf. Secretary. Tel: (01) 303-497-3295, Fax: (01) 303-497-6421.

# 29 Hot New Ways to Build Really Cool Power Amplifiers

High efficiency. And low thermal resistance. Together, they keep our new Power FETs running cool — and that can help put your amplifier reliability problems ON ICE.

Our *D Series* devices are ideal for C-Band analog applications, while our *DD* parts *guarantee* outstanding IM3 performance and flat gain in your digital and mixed-signal circuits.

For space applications, our "C" Series FETs deliver Power Added Efficiencies (PAE) of 56%.

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Circle the number below, or call your nearest CEL Sales Office. We'll send you out data sheets and a Product Selection Guide immediately.

| Frequency | Part No. | P1dB | GL(dB) | PAE |
|---|---|---|---|---|
| 3.7 to 4.2 for Space | NEZ3742-10C | 10W | 15.0 | 56% |
| | NEZ3742-5C | 5W | 15.0 | 56% |
| | NEZ3742-2C | 2W | 15.0 | 56% |
| 4.4 to 5.1 | NEZ4450-15D, DD | 18W | 10.0 | 35% |
| | NEZ4450-8D, DD | 9W | 10.5 | 37% |
| | NEZ4450-4D, DD | 4.5W | 10.5 | 40% |
| 5.9 to 6.4 | NEZ5964-15D, DD | 18W | 9.0 | 33% |
| | NEZ5964-8D, DD | 9W | 9.5 | 35% |
| | NEZ5964-4D, DD | 4.5W | 10.0 | 37% |
| 6.4 to 7.2 | NEZ6472-15D, DD | 18W | 8.0 | 31% |
| | NEZ6472-8D, DD | 9W | 8.5 | 33% |
| | NEZ6472-4D, DD | 4.5W | 9.0 | 35% |
| 7.1 to 7.7 | NEZ7177-8D, DD | 9W | 8.0 | 31% |
| | NEZ7177-4D, DD | 4.5W | 8.5 | 33% |
| 7.7 to 8.5 | NEZ7785-8D, DD | 9W | 7.5 | 31% |
| | NEZ7785-4D, DD | 4.5W | 8.0 | 33% |

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INFO/CARD 19

**Please see us at RF Expo East '93, Booth #308**

**TDAB Holds First Meeting** — Telecommunications representatives met in Geneva on the 6th and 7th of July, 1993 to set up a strategic consultative body called Telecommunication Development Advisory Board (TDAB). The board will advise ITU on priorities and strategies for telecommunication development, to advise ITU Member countries on how best to step up telecommunications development and to reinforce the role of the development machinery of the Union in this area. TDAB is expected to draw upon the resources and experiences of both governments and the private sector.

**Open Network Standards** — Motorola has submitted a recommendation to the Telecommunications Industry Association (TIA), for a cellular standard that will encourage competition in the Pan American cellular market. The recommendation will establish a common approach to connecting cellular switches to cellular base stations. The recommended standard is referred to as the "A+ interface" and resembles the "A interface" which has been deployed by a number of the world's switch and cell site vendors.



**Proposed Mexican Production Facility** — RF Industries, in partnership with a coax connector manufacturer in Taiwan, will shortly be forming a joint venture manufacturing operation in Tijuana and Monterrey, Mexico. The plant will mold, plate and assemble the company's products. For more information, contact RF Industries, Ltd. at their new San Diego location. Their address is 7620 Miramar Road, San Diego, CA 92126-4202, Tel: (619) 549-6340 - Fax: (619) 549-6345.

**New Identity For Differential GPS Service** — Magnavox Electronic Systems Company and CUE Network Corporation unveiled a new name and identity program for their nationwide precise satellite positioning and location service. The service, formerly called "Pinpoint," will be marketed jointly by the two companies throughout North America under the brand name ACC•Q•POINT. The new name was chosen to clear up possible market confusion with other companies using the generic name "Pinpoint." ACC•Q•POINT is a real-time differential GPS (DGPS) system which permits a subscriber to determine geographic location within a few feet, using signals from U.S. navigational satellites and error correction data broadcast via subcarriers from commercial FM radio stations.

**Iridium Finds Financing** — Motorola announced it has signed more than two dozen investors and secured $800 million in financing for it's 6-year old Iridium global wireless communications project. The investments in Iridium Inc. will permit the project to move from the R&D phase to the manufacturing phase. Before the project is finished, a second round of financing will likely be needed to secure another $800 million. By 1998, Motorola and the other investors, plan to have a system of 66 satellites circling the Earth.

**Expansion For Taconic Plastics** — Taconic Plastics, manufacturer of

# For ten years we've been supplying our NE710 customers with the industry's leading FET.

# What more could they ask for?

***"My needs are simple: A FET that delivers absolutely the best noise performance available. Period."***

Introducing the new extra-low noise Pseudomorphic Heterojunction

NEW

| | NF | Ga | P1dB | Freq |
|---|---|---|---|---|
| NE326 | 0.5 | 11.5 | 10.75 | 12 |

*In chip and low cost ceramic and hermetic packages*

***"How about a chip that does it all? Good noise figures, good gain, moderate output power, and bandwidth enough for my 6 - 18 GHz amplifiers."***

You've got it. Introducing the new Heterojunction version:

NEW

| | NF | Ga | P1dB | Freq |
|---|---|---|---|---|
| NE332 | 0.8 | 10.5 | 12.5 | 12 |

*In chip and low cost ceramic or hermetic packages*

***"How about a low cost plastic version? One I can order in volume for applications to 14 GHz."***

Our high-yield Ion Implant process produces a consistant, uniform MESFET at a new low cost.

| | NF | Ga | P1dB | Freq |
|---|---|---|---|---|
| NE760 | 0.8 | 15.5 | 14.5 | 4 |

*In plastic, low cost ceramic and hermetic packages and chips*

NOTE:
Noise Figures (NF) in dB
Gain (Ga) in dB
Power Out (P1dB) in dBM
Frequency in GHz

***"How about a part with Heterojunction-style performance for my DBS, MMDS and commercial communications applications—at a MESFET price."***

Meet the ideal second stage device:

NEW

| | NF | Ga | P1dB | Freq |
|---|---|---|---|---|
| NE424 | 0.8 | 11.5 | 11.0 | 12 |

*In low cost ceramic packages*

***"I'm building mixers at 900 MHz and need dual gate MESFETs by the thousands. Can you fix me up—at under 50¢ per part?"***

We can—two different ways:

| | NF | Ga | P1dB | Freq |
|---|---|---|---|---|
| NE251 | 0.8 | 12.0 | 12.0 | 0.9 |
| NE253 | 0.8 | 12.0 | 15.0 | 0.9 |

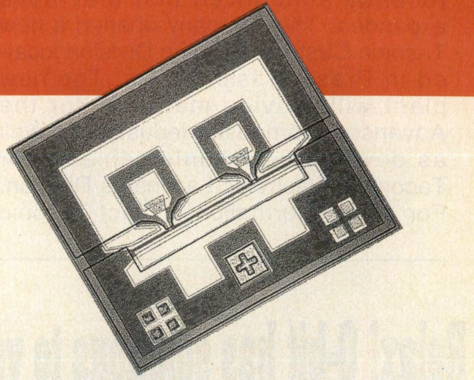
*In low cost plastic packages*

***"I need a <u>low</u> noise FET for the LNA in my TVRO system. And, of course, I need it in a high volume, tape and reel package."***

How about a 0.6 dB noise figure at 12 GHz? Introducing the new:

| | NF | Ga | P1dB | Freq |
|---|---|---|---|---|
| NE324 | 0.6 | 11.5 | 11.0 | 12 |

*In chip and low cost ceramic and hermetic packages*

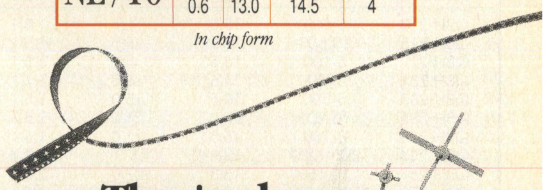


***"Don't mess with the original! It's designed into more of my circuits than any other FET. The 710 is a workhorse; solid, consistent–and I still use 'em by the truckload. Don't change!"***

Here you go:

| | NF | Ga | P1dB | Freq |
|---|---|---|---|---|
| NE710 | 0.6 | 13.0 | 14.5 | 4 |

*In chip form*



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INFO/CARD 21

**Please see us at RF Expo East '93, Booth #308**

microwave circuit board materials, recently announced that they have expanded. The company opened a new Taconic Custom Weaving Division located in Presque Isle, Maine. The new plant will provide materials for the Advanced Composite industry, as well as develop new reinforcements for Taconic's Electronic Laminates Division. For more information contact Taconic Plastics, Ltd. P.O. Box 69, Coonbrook Road, Petersburgh, New York, 12138, Tel: (800) 833-1805 or (518) 658-3202 Fax: (800) 272-2503 or (518) 658-3204.

**Celeritek Relocates** — Celeritek, Inc. has relocated the company from San Jose, California to larger facilities in Santa Clara, California. The company manufactures wireless communications products for commercial and military applications. The new facility will provide expanded capacity for research and development, equipment manufacturing and semiconductor operations. For more information contact Robert Jones, Vice President, Marketing, 3236 Scott Boulevard, Santa Clara, Calif. 95051, Tel: (408) 986-5060, Fax: (408) 986-5095.

**New Company Name** — John Fluke Mfg. Co., Inc. has changed its corporate name to Fluke Corporation. In addition, the company has changed its stock exchange symbol from "FKM" to "FLK." The stock, traded on the American and Pacific Stock Exchanges, has been trading under the new symbol "FLK" since August 13, 1993.

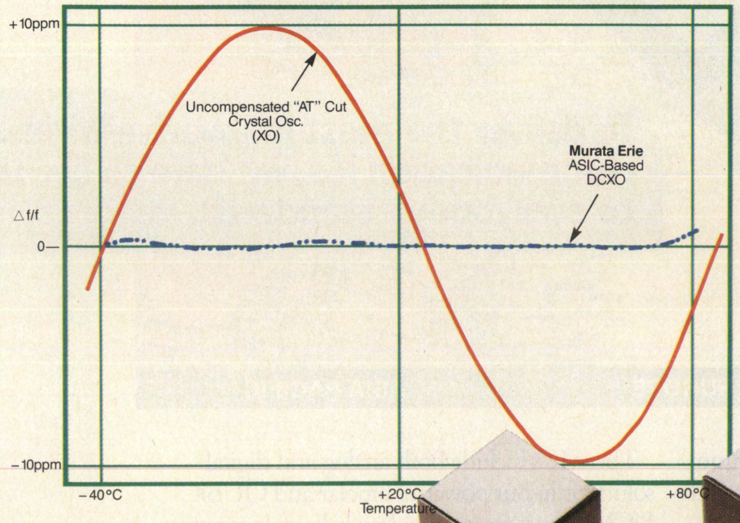
**Expanded Facility** — Hittite Microwave Corporation has expanded its facility and is now occupying the entire building at its current location of 21 Cabot Road, Woburn, Massachusetts. Hittite Microwave Corporation is a supplier of microwave monolithic integrated circuits (MMIC) and multi-chip modules made of MMIC chips.

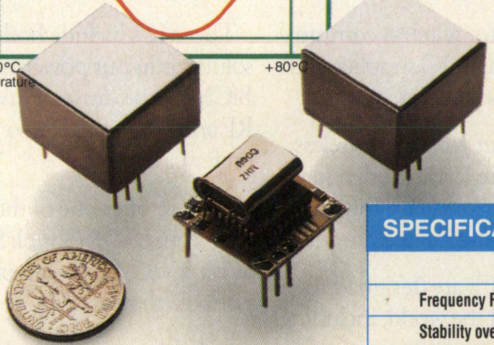
**ISO 9001 Certification** — M-tron Industries and Densitron Microwave Ltd have received certification registration to ISO 9001, from The British Standards Institution/Quality Assurance. ISO is the abbreviation for the International Standards Organization.

**The European Market For Radio Paging** — According to a new report, "The European Market for Radio Paging," by international market analyst Frost & Sullivan, the public services and private systems (on-site and wide area) paging markets are both under pressure from other technologies that want a share of the market. The report concludes that it seems likely that the paging markets in Europe will decline after 1997. Frost & Sullivan predicts that in the late 1990s the public services sector can expect very low growth, with zero growth by the year 2000. In the private systems sector only marginal growth is expected, and the wide-area market is anticipated to decline. However, demand remains strong for very low-cost systems for small users requiring only 20 or 30 receivers in an on-site system, and in those nations where other public services and systems are less effective, such as the former East Germany, Portugal and Spain. For more information

# No compromise necessary.

## Murata Erie's new ASIC-based DCXO



Excellent frequency stability has been a fact of life with crystal oscillators for years. So has small size (a relative term), low cost and flexibility. **But**, did you ever try to get everything that you needed in a single oscillator/package? No way. One, maybe two desirable characteristics but certainly not all. You had to compromise.

**Now, compromise no more.**

Murata Erie has the solution...our new ASIC-based, digitally compensated crystal oscillator is exceptionally stable, truly tiny, low in cost and extremely flexible. It is basically everything that you always wanted in a crystal oscillator but always had to compromise to get.

Check some of these features and specifications:

- To ±0.25ppm stability
- Less than 5.0 cubic centimeters volume
- 100 mW max power
- Built-in interface for reconfiguration capability and software
- Single +5VDC power supply

**SPECIFICATIONS**

| | DC2210 AH |
|---|---|
| Frequency Range | 10MHz to 25MHz |
| Stability over −40°C to +85°C | ±0.5ppm |
| Output | "HC" CMOS |
| Power Dissipation | +5V @ 15mA |
| Short Term Stability | 1×10E−9 @ T=1 sec. |
| Package Size | 0.79" × 0.79" × 0.45", PC PINS |

Find out more about how you can take the compromise out of your crystal oscillator specifications. Write for complete details on Murata Erie's new ASIC-based DCXO or call 1-800-831-9172, FAX: 1-404-684-1541.

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INFO/CARD 23


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contact Amy Arnell at, Tel: (415) 961-9000, Fax: (415) 961-5042.

**Swedish Scientists To Use GPS In Geodetic Network** — Fifteen navigation receivers designed to generate precise position and timing information from signals broadcast by the satellite-based Navstar Global Positioning System (GPS) were recently delivered to Sweden's Chalmers University of Technology in Goteborg. The receivers, supplied by Allen Osborne Associates, Inc., are being installed in a nationwide geodetic network stretching from Hassleholm and Onsala in the southern part of the country to Esrange and Overkalix above the Arctic Circle in the country's far north. The new GPS receivers will help Sweden's National Land Survey and Chalmers University, joint operators of the 20-station network, called SWEPOS, to detect horizontal and vertical land movements as small as one centimeter in a year throughout Scandinavia.

**Video Network In The Middle East** — Orbit Communications Company Limited (OCC) has selected Scientific-Atlanta to provide a consumer digital video compression satellite network to distribute video programming throughout the Middle East. Under the terms of the contract, Scientific-Atlanta will provide digital compression uplink equipment, integrated receiver decoders (IRD's) and head-end equipment and will be responsible for the installation, operation and maintenance of the network. The network will utilize the company's encryption and conditional access system enabling only authorized subscribers to receive signals.

**Navy Signs Eastern Microwave Corporation** — Eastern Microwave Corporation has been awarded a contract by the U.S. Navy to design, develop, and manufacture antenna array and positioning systems. The systems will be incorporated into surface search radars for both domestic and international installations.

**Harris Wins Contract For Portugese Air Traffic Control Center** — Harris RF Communications has signed a $6.9 million contract with the Portugese Aeroportos e Navegaçao Aérea (ANA) for the design, engineering and installation of a new Oceanic Radio and Telephonic Communications System. The communications system, which will include both air-to-ground and ground-to-ground components, will be installed at the new Oceanic Air Traffic Control Center to be established in Lisbon. Harris will provide HF and VHF air-ground radio systems for links with aircraft during oceanic flight, a microwave system connecting remote transmit and receiver sites outside Lisbon, and telephone and switching system for communications among controllers, operators, supervisors and the technical support staff.

**A New location** — The consulting office of Earl McCune announced that it has relocated. The new facility provides expanded capacity to serve and support their clients and develop RF communications products and wireless hardware. The new address is 2383 Pruneridge Avenue, Suite 3, Santa Clara, CA 95050-6461, Tel: (408) 983-1076.



INFO/CARD 25

IF YOUR SIGNAL SOUNDS LIKE THIS, YOU SHOULD BE TUNING IN TO US.

# Our fully-programmable PLL frequency synthesizers are the clear choice in signal stability.

**Motorola's New PLL Frequency Synthesizers**

| Part Number | Max. Frequency | Input Sensitivity | Main Supply | Current Consumption (Nominal) | Phase Detector Supply | On-Board Prescaler | Low-Power Standby |
|---|---|---|---|---|---|---|---|
| MC145170 | 160 MHz | 500 mV$_{p-p}$ | 2.5*– 6 V | 2.5 mA @ 3 V<br>7 mA @ 5 V | 2.5 – 6 V | 4/5 | No |
| MC145170-1** | 180 MHz | 500 mV$_{p-p}$ | 2.7*– 5.5 V | 2.5 mA @ 3 V<br>7 mA @ 5 V | 2.7 – 5.5 V | 4/5 | No |
| MC145190 | 1.1 GHz | 200 mV$_{p-p}$ | 4.5 – 5.5 V | 7 mA @ 5 V | 8 – 9.5 V | 64/65 | Yes |
| MC145191 | 1.1 GHz | 200 mV$_{p-p}$ | 4.5 – 5.5 V | 7 mA @ 5 V | 4.5 – 5.5 V | 64/65 | Yes |
| MC145192 | 1.1 GHz | 200 mV$_{p-p}$ | 2.7 – 5.0 V | 7 mA @ 3 V | 2.7–5.5 V*** | 64/65 | Yes |
| MC145200** | 2.0 GHz | 200 mV$_{p-p}$ | 4.5 – 5.5 V | 12 mA @ 5 V | 8 – 9.5 V | 64/65 | Yes |
| MC145201 | 2.0 GHz | 200 mV$_{p-p}$ | 4.5 – 5.5 V | 12 mA @ 5 V | 4.5 – 5.5 V | 64/65 | Yes |

*100 MHz @ 3 V **Engineering samples now, production 4Q93 ***Phase detector A = 4.5 – 5.5 V, Phase detector B = 2.7 – 5.5 V

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RF10/93

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Application Requirements__________

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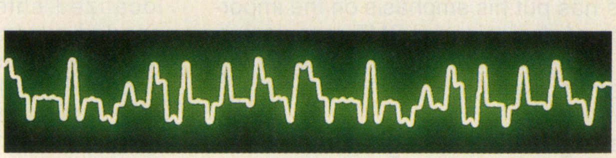

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***MOS Digital-Analog Integrated Circuits Division***



INFO/CARD 26

# Surface-Mount Technology Reaches 'Standard' Status

*By Gary A. Breed*
*Editor*

The use of surface mount technology was pioneered by RF engineers, who were its first major users with chip capacitors and strip-leaded semiconductors. High volume SMT manufacturing has only become common in the past few years. A look inside a pager, personal computer or television receiver demonstrates the progress — in 1983, SMT assembly would be hard to find; in 1988, SMT use was mixed with standard through-hole construction; and today, you would be rarely find one of these products without extensive (or exclusive) use of SMT. This method of electronic assembly has reached the status of "standard operating practice."

SMT acceptance has come, somewhat surprisingly, in spite of the lack of standard package sizes and footprints. There are numerous standards, but each has many variations from one manufacturer to another. For example, clearance under the package and package height can vary widely for parts in the "same" standard package.

Conversations with Bob Barron of Stanford Telecom's MQA contract manufacturing facility revealed additional insight into the role of SMT: The density of circuit boards, the positioning of components with respect to one another, and the mixing of digital and RF circuitry within a small area are problems that must be addressed by designers. It is far too costly to deal with these problems after a design is completed and supposedly ready for production. Barron's has put his emphasis on the importance of designing for SMT manufacturing into writing in this issue's Cover Story on page 65.

## Dealing With Packaging

It is interesting to note that SMT manufacturers have worked around the problem of non-standard devices. Some parts have varying package footprints, some are not available in surface-mount packages, and some are unusual components that require hand placement and soldering.

In order to meet the demand for smaller, high volume products, manufacturers have simply learned to handle these disparate components. Machine vision has been employed to verify the placement of parts. The camera/computer combination can recognize whether the components are aligned with the circuit board pads, and can adjust placement as programmed by the operator.

Combination SMT and through-hole mounting is routine, despite being considered a major problem area in the early days of SMT manufacturing. SMT mounting on both sides of p.c. boards is also a technique that has been developed to meet the needs of designers who require specific parts placement to meet design goals, as is the case with many RF circuits.

***SMT manufacturers have worked around the problem of non-standard devices***

Hand assembly and soldering has been accepted as a necessity by RF product manufacturers. Many RF components are modular or have packages dictated by the physical construction of the devices themselves (e.g., inductors, connectors, filters, isolators). In addition, localized shielding covers are often needed over circuits that are prone to radiation or susceptible to external influences. Manufacturing facilities keep assembly technicians on staff who are trained in proper procedures for these devices.

Of course, all hand operations add to the cost of the finished product, and work continues to provide more components in SMT packages. Cost is one of the primary goals that justifies manufacturing with SMT. Overall size and weight reduction cuts the cost of an enclosure, which can be the single most expensive component in some products. Weight reduction cuts shipping costs and storage space requirements.

## EM Fields In Small Spaces

A particular problem with miniaturization of all kinds is coupling between circuit elements, whether SMT, chip-and-wire hybrids or multi-chip modules (MCMs). At first, one might think that coupling, parasitic effects and radiation from smaller circuits would be less than their larger predecessors. While conductors may be shorter, they are also narrower, which results in higher inductance per unit length. Closer spacing means that fields coupled from one part of the circuit to another can be higher, since field strength increases by $R^2$ instead of geometrically, although this is mediated somewhat by the generally smaller power consumption and lower circuit currents with newer components.

Jim Muccioli of JASTECH has researched radiation from digital integrated circuits using new small-geometry, high-speed processes. In comparing a 1.95 micron microprocessor with its 1.5 micron version, he found that the smaller geometry device had much greater emission levels at high frequencies. For example, above 200 MHz, the 1.5 micron process device had emissions that were 10-15 dB greater than the 1.95 micron device. The message for engineers designing smaller circuits is that some of the SMT components rely on smaller device geometries to minimize die size. They are inherently faster devices which emit greater high-frequency noise that their predecessors.

Dealing with problems like these are among new requirements being placed on design engineers. Designing for SMT has a whole new set of mechanical constraints, as well. SMT circuits are rigid structures. Board flexing, different temperature coefficients, shock and vibration resistance are more important than before. RF engineers have a significant challenge before them. *RF*

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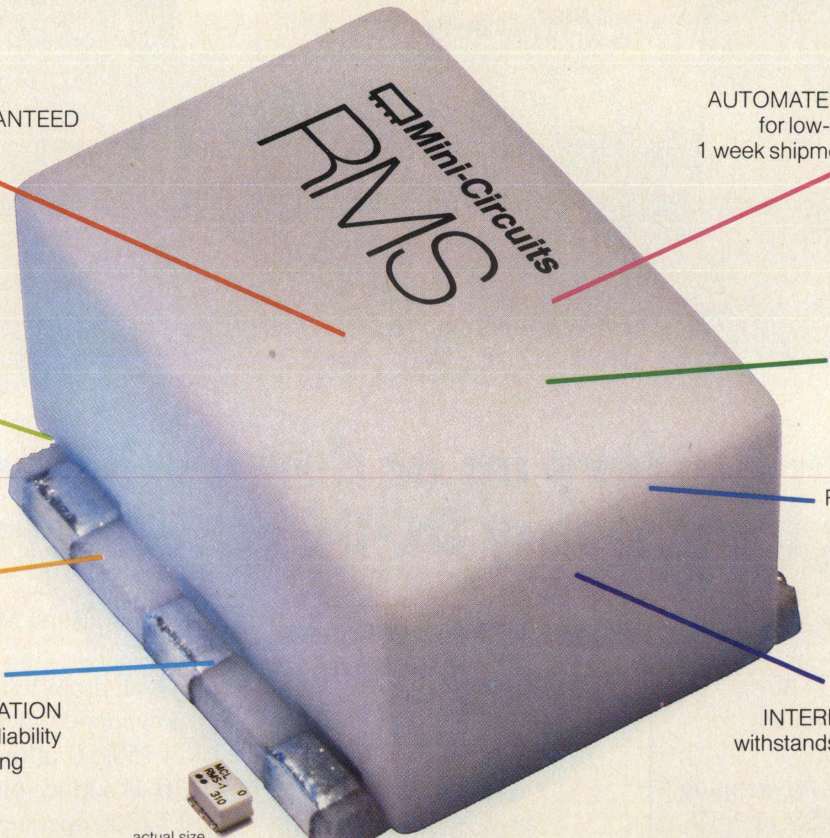


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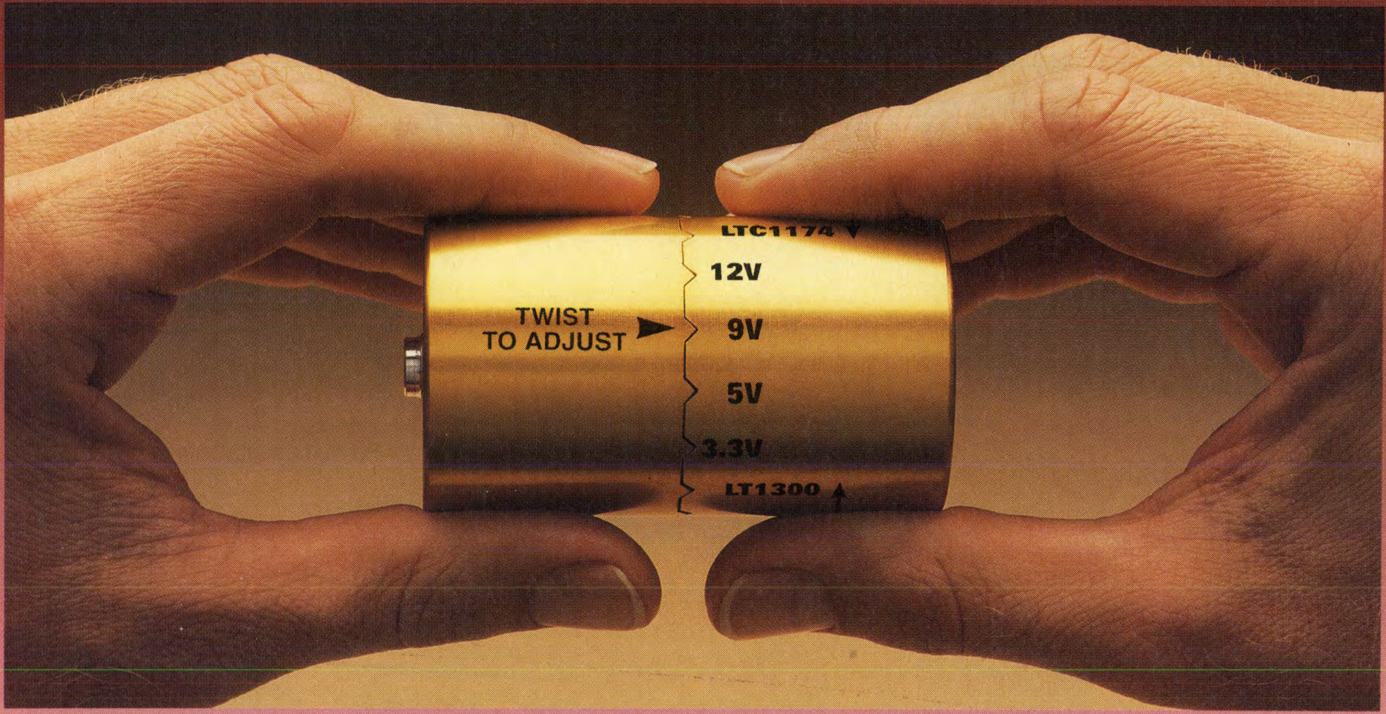
INFO/CARD 27

# Any Battery Any Voltage
# It's That Simple!

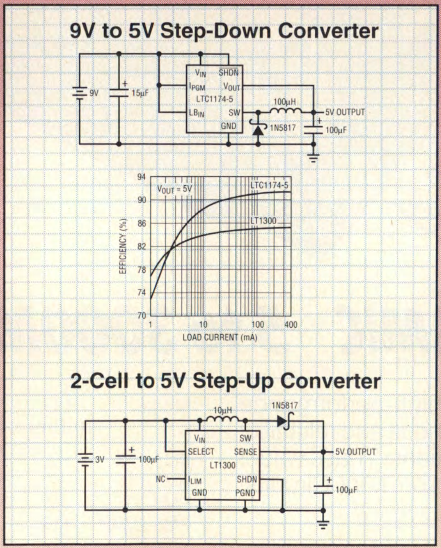



## Step your battery up or down with our new high efficiency DC-DC converters.

Low quiescent current, high efficiency, and minimum board space distinguish these two DC-DC converters designed specifically for battery-powered systems.

The LTC1174 is ideal for stepping battery voltages down to 5V or 3.3V at up to 400mA output current, or for converting 5V to 3.3V or even -5V. Thanks to Burst Mode™ operation, no load supply current is just 130µA. In shutdown, the LTC1174 draws an industry low 1µA. True current mode operation at 180kHz allows smaller inductors and safe control of short-circuit currents. And 100% duty cycle operation makes this the industry's only low dropout DC-DC converter.

The LT1300 step-up DC-DC converter provides 5V or 3.3V from a 2-cell battery at up to 250mA output current. Burst Mode operation maintains quiescent current at a low 120µA, which drops to only 10µA in shutdown mode. A switching frequency of 150kHz allows the use of small 10µH surface mount inductors. The LT1300 operates down to a low 1.8V input, so you can squeeze all the energy out of the battery.




Take your best step up (or down) in DC-DC conversion with LTC's Burst Mode DC-DC converters.

Both the LTC1174 and LT1300 are available in industry standard 8-lead DIP or SOIC packaging. Pricing starts at $3.63 in 1000s for the LTC1174 and $2.87 for the LT1300.

For details, contact Linear Technology Corporation, 1630 McCarthy Blvd., Milpitas, CA 95035/ 408-432-1900. For literature only, call **800-4-LINEAR.**

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# Tampa Conference Features Special Track on Space Applications

**Here is the updated schedule of papers for RF Expo East '93, to be held in Tampa, Florida from October 19 to 21. Interested engineers can register by telephone at (800) 525-9154 or (303) 220-0600. Fax inquiries can be sent to (303) 770-0253.**

**Tuesday, October 19**
**Digital Communications and DSP**

**Session A-1— 8:30-11:30 a.m.**
**Digital Communications**

*An ISM Band Design for WLAN and PCS*
Robert Zavrel
GEC Plessey Semiconductors

*Integrated Modem/RF Design Architectures for Reduced Power, Increased Capacity F-QPSK Wireless Systems*
Kamilo Feher, University of California, Davis

**Session B-1— 1:30-4:30 p.m.**
**Digital Communications and DSP**

*A DSP Microprocessor Based Receiver for a Cosine Transition-shaped BPSK Signal*
Bruce H. Williams, Roy E. Greeff
Paramax Systems Corp.

*Designing a High Performance Monolithic PSK Modulator*
Robert Zavrel, GEC Plessey Semiconductors

**Tuesday, October 19**
**Test & Measurement**

**Session A-2 — 8:30-11:30 a.m.**
**System Performance**

*Methods for Estimating and Simulating the Third Order Intercept Point*
Carl Stuebing, Mojy C. Chian
Harris Semiconductor

*Low Cost Phase Noise Measurement Technique*
Jim H. Walworth, Tampa Microwave Lab, Inc.

*Externally-Induced Transmitter Intermodulation: Measurement and Control*
Ernie Franke, E-Systems ECI Division

**Session B-2 — 1:30-4:30 p.m.**
**Test Methods and Equipment**

## RF Expo *PLUS* — Extra Technical Sessions on New Space Applications

Tuesday, October 19: 8:30-11:30 a.m.

**EXPO *PLUS* Session 1 —**
**Satellite and Space Systems**

*Overview of Current Satellite Systems*
A review of major satellite systems with commercial applications, including GPS, VSAT, LEO and geosynchronous technologies. New and existing systems are described in a general manner, covering the basic mission and capabilities.

*The GEOSAT Follow-on (GFO) Altimeter*
Dan Walker, E-Systems, Inc.
The Navy GEOSAT mission demonstrated the ability of a radar altimeter to measure ocean features with an accuracy of 3 cm. The GFO radar under develoment achieves the same precision with 1/3 the weight and 1/2 the power consumption.

Wednesday, October 20: 8:30-11:30 a.m.

**EXPO PLUS SESSION 2 —**
**Components for Space Applications**

*Designing Microwave Circuits for Geosynchronous Space Applications*
Ron Ogan, Trak Microwave Corp.
Component design for the harsh envrironment of space must allow for radiation, outgassing, and exceptionally high reliability. Design and test criteria for these components are covered in this presentation.

*Low Cost Plated Plastic Diplexers for Use in Commercial Mobile Satellite Applications*
Chip Scott, Teledyne Microwave
Light weight and low cost were the forces driving the development of a new family of filters, diplexers and other cavity devices for mobile and fixed satellite earth terminals.

Thursday, October 21: 8:30-11:30 a.m.

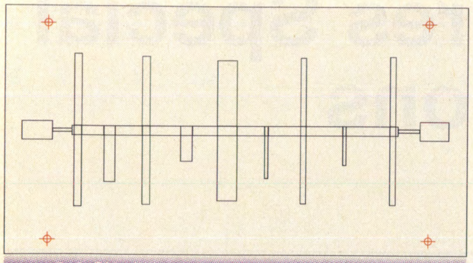
**EXPO PLUS SESSION 3 —**
**Satellite and Space System Performance**

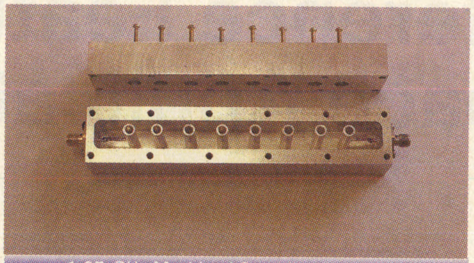
*Satellite Channel Utilization in the Presence of Rain Attenuation*
Kaivan A. Karimi, Valentine Aalo,
Florida Atlantic University
Rain is the most dominant cause of signal degradation in Ka band satellite links. This paper descibes an adaptive rain-fade countermeasure based on effective utilization of channel capacity.

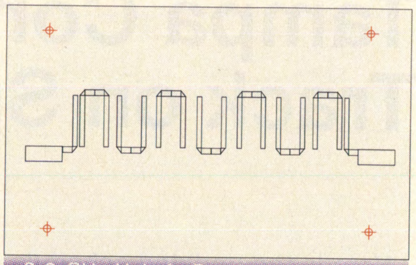
*Hardware Verification of Communication System Simulations*
Henry Helmken,
Florida Atlantic University
The mission of the Space Communications Technology Center at Florida Atlantic University is sponsored by NASA to develop systems for digital satellite communications. Hardware cost is increasing the need for computer simulation. Those simulation efforts are described in this paper.

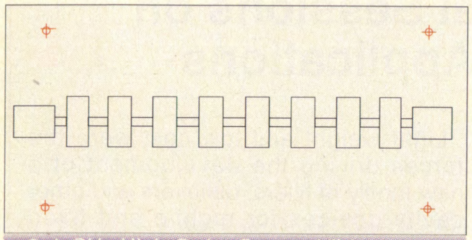
*GFO Water Vapor Radiometer*
Muhammad A. Malik, E-Systems, Inc.
Measuring sea surface topography is the mission of the Navy GEOSAT follow-on program. This paper describes the signal velocity error-correction capabilities of a water-vapor radiometer system.

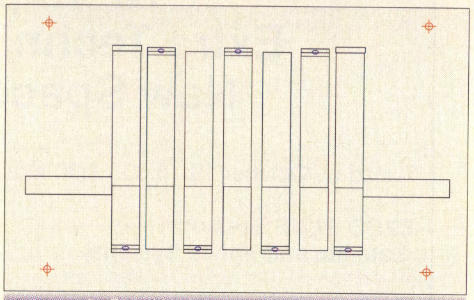
# WHAT DO THESE FILTERS HAVE IN COMMON?

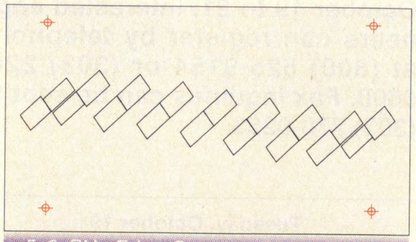

5.6 GHz Elliptic Bandpass on Teflon Board

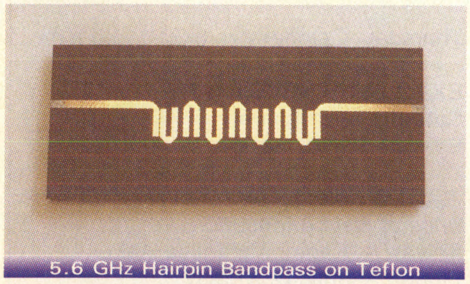

1.27 GHz Machined Combline Bandpass

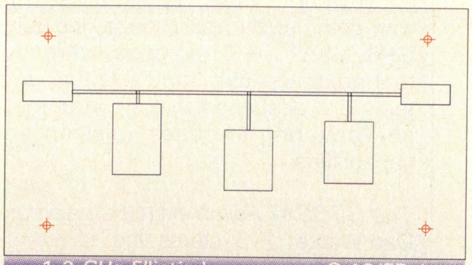

2.2 GHz Hairpin Bandpass on Teflon

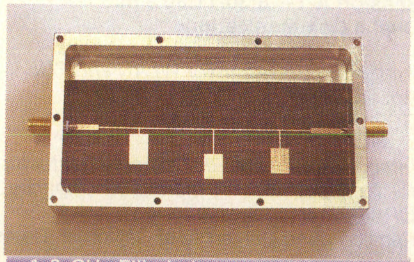

6.4 GHz High-Order Stepped-Z Lowpass


10 GHz Wideband Interdigital BP


5.6 GHz Edge-Coupled BP on Teflon


5.6 GHz Hairpin Bandpass on Teflon


1.2 GHz Elliptic Lowpass on G-10/FR-4


1.2 GHz Elliptic Lowpass (Mounted)

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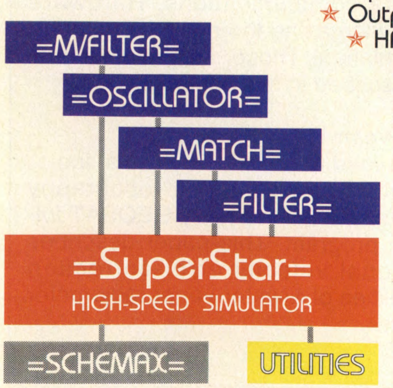

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*A 3 GHz 50 ohm Probe for PCB Measurements*
Joel Dunsmore, Robert Kornowski, Chuck Tygard, Hewlett-Packard Co.

*Low Cost RF Tuner System for JDC Load Pull and SSPA Design*
C. Tsironis, Focus Microwaves Inc.

*Noise Figure and Gain Measurement on High Speed Bipolar Junction Transistors*
Wayne Jung, Tektronix, Inc.

**Tuesday, October 19**
**Essential RF Circuits**

**Session A-3 — 8:30-11:30 a.m.**
**Amplifier Design**

*The Current-Feedback Op Amp, A High-Speed Building Block*
Anthony D. Wang, Burr-Brown Corporation

*The SLAM: A New Ultralinear Power FET Module Concept for HF Applications*
Adrian I. Cogan, Lee B. Max
MicroWave Technology, Inc.

*Tradeoffs in Practical Design of Class E High-Efficiency RF Power Amplifiers*
Nathan O. Sokal, Laszlo Drimusz, Istvan Novak, Design Automation

**Session B-3 — 1:30-4:30 p.m.**
**RF Power**

*High Power, Low Frequency Microstrip Switches*
S. Irons, E. Higham, M/A-COM

*Class-E Power Amplifier Delivers 24 W at 27 MHz at 89-92% Efficiency, Using One Transistor Costing $0.85*
Nathan O. Sokal, Ka-Lon Chu
Design Automation

*The CAM: A UHF/L-Band FET Module for Pulsed Power Avionics Applications*
Frank Sulak, Ken Sooknanan, Adrian I. Cogan, MicroWave Technology, Inc.

**Wednesday, October 20**
**Wireless Personal Communications**

**Session C-1 — 8:30-11:30 a.m.**
**Wireless Communications Systems**

*Frequency Synthesizer Strategies for Wireless*
Bar-Giora Goldberg, Sciteq Electronics

*TDMA Transmitters — Characterizing Power, Timing and Modulation Accuracy*
Helen Chen, Hewlett-Packard Co.

*Performance Simulation of a Low-Power In-Building Wireless Centrx System*
Douglas Alston, BellSouth Telecommunications

**Session D-1— 1:30-4:30 p.m.**
**Components for Wireless**

*Practical Applications of a Low Cost Low Noise GaAs PHEMT MMIC for Commercial Markets*
Al Ward, Henrik Morkner, Hewlett-Packard Co.

*Highly Integrated GaAs MMIC RF Front End for PCMCIA PCS Applications*
Thomas Kotsch, Andy Laundrie, Steve Geske, Howard Fudem, Jim Blubaugh, Sanjay Moghe, Northrop Corp.

*Low Power Transmitter Design Using SAW Devices*
Earl Clark, RF Monolithics, Inc.

**Wednesday, October 20**
**Specialized Design Techniques**

**Session C-2 — 8:30-11:30 p.m.**
**RF Applications**

*FMCW Radar Architecture*
Ken Puglia, M/A-COM

*Filter Comparator Network for Beam Position Monitoring*
Michael Ferrand and Mark McWhorter
Lorch Electronics

*Digital Temperature Compensation of Oscillators Using a Mixed Mode ASIC*
Steve Fry, Murata Electronics North America

**Session D-2 — 1:30-4:30 p.m.**
**Synthesizers**

*A Synthesizer Design Program With Detailed Noise Analysis*
Terrence Hock
National Center for Atmospheric Research

*PLL Settling Time: Phase vs. Frequency*
Donald E. Phillips, Rockwell International

*Linear Frequency Modulation — Theory and Practice*
Bar-Giora Goldberg, Sciteq Electronics

**Wednesday, October 20**
**Analytical Methods**

**Session C-3 — 8:30-11:30 a.m.**
**Modeling for CAD**

*RF Active Device Modeling for CAD, A Coming Necessity*
Gary Roberts, Hewlett-Packard Co.

*Regression Based Algorithms for Inductor Modeling*
Edmund (Joe) Tillo, Ford Motor Co.

*Computer Aided Design Tools for Small Signal RF Matching Networks*
Michael Rothery, Sam Ritchie, Madjid A.Belkerdid, University of Central Florida

**Session D-3 — 1:30-4:30 p.m.**
**CAD Methods**

*SAW Resonator Oscillator Design Using Linear RF Simulation*
Alan R. Northam, RF Monolithics, Inc.

*Embedding RF Design Tools in an IC Design System*
Mojy C. Chian, Steve S. Majors, Alan G. Whittaker, Harris Semiconductor

*Electromagnetic Simulation for High Frequency Planar Circuits*
Daren McClearnon, Hewlett-Packard Co.

**Thursday, October 21**
**Wireless Applications**

**Session E-1 — 8:30-11:30 a.m.**
**Wireless Applications**

*A Monolithic 915 MHz Direct Sequence Spread Spectrum Transmitter*
Stephen Press, Tektronix, Inc.

*A Low Power RFID Transponder*
Raymond Page, Wenzel Associates, Inc.

**Thursday, October 21**
**RF Circuits and Systems**

**Session E-2 — 8:30-11:30 a.m.**
**RF Circuits and Systems**

*Multi-Component Module for High Speed Passive Design*
Mark Brooks, Thin Film Technology

*Design of a Search Based PLL*
Mike Black, Texas Instruments

*Analysis of Transversely Coupled SAW Resonator Filters Using COM Techniques*
V.Narayanan and S.M. Ritchie
University of Central Florida

**Thursday, October 21**
**Practical RF CAD**

**Session E-3 — 8:30-11:30 a.m.**
**Practical RF CAD**

*Basics of CAD at RF for Wireless Circuits and Subsystems*

*Defining Circuit abd Subsystem Specifications for Cordless Telephone Applications*

*Design Examples of Small-Signal Circuits Operating from 3.3 to 4.5-Volt Supplies*

*Design Examples of Large-Signal Circuits*

*Using CAD for Circuit Layout and Packaging Design*
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INFO/CARD 36

# Designing with Ultraminiature SMT Semiconductor Packages

*By Terry Cummings*
*California Eastern Laboratories*
*and Paul Edwards*
*ARXE, Inc.*

*The benefits of today's miniature semiconductor devices cannot be denied. Small plastic packages are inexpensive, easy to use, weigh less and require less real estate; important considerations in the design of portable handheld products. A small package size can also improve a device's repeatability. As parts shrink, interconnections are smaller and closer, and the mechanical tolerances between leads and solder pads are reduced, all of which help to reduce device parasitics. This article describes the problems engineers will encounter when designing with ultraminiature packages, and identifies the solutions.*

Solder joint volume, often overlooked by designers, plays a more important role in circuit performance as frequencies increase. Smaller parts mean smaller pads, and that means smaller solder volumes. When the variation in solder volume as a percentage of wavelength decreases, parasitics are reduced even further. From a mechanical standpoint, substrate choice is not as critical with smaller parts; their smaller footprints make their solder blocks less inclined to break if the board should warp or flex.

Electrically, smaller packages should have less lead inductance and lower package capacitance. So potentially, more performance can be squeezed from the encapsulated die. Ideally, as package size is reduced, performance should approach that of the chip itself. In reality, gain and noise figures do improve somewhat as the package size is reduced — when specific package configurations are compared. The important point is that RF performance is not compromised when choosing these miniature parts. See Table 1.

On the other hand, heat dissipation does suffer, but only to a degree. Smaller packages have less surface area to

**NE680 Package Comparison vs Performance MAG**

| F MHz | 19 Pkg | 30 Pkg | 33 Pkg |
|---|---|---|---|
| 500 MHz | 20.2 dB | 20.1 db | 19.7 dB |
| 800 | 18.0 | 17.7 | 17.0 |
| 1000 | 16.8 | 16.1 | 15.7 |
| 1500 | 14.4 | 13.4 | 12.7 |
| 2000 | 10.9 | 10.0 | 9.6 |
| 2500 | 8.9 | 8.2 | 8.0 |

Vce=2.5 V Ic=3mA

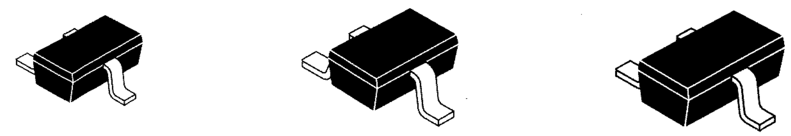

**Table 1. NE685XX performance table and package outlines.**

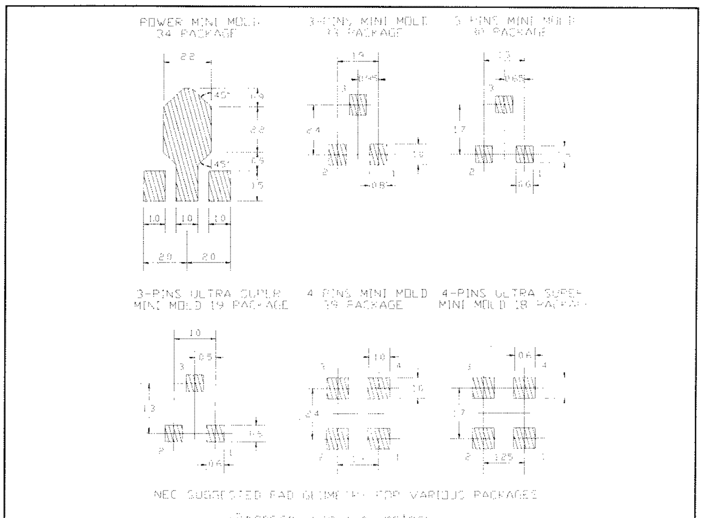


**Figure 1. NEC suggested pad geometry for various packages.**

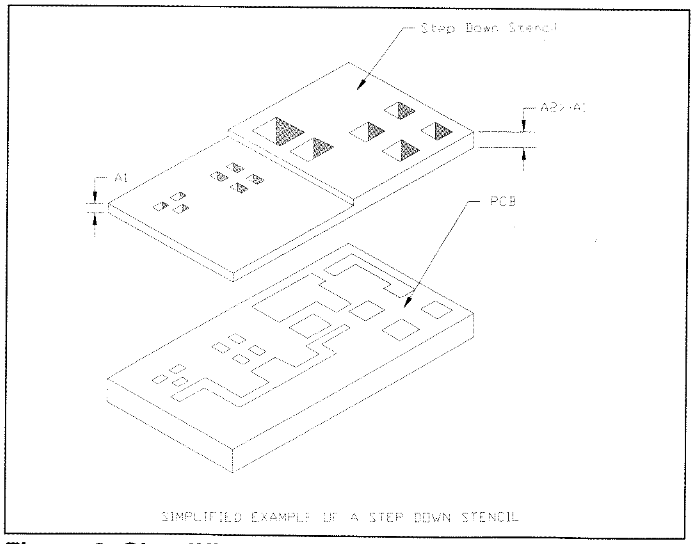


**Figure 2. Simplified example of a step down stencil.**

dissipate heat, and the collector leads are smaller making the $R_{th}$ higher. In most small signal applications, these devices consume so little power, heat dissipation is not an issue. But if heat dissipation becomes a concern, as in the case of oscillator design, the part can be epoxy bonded to the PCB to create a better thermal path.

As frequencies increase and the gap between theoretical and actual circuit performance widens, interaction between designer and assembler becomes more and more important. Automated assembly at these frequencies present unique challenges. Concurrent engineering is the only way to tackle them. These challenges fall into two general categories: performance and manufacturability.



## Circuit Performance

Above UHF, circuit performance must take into account topics that aren't much of a concern in designs at lower frequencies: solder volume, solder screens, the solder paste itself, pad geometry and size, and device alignment.

As previously mentioned, solder joint volume affects the circuit parasitics and energy reflections at the device/PCB interface. Solder volume is controlled by the pad size and the solder mask aperture (defined by the designer) and by the solder paste, the solder screen, and the stenciling process (defined by the assembler).

Device manufacturers can provide recommended PCB layout geometries that show the location and size of solder pads for the parts you've specified (Figure 1). Share them with your assembler, but remember, they're only recommendations. Your assembler's unique experience may lead to modifications.

Once pad geometry is decided, a solder mask is created on the PC Board. Tiny apertures in the solder mask align with the pads in your circuit. Solder paste is laid into these apertures, the thickness of the mask acting as a wall, or dam, to keep the solder from spreading. Obviously, the size and position of these holes are critical — especially for miniature parts. Look closely at your parts' leads. Variation in size can be nearly indistinguishable, but if one is larger — generally the collector — it will require a larger pad and larger hole in the mask.

If the apertures are too small, there's a chance that not enough solder or bonding area will be available to make effective solder fillets. This can result in a solder joint that makes no contact, or worse, makes a mechanically weak contact that's unreliable and costly to trace.

If an aperture in the mask is too large, a lead could get too much solder. When heated, the surface tension of the extra solder could be significant enough to pull the part out of alignment, and/or out of contact with its pads.

While size and position of the apertures in the solder mask are critical, control of the volume of solder paste is just as important. Controlling solder volume is especially tricky when miniature devices are combined with large ones that require relatively massive amounts of solder. Too much solder can cause tiny parts to float out of alignment during the reflow process — and increase parasitic capacitance from the pad to the lead frame or die, a condition that can unpredictably effect performance.

Solder paste is applied to the PCB by

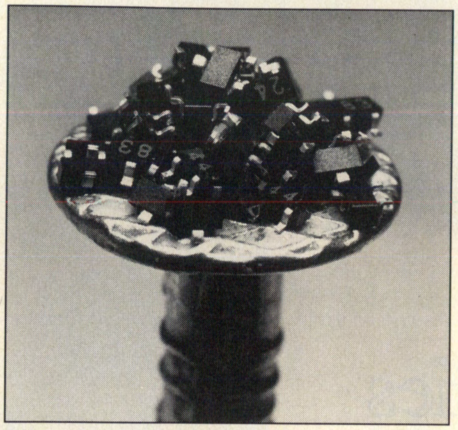

**Figure 3. NE68519 packages on the head of a nail for size reference.**

forcing it through tiny apertures in a solder stencil. Obviously these apertures must be aligned with those in the solder mask. Solder volume is determined by the size of the aperture in the stencil, thickness of the stencil, and the formulation of the solder paste itself.

Experienced assemblers understand the nuances of solder paste screening and the resulting reflow characteristics. Smaller parts mean smaller areas of paste, so a more viscous formulation with high metal content is used to assure sufficient volume.

Once the solder formulation is determined, the assembler can specify one of two different kinds of stencils to control its volume. The first is a stencil of uniform thickness that features apertures sized to leave exactly the volume of solder required. This aperture size is a function of the solder formulation, the stencil manufacturer's capabilities, the assembly process ... and a good deal of the "black magic" that comes from experience.

The alternative is a step down stencil. A step down stencil is thin in the area of your miniature devices, limiting the volume of solder that's laid down, and thicker — sometimes twice as thick — under your large parts, providing deeper apertures for the larger volumes of solder paste these parts require (Figure 2). The drawback is that you must try to isolate tiny devices on the board, away from the PLCCs (Plastic Leaded Chip Carriers), tantalum and electrolytic capacitors, castellated trimmer caps, canned mixers, and other large components.

## Manufacturability

The manufacturability of an assembly takes into account its "cost-of-design" and its "survivability". Cost-of-design addresses the equipment and process required to produce the assembly. Sur-

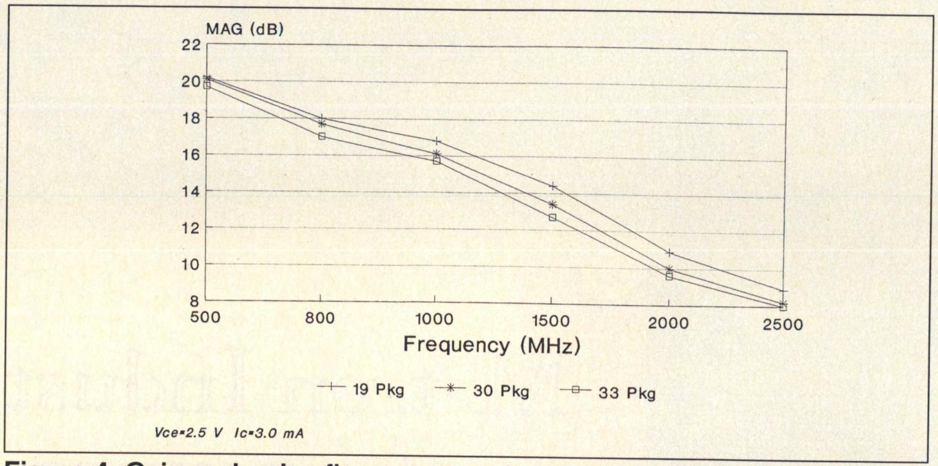


**Figure 4. Gain and noise figure comparison of the NE680 by package.**

vivability covers ESD, parts handling, moisture, and part placement.

ESD is always a concern with higher frequency devices. The enormous voltages generated can wreak havoc on the tiny geometries of high frequency transistors. FETs are especially susceptible. With a few precautions, static problems can be avoided.

First, make it clear to your assembler — ahead of time — that they'll be working with static-sensitive high frequency parts, and even though they may be bipolars, ESD precautions cannot be relaxed. Your device vendor can provide you with specifics, or refer to Military standard guidelines such as DOD-HDBK-263 (available through Naval Sea Systems Command, SEA 3112, Department of the Navy, Washington, DC 20362).

Next, don't give your assembler small parts in bulk bags. Their size makes them difficult to handle, and extra handling increases the potential for ESD problems. Whenever possible, see that parts are provided on tape and reel—and handle those reels like eggs.

On NEC reels, parts sit firmly in perfectly proportioned pockets on the tape. But occasionally, a pocket can be a bit loose. Pick at the end of the tape too much and those flea-sized parts can jump like fleas. Automated pick and place equipment depends on parts being in the correct position in their pockets. If they are out of position, they'll literally have to be tweezered back into their pockets — or hand-

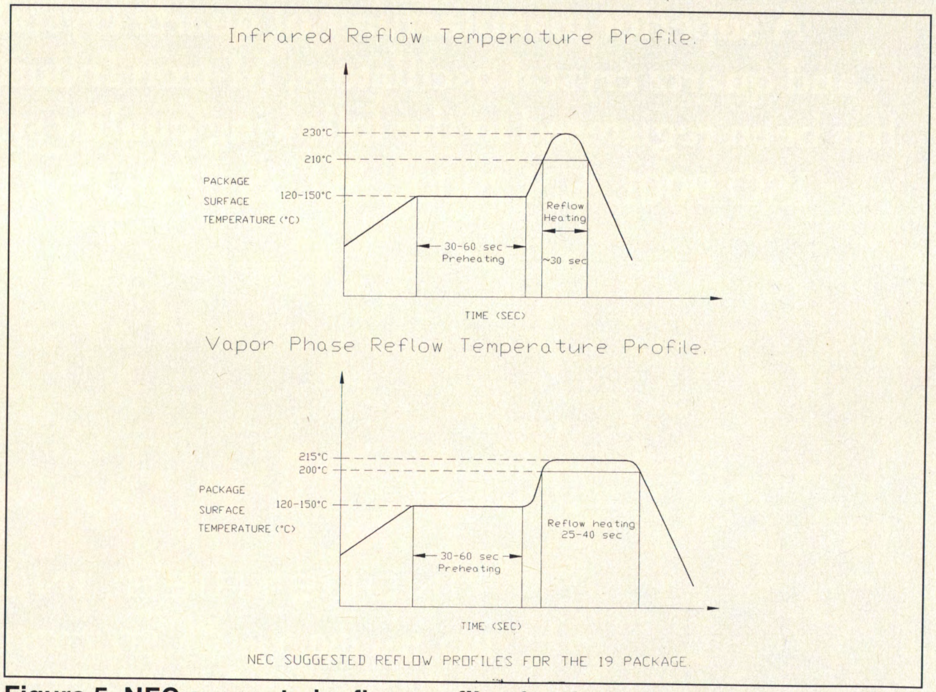


**Figure 5. NEC suggested reflow profiles for the 19 package.**

placed on the board. To avoid the problem, keep the end of the tape tucked in or taped down whenever the reel is transported or stored.

Like movie projectors, automated pick and place machines require a certain amount of leader when threaded. A general rule of thumb is to provide 18 inches of cover tape — the clear tape that holds the parts in their pockets — as a leader. If extra leader is missing from your reel, you may have to sacrifice 18 inches worth of parts to provide it. With some machines you can avoid this by simply attaching a spare piece of cover tape to your parts tape. Your assembler will provide you with the specific needs of his equipment.

## Moisture

Moisture is a critical concern when working with plastic parts. In the reflow processes, parts are quickly blanketed with intense heat of 220° to 250°C. In the wave solder process temperatures rise even faster, and the part is encapsulated in molten solder. Both processes can cause any moisture trapped inside the part to boil and expand, cracking the plastic package. Typical failure points are the fissures and distortions found along the lead frame and the encapsulant interface. The damage may not effect the part immediately, but with its environmental seal lost, reliability is compromised.

The solution is to prebake the parts before assembly. By slowly heating the part to a specified temperature, then holding it at that temperature for a specified period of time, the moisture inside is evaporated. Many manufacturers prebake their devices before packaging. If so, they'll have specific storage, handling and "time-to-reflow" requirements that must be observed. If the prebake is to be done by the assembly house, the manufacturer's bakeout requirements must be followed to avoid damaging the part.

Interestingly, the NEC 19 package (Figure 3, 4) is so small, moisture concerns are minimized. While NEC recommends these parts be stored at 5 to 30° C, and at relative humidities less than 65%, prebaking is not required.

For parts that do require prebake, their embossed tapes usually have a small hole under each part that allows moisture to escape. Since these tapes are not air tight, the reels should be stored in a moisture-minimized environment. This is especially true for reels that have been opened and partially used. Work with your device vendor and assembler to see that everyone understands, and follows, the manufacturer's instructions.

The manufacturer's reflow specifications are also critical to the assembly process. A part's reaction to reflow is heavily dependent upon its size, shape, and composition. Device manufacturers provide guidelines for specific devices, and for the different reflow processes (See Figure 5). These specifications provide reflow temperature versus time profiles and peak temperatures. Again, working closely with your vendor and assembler will help assure success.

## The Assembly Process

Many machine-placed parts will self-center themselves on the PC board during the reflow process. The surface tension of the solder is actually strong enough to pull a component into position on its pads. With miniature parts, the

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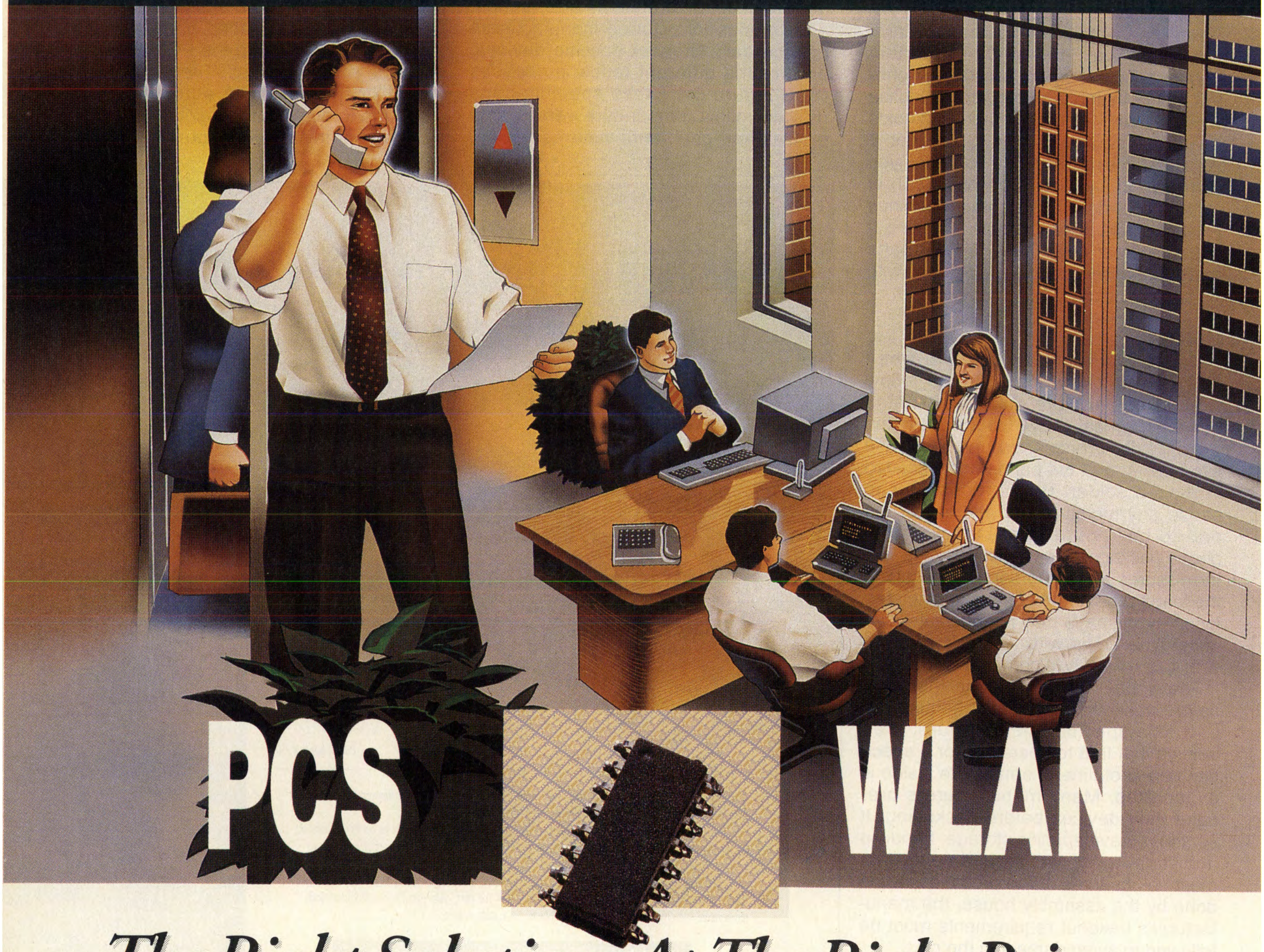


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|---|---|---|---|---|---|---|---|---|---|---|
| **1.7-2.0 GHz** | | | | | | | | | | |
| CAS1401 | 22.5 | 23.5 | 25 | 21 | 2.0:1 | 16 | 20 | 0.8 | 1.0 | 100 |
| CAS1402* | 22.5 | 23.5 | 25 | 21 | 2.0:1 | 16 | 20 | 0.8 | 1.0 | 100 |
| CMM1301** | 23.5 | 24.5 | 30 | 22 | 2.0:1 | 16 | 20 | — | — | — |
| **2.3-2.5 GHz** | | | | | | | | | | |
| CAS2401 | 22 | 23 | 20 | 18 | 2.0:1 | 16 | 20 | 0.8 | 1.0 | 100 |
| CAS2402* | 22 | 23 | 20 | 18 | 2.0:1 | 16 | 20 | 0.8 | 1.0 | 100 |
| CMM2301** | 23 | 24 | 25 | 19 | 2.0:1 | 16 | 20 | — | — | — |

*0.5 mA current required for switching. **Amplifier without switch.

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tiny amounts of solder often don't produce enough surface tension to overcome the part's inertia. This is especially true if there is any flashing left on the package from the manufacturing process. While a tiny burr is insignificant on a large package, it can literally act as a parking brake on a small one.

As a result, placement accuracy is more critical with smaller parts. An assembly machine's jaws can mechanically center the parts to within ±4 mils, but it depends on clean and consistent parts for precision centering. Flashing can cause misalignment in the jaws and throw off the placement, or cause parts to get stuck in the jaws.

The NEC 19 packages have little or no flashing. They can easily be picked and placed with a Zevatech FS710 assembler by handling them as thick 0603 type packages. This FS710 required no elaborate set-up; the NEC 19 packages were loaded and the machine placed them on a PC board in under 15 minutes.

For standard SOT packages, special alignment jaws are available that are designed to fit the SOT's leads. For non-standard packages, the part handling capability of pick and place equipment varies. Generally an assembly machine designed to handle 0603 packages can be made to handle other miniature devices by using a four-jaw chuck and a low gripping force.

Part centering accuracy can be greatly improved with optical centering placement equipment. Optical centering is an expensive add-on to low-volume mechanical-centering assembly equipment, but it's typically standard on high-volume high-speed machines. Optical centering makes placement more precise (±2 mils), and it can help speed the set-up process, and improve the throughput of mechanically inconsistent devices or those with excessive flashing.

## Summary

Automated assembly of ultra miniature semiconductor devices can be simple and straightforward, if you're aware of the potential pitfalls and plan for them accordingly. Choose an assembler with the equipment and the experience to handle small parts, then work together, along with your semiconductor vendor, to see that everyone's needs and expectations are clearly defined and met. *RF*

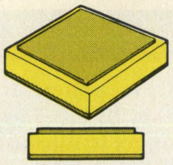
## Acknowledgment

The authors would like to acknowledge that Khanh Luu of CEL helped with the drawings and diagrams.

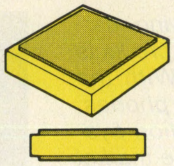
### About the Authors

Terry Cummings has a BSEE from San Jose State University in California. He currently is an applications engineer at CEL and has over 10 years experience in design engineering, production, and product marketing. Khanh Luu is a senior at Cogswell College. He currently is a technician at CEL with 5 years experience.

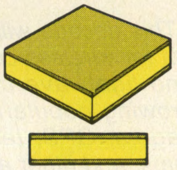




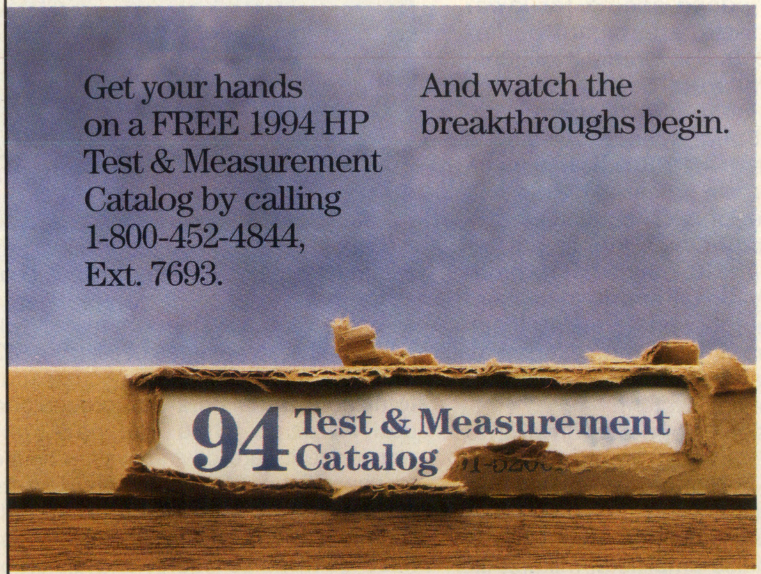












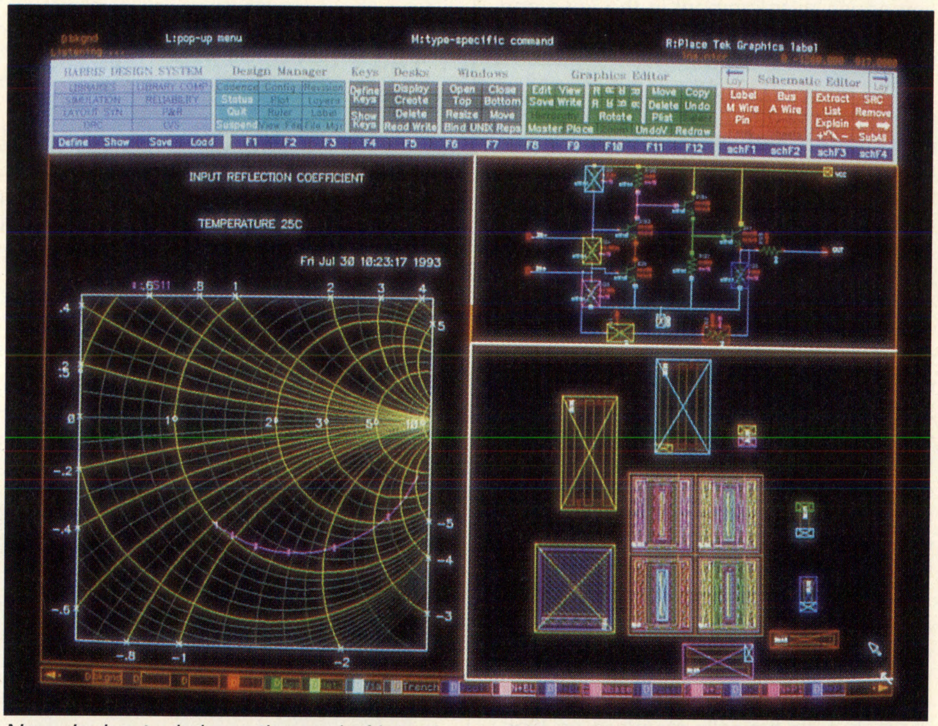
# Merging RF and IC Design Tools for ASIC Development

By Mojy C. Chian and Deborah A. Chian
Harris Semiconductor

*This decade will foster a growing market for wireless communication products and lightwave components. With the growing popularity of cellular phones, pagers, GPS, telecommunication systems, and other applications, the market for wireless products is ever-expanding. There is also an increasing need for optoelectronic components for fiber optic based telecommunications, cable television, and other optical data transfer systems.*

The search for more efficient design methodologies has created a paradigm shift for the designers of RF products. The rapid growth of the wireless market has created a dynamic environment; a fast paced search for lower cost, smaller size, and higher performance products. However, traditional discrete designs are quickly reaching the physical limits of size, parasitics, and electrical performance. A solution that integrates many RF subsystems on a single silicon die promises dramatically smaller size, greater manufacturability, and in many cases higher performance. This article identifies the problems faced by circuit and IC designers as they attempt to work in each other's realm, and shows how Harris Corp. has attempted to address those problems in its Fastrack design system.

For applications from tens of MHz to over 1 GHz, the major design options are discrete components and silicon ASICs. RF designs are typically comprised of many individual subsystems such as low noise amplifiers, mixers, filters, and automatic gain controls, each generally with no more than 10 transistors. In traditional RF designs the subsystems are realized with discrete components, a low level of integration. IC designs typically offer a higher level of system integration and obtain increased performance through higher complexity. With new IC processes offering transistors with an $f_t$ around 10 GHz, silicon ASICs can realistically combine many or all of the RF subsystems on a single die.



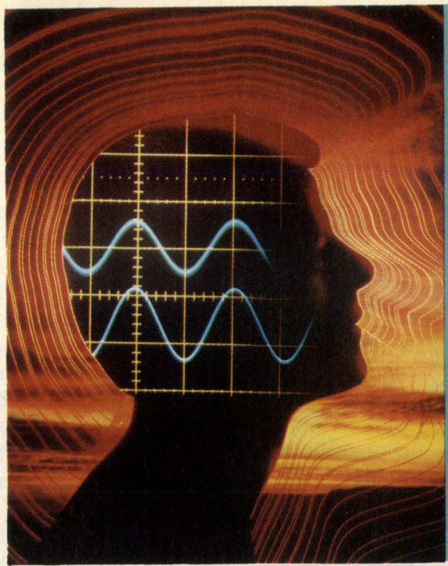

*New design tools have the goal of introducing RF-specific design functions into established IC design, layout and simulation systems.*

The major advantages of silicon ASICs over discrete designs are size, customizable transistors and predesigned cell libraries. Size and packaging requirements are driving system houses to use silicon ASICs. RF and low frequency microwave circuits typically require customized transistors to optimize gain, noise figure, and distortion. Predesigned cells in an ASIC vendor's cell library can dramatically reduce the overall design time of an RF system. IC foundries offering high frequency processes, variable geometry devices, predesigned cells, and a front-to-back design system are an essential ingredient in the design of new RF systems.

## Combined IC and RF Tools

Today, both IC and RF designers are compelled to design in each other's domain. With IC applications approaching 1 GHz or higher, IC designers are advancing into RF design. On the other side, the appeal of monolithics is encouraging RF designers to use silicon ASICs and consequently, IC design systems. IC CAD vendors like Cadence Design Systems, Mentor graphics, View Logic, etc. and RF CAD vendors such as EEsof, Compact Software, and Hewlett Packard, have traditionally focused on one domain or the other. But, a combined system is needed to provide a consistent design environment in which both IC and RF designers can work with familiar tools. There are two ways to create such a combined system:

*Direct Integration* — With direct integration of IC and RF design tools, the



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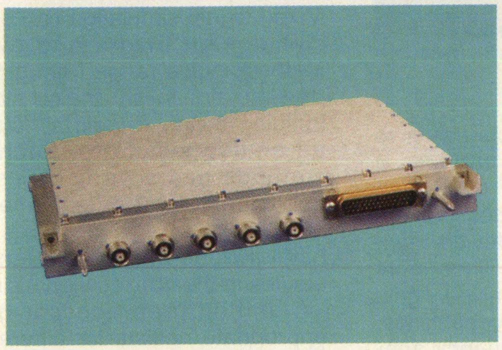
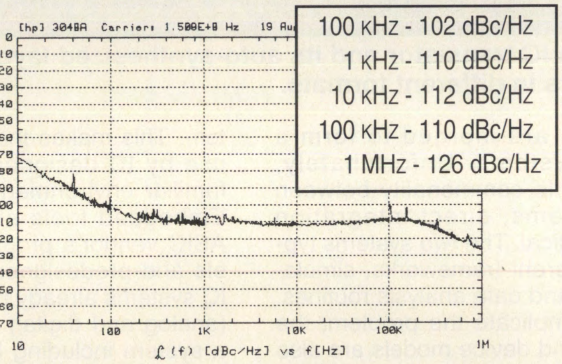
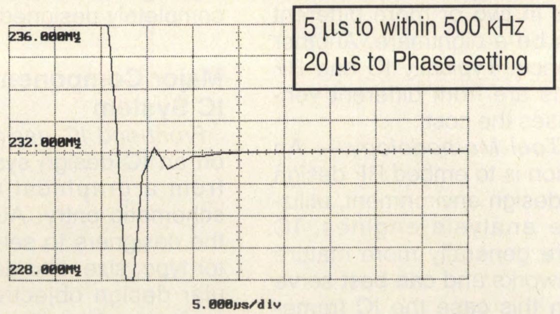





### SPECIFICATIONS

| | |
|---|---|
| Frequency Range: | 1500-1800 MHz |
| Frequency Step Size: | 2 MHz |
| Switching Speed: | 25 µsec |
| Between channels: | 500 nanoseconds |
| Output Power: | +12 dBm |
| Spurious: | -60 dBc |
| Harmonics: | -30 dBc |
| Reference: | L Band |
| Size: | 6" x 9" x 1.2" |

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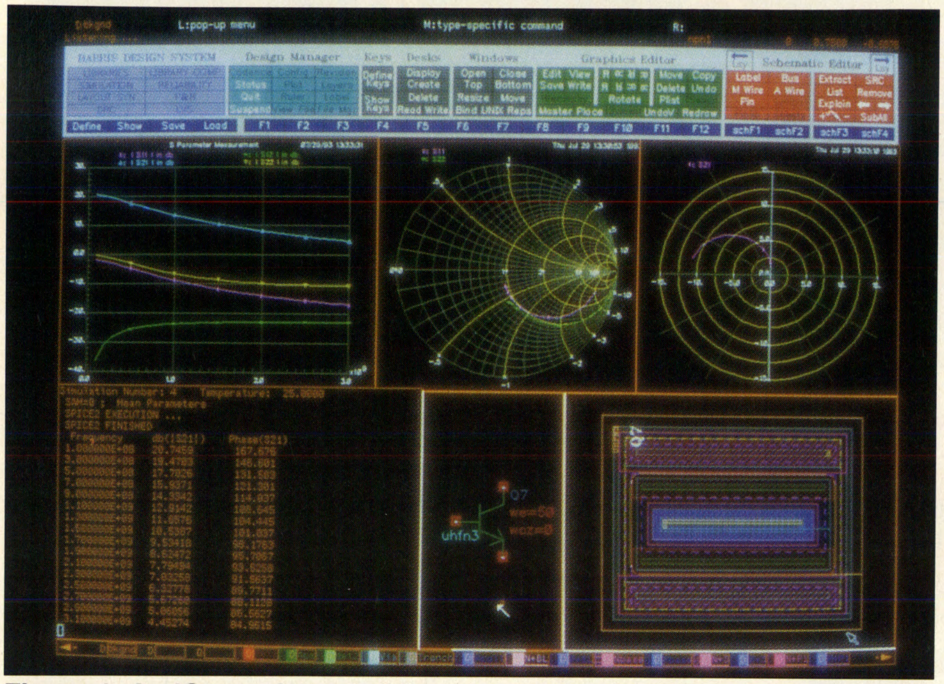


**Figure 1. An IC transistor and its auto-synthesized layout. Plots include S parameters in different formats.**

two systems are merged to form a superset system. Unfortunately, because of little commonality between the two systems, direct integration seems impractical. The two systems typically use different frameworks, simulation engines, and data analysis routines. To further complicate the problem, the cell libraries and device models are also different. Maintaining different sets of device models in two or more different simulators can be a nightmare. Another factor is the price. Typically IC and RF design systems are from different vendors, which raises the cost.

*Embedded Tool Methodology* — An alternate solution is to embed RF design tools in an IC design environment, utilizing the same analysis engines. IC frameworks are generally more mature than RF frameworks and can best serve as the host. In this case the IC framework, simulation engine, data analysis tools, etc. are enhanced and modified to provide RF-specific design tools. In an embedded methodology, the simulation data base and cell libraries are the same, regardless of the type of application (RF or IC). Also, a single analog simulator greatly reduces the device modeling problems. It is easier to implement a new model, easier to maintain existing models, and there is never an issue with model consistency. The other main advantage of this system is that it provides an RF extension to an IC system. This makes the RF tools easier to use by IC designers, and provides a familiar environment for RF designers. Embedded tools allow access to the ASIC vendor's proprietary device models and predesigned cells. Since most IC systems already provide mixed signal (analog and digital) design capabilities, a system including low frequency analog, digital, and RF subsystems can be completely designed and simulated.

## Major Components of the IC System

*Front-end IC design tools* — Designs on an IC design system generally start from a graphical user interface for schematic entry. Automated tools help the designers to select the best transistor type, size, and geometry for a particular design objective. Other tools will perform schematic capture, generate a netlist, and verify that the schematic follows design rules. The analog simulator can then be invoked for routine AC and DC analysis. Sensitivity analysis, parametric analysis, pole/zero analysis, and noise analysis may also be run. Sensitivity analysis allows the designer to observe the sensitivity of a performance objective to a circuit parameter (e.g., an element value or transistor size). Parametric analysis is a powerful tool for observing the variation of performance objectives as a circuit parameter varies (e.g., temperature or element values) within a specified range. Pole/zero analysis determines the poles and zeroes of the transfer function of the system; useful for identifying and correcting an undesired frequency response. Statistical simulations, using process-dependent parameters and correlation coefficients enables designers to observe the statistical variation of the design objectives and predict manufacturing yield.

For large circuits, a whole chip simulation at the transistor level may be very CPU intensive. In such cases, IC designers replace selected blocks of the circuit with their macro or behavioral models. This modular approach significantly reduces the simulation time while retaining a comparable level of accuracy. Choices from very fast but less accurate to slower but more accurate allow the user to select the appropriate speed-accuracy requirements for simulation. Mixed signal simulation can uncover circuit design errors and oversights. Methodologies range from gluing a digital simulator to an analog simulator, to building a native logic simulator inside an analog simulator. Each of these methods has a particular domain of applicability and effectiveness. Some design systems (e.g., Harris Fastrack) have included many different methodologies in their mixed signal simulator.

A unified preprocessor generates appropriate netlists for different simulators without any need for user intervention. A unified graphical post processing tool displays the results regardless of the type of simulation, simulator, or analysis. A full-function waveform calculator also allows the user to perform mathematical processing of the results.

*Backend IC design tools* — The layout of the active and passive devices are automatically generated from the schematic. Many predesigned cells have parametric layouts, and the entire layout of the cell is automatically synthesized. Even though analog routing of the layout may still be manual, many automated and dynamically operating tools are provided to simplify this task. Another tool allows real time verification between the schematic and the layout. An electromigration checking tool can examine selected nets and point out any electromigration design rule violations.

The designer can then proceed to use the layout parasitic extraction tool to determine the parasitic resistances and capacitances of the routing lines. A layout driven netlist which includes the pri-

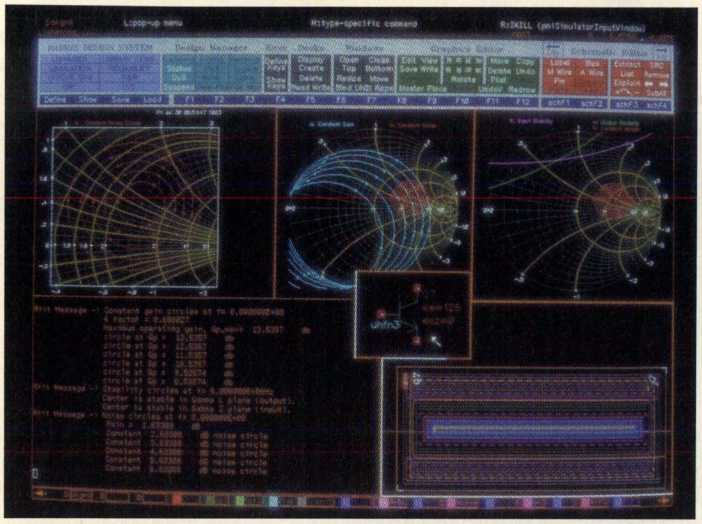

**Figure 2. IC transistor simulation with noise, power gain and stability circles.**

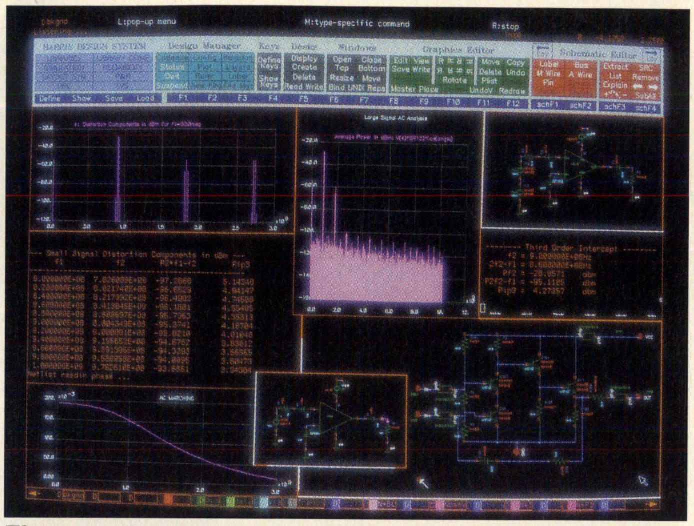

**Figure 3. Low noise amplifier, with results of large-signal and small-signal distortion analysis.**

mary and parasitic elements is generated. A post layout simulation, using this netlist, can determine the effect of the layout parasitics and point out any performance degradation.

### Components of the Embedded RF System

Even though execution of certain RF functions is possible through IC design practices and CAD tools, it is neither easy nor efficient. Also, RF designers typically use different design goals and different data representation, analysis, and interpretation than IC designers. Therefore it is essential to provide an RF design tool set within any design system for silicon ASICs for RF applications. Many RF-specific design tools use the same mathematical basis and numerical algorithms that exist in IC tools, although the procedure to extract the required data can be different. In general, the user interface, analog simulation engine, and the data representation routines in the IC design system are enhanced and modified to provide an RF design environment and RF-specific design tools. The 3.6 release of Harris Fastrack is the first IC design system to provide a large set of embedded RF design tools.

*S parameter measurement* — S parameter measurement enables the designer to determine all four S parameters for a transistor or a two port network. The system establishes the proper loading and excitation (and auto-biases single transistors), runs multiple simulations and processes the results to determine the S parameters. Once the S parameters are determined they can be printed or plotted to the screen or to a file in any user specified format. The S parameters are also used to generate constant power gain and stability circles, plot and print K factor, input and output VSWR, and input and output impedances (Figure 1).

*S parameter representation* — Frequently, RF designers require the capability to represent a two-port network by its S parameters in a tabular form. In the embedded RF tools, a generic macromodel provides the capability to represent the external behavior of any two-port network using an S parameter file. This file can either be generated by the S parameter measurement utility described above, or from a data sheet for other vendors' parts. The modeling is based on the equivalent two port Y parameters which in turn are generated from the complex tabular S parameters. For AC small signal analysis, the interpolated values of the tabular Y parameters are used.

*Noise parameter measurement* — RF designs are frequently driven by noise specifications. Even though IC and RF designers basically use the same noise models and AC small signal noise analysis, the interpretation of the results, the type of requested information, and the noise information processing are different. RF designers are typically interested in noise figure at a given frequency, minimum noise figure, noise resistance, and optimum reflection coefficients versus frequency. The noise parameter measurement capability in the embedded RF tools enables the designer to measure RF specific noise parameters for a single transistor or a two port network. As with S parameter measurement the system will automatically connect a bias circuit to a single transistor. Noise figure, minimum noise figure, and noise resistance versus frequency can be plotted or printed to screen or file. Optimum reflection coefficient can be printed or plotted on a rectangular plot or a Smith chart. The user can also plot the constant noise circles on a Smith chart. In addition, constant noise circles can be overlaid on the constant gain or stability circles to allow for a visual design trade off between gain, stability, and noise performance of a circuit (Figure 2).

*Large signal AC analysis* — This analysis enables the RF designer to observe the nonlinear performance of a circuit in the frequency domain. It can be used for single tone circuits to observe harmonic distortion or for multitone circuits to observe the intermodulation products. The third order intercept point is automatically calculated from the frequency spectrum. The power spectral density can be plotted (in a spectrum analyzer type bar plot) or printed to screen or file. The system determines the desired frequency spectrum by invoking an internally controlled and automated nonlinear transient analysis followed by a Fast Fourier Transform (FFT). The transient analysis and FFT are designed to internally control each other for optimum accuracy and efficiency. While most FFT based methods only offer about 60-80 db of dynamic range, this new method offers around 160 db of

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dynamic range. This methodology is totally different from Harmonic Balance, although both methods use a combination of FFT and nonlinear transient analysis. Harmonic Balance based simulators are more suitable for smaller circuits that either take a long time to reach steady state or when the beat frequency in a multi-tone application is several orders of magnitude smaller than the tones. Large signal AC analysis in the embedded RF tools is as efficient as typical SPICE transient analysis for larger circuits.

*Small signal distortion* — This analysis enables the designer to observe the nonlinearity and distortion for quasi-linear circuits. Quasi-linear circuits are a class of circuits that are ideally linear, but exhibit some undesired nonlinearities. Small signal distortion analysis exploits this nonlinearity by determining the undesired effects of each nonlinear component in the circuit on an output load. This analysis uses linear transfer functions and, consequently, it is significantly faster than FFT based analyses. At a given fundamental frequency, the frequency spectrum, comprised of the major intermodulation components, can be plotted or printed. The fundamental frequency can be varied over a specified range, and the intermodulation components can be plotted or printed versus the fundamental frequency. The third order intercept point can also be automatically calculated (Figure 3).

### Other Design Tools

The following describes some other RF design tools that can be supported inside an IC design system:

*Optimization* — A circuit optimizer is an invaluable tool that determines the optimum solution to design tradeoffs. Typically, circuit optimizers are ineffective (very slow) on large IC circuits and are mainly used at the cell development stage. Because of the small size and small signal AC optimization objective functions (inexpensive in terms of simulation CPU time), RF designs are ideal candidates for circuit optimization.

*Package modeling and simulation* — The prediction of the effects of the package and bond-wire parasitics on the electrical performance of the circuit is essential for RF designs. The designers must to perform pre and post layout simulations on chip level schematics (as opposed to the traditional die level schematic.

*Electromagnetic simulation* — Discrete designers typically use electromagnetic simulators to analyze and model board level RF layout structures. These simulators could also be used to accurately model on-die structures such as spiral inductors, transmission lines, bends, etc.

*System level simulation* — System level simulation is an essential requirement for top-down design of large RF systems. Parametric macro/behavioral models can represent RF subsystems, and the user can choose from generic building blocks such as LNAs, mixers, filters, comparators, and delay blocks, then customize them by entering the desired RF specifications such as noise figure, VSWR, S parameters, and filter specs. Any number of these blocks in any arrangement (including topologies with local and global feedback) can be used for a complete functional simulation of the entire RF system.

### RF Designs in Silicon ASIC — The Challenges

Currently, RF designs in silicon ASICs require iterating between RF and IC design tools. Many RF-specific design steps are simply not available in IC design tools. On the other hand, RF design tools do not perform IC statistical simulation and yield prediction, parasitic extraction, layout-driven simulation, electrical and physical design rule checking, and layout-to-schematic verification. RF designers typically must analyze the steady state response of a circuit to multi tone sinusoidal excitations in either the large or small signal domains. IC designers, on the contrary, are more interested in the transient and small signal AC behavior of a circuit. While IC designers are interested in propagation delay and open loop gain of a circuit, RF designers look for noise figure and the third order intercept in a large signal AC application.

Two other factors further complicate RF designs in silicon ASICs. First, generally, there is a lack of high frequency design knowledge in the IC design community and a lack of IC fabrication process-dependent design knowledge in the RF design community. Second, since breadboarding is not possible in ASICs and redesign after fabrication is very costly, accurate prediction of performance requires robust simulation capabilities and accurate device models.

*Transistor selection* — To optimize a design, the most appropriate transistors and their ideal sizes must be selected. When customizable transistors are available, IC design tools are needed to

**Advantages**

Small size
System solution
Lower cost for high volume
Higher performance
Reduced package parasitics
Negligible interconnect parasitics
Highly accurate device matching
Predesigned cell library
Variable geometry devices
Combined IC and RF design tools
Accurate statistical simulation
Post layout simulation

**Drawbacks**

Lack of RF knowledge in IC community
Lack of silicon knowledge in RF community
New design paradigm
Restriction to one vendor
High absolute error tolerance
Lack of large passive elements
Passive element parasitics

**Table 1: Comparison of RF designs in silicon ASICs versus discrete based designs.**

select a transistor that meets proper the low frequency and DC requirements. Automated tools let the designers optimize transistor size for specified DC requirements such as, minimum geometry for non-saturation, specified values for RC and RB etc. Other automated tools can determine the optimum operating conditions. An example of such automated tools are Device Design and Device Test in Fastrack. On the other hand, RF design tools are used to make tradeoffs between transistor noise figure and gain, to design impedance matching networks, and to design for intermodulation performance. Using separate RF and IC design systems for selection of a transistor would require many iterations between RF and IC tools.

*Statistical and parametric simulations* — After the initial selection of the topology of the circuit and transistor sizes, statistical and parametric simulations are required to verify the performance of the circuit. Statistical simulations are needed to investigate the effects of process variations on the design objectives. Parametric simulations allow designers to vary the device sizes and/or element values and observe the corresponding effects on design objectives. The design objectives can be RF-specific design goals as well as DC and low frequency design goals. IC design systems provide process dependent statistical models and powerful statistical simulation capability. RF design systems, on the other hand, support AC and S parameter analyses but lack true statistical simulation capability. They typically provide a Monte Carlo simulation, but this may lead to incorrect and misleading results because the models are not process dependent and the statistical correlation between the device model parameters is ignored. As a result, for cases like statistical distribution of third-order intercept point for a mixer, iterating back and forth between the design tools will not provide the required information.

*Post-layout simulation* — After the circuit layout is complete, the IC design system can extract the layout parasitics (resistors and capacitors) and back annotate them into the schematic and augment the netlist. This modified netlist is the true representation of the circuit, on which the final performance verification must be based. If the IC design system does not provide RF-specific simulation tools, the designer has to port the netlist to another system to perform an RF-specific simulation. In such cases, investigating the effects of layout parasitics on a circuit's performance criteria, predicted by an RF-specific simulation tool, requires switching between the IC and RF design systems.

*Device Models* — RF designers often use generic Spice model parameters supplied by the component vendors. This allows the use of many simulators with the same device models. However, IC foundries typically use process based proprietary device models for the intricate and often critical behavior of devices at high frequency. If the ASIC vendor does not supply all the RF tools needed to complete the design, a transfer of the design to another simulator is necessary. The designers have to use a simplified version of the vendor's proprietary models and sacrifice accuracy. An exception to this rule is when linear simulators are used, which can use S parameters to model nonlinear devices. The S parameter models can either be supplied by the vendor or generated from the simulation.

*Libraries* — ASIC vendors typically provide a large library of predesigned cells including mixers, multipliers, opamps, comparators, voltage references, analog-to-digital, and digital-to-analog converter, sample and hold circuits, and a full set of logic gates. These predesigned cells are often customizable to meet the user's specific performance criteria. Predesigned cells can dramatically simplify and increase the level of integration for RF designs, a major factor that should be considered by RF designers using discrete components. However, using an ASIC vendor's predesigned cells increases the dependency of the overall performance of the RF system on the vendor's cell library and device models. For consistent results and documentation, the entire design should be completed on one system.

## RF Designs in Silicon ASICs — The Drawbacks

Even though, the paradigm shift from discrete based designs to silicon ASICs offers many advantages, it is not without restrictions and shortcomings. Some of the advantages of discrete based designs may not be available. In this new paradigm, many traditional design practices which were based on the merits of discrete devices need to be abandoned.

*Restriction to one vendor* — Traditionally, discrete designers have used passive and active devices from many different vendors. For example, a few low noise devices from one vendor can be combined with a few high power devices from another vendor in an RF subsystem to satisfy noise and power specs. With silicon ASICs, the designer is restricted to one vendor's offerings.

*Absolute error tolerance* — Discrete passive elements with absolute error tolerances of 1 percent or less are commonplace. In silicon ASICs, even though the relative error tolerances (matching) can be kept to about 1 percent, low absolute error tolerances for passive elements are hard to achieve. Tolerances of 10 percent or higher are not uncommon for IC resistors, capacitors, and inductors. High precision passive elements can be achieved by laser trimming after fabrication, but this increases the price of the ASIC.

*Large passive elements* — The inability of silicon ASICs to offer economical high capacitance, resistance, and inductance per unit area precludes the use of on-chip large passive elements. The per unit capacitance and resistance (NiCr) are typically around 0.1 pF/mil and 125

ohms/mil respectively. Because of the these restrictions and the high cost of real estate on a die, large capacitors, inductors, and resistors have to go off-chip.

*Passive element parasitics* — Ideal passive elements, although routinely used in the conception phase of a design, are impossible to produce. Real passive elements contain parasitic resistors, capacitors, and inductors. As the frequency of operation increases, the effect of the parasitic elements will be more apparent and in many cases can dominate the element itself. This effect is drastically more prevalent in IC devices than in discrete devices. For example, on-chip inductors (spiral) cannot be produced with high Q.

The above restrictions and shortcomings necessitate new thinking. Instead of trying to find IC solutions that can be used to realize conventional discrete design procedures, a new design methodology has to be adopted. In the new methodology, circuit topology has to has to follow the device selection. This is contrary to discrete design in which the device selection follows the circuit topology. Discrete based designers typically conceive the topology, determine the requirements for the passive and active elements and then search for them in vendor's data books. Since this design practice is not possible with silicon ASICs, the designers have to determine the capabilities and performance limits of the available devices and then select a topology that can use the given devices to meet the design objectives.

## RF Designs in Silicon ASICs — The Advantages

*System level advantages* — The single most important advantage of silicon ASICs over discrete designs is reduced size. As RF systems become more sophisticated, the element count increases and the physical size becomes a critical issue. In many portable wireless applications, this fact alone can justify using ASICs. The other important system level advantages are reduced time-to-market and lower cost for high volume products. The predesigned and parametric cells in an IC vendor's library can dramatically reduce the overall design time. As the production volume of an IC product increases, the per unit price drops significantly, at a rate much higher than for discretes.

*Packaging and interconnects* — Performance is another important advantage of ASICs over discrete designs. By eliminating most individual packages, the performance degradation due to package parasitics is minimized. Equally important is the effect of interconnect parasitics. The element-to-element interconnect in an IC is several orders of magnitude shorter than in p.c. boards (microns versus tenths of inches).

*IC specific design advantages* — Integrated circuits offer specific design advantages such as highly accurate device matching and variable geometry devices. IC fabrication technologies can offer relative error tolerances of 0.1 percent or lower for high performance differential applications. Some IC foundries, including Harris Semiconductor, offer variable geometry devices. RF designers can capitalize on this feature to select the appropriate size and geometry for devices to optimize performance.

*Design tools* — Another significant advantage of ASIC based designs is the robustness and completeness of the design tools. Two powerful IC design tools, typically unknown to RF designers, are the statistical (with correlated fab data) and post layout simulation tools. Combining traditional IC tools with embedded RF design tools provides a complete front-to-back system design environment and offers a powerful circuit performance prediction capability.

## Conclusions

A major step in the evolution of RF designs is emerging. The recent introduction of IC technologies offering high frequency transistors with $f_t$ in the vicinity of 10 GHz has opened new opportunities for higher integration of wireless communication systems. Fast silicon IC devices make possible the integration of many RF subsystems on a single die and offer a total solution to mixed frequency (low frequency and RF) and mixed signal systems. Migrating from discrete RF design to RF designs utilizing silicon ASICs reduces the overall size of the system, minimizes the effect of packaging, and can speed up the design cycle. As more of the ASIC vendor's predesigned cells are used in the RF system, the overall cost decreases and the time-to market improves.

To realize this opportunity, IC design systems have to be enhanced to accommodate RF-specific design tools. Reliable and accurate tools for predicting the IC performance before fabrication is essential. Since RF and IC designers are forced to work in both domains, it is essential to provide a unified design system where each designer will find their familiar environment as well as the ability to traverse to the other domain. In the years to come, we will see widespread use of silicon ASICs for RF systems. This will encourage the IC vendors to invest more heavily in the fabrication process, simulation models, and design tools. The performance of IC devices are far from their physical limits and many advancements in providing higher ft, lower noise figure, and distortion are forthcoming.

As the frequency of operation increases, there is a need for more accurate models in general, and physically based models in particular, for active and passive devices and interconnects. At the same time, as RF systems grow in complexity, the simulation efficiency of the models becomes a necessity. This requires a modeling methodology that can accurately translate physically based models (which are typically very complex) to efficient circuit simulator compatible models. The growing market for RF designs in silicon ASICs will also motivate the traditionally separate RF and IC CAD tool vendors to strive for a combined system solution. ***RF***

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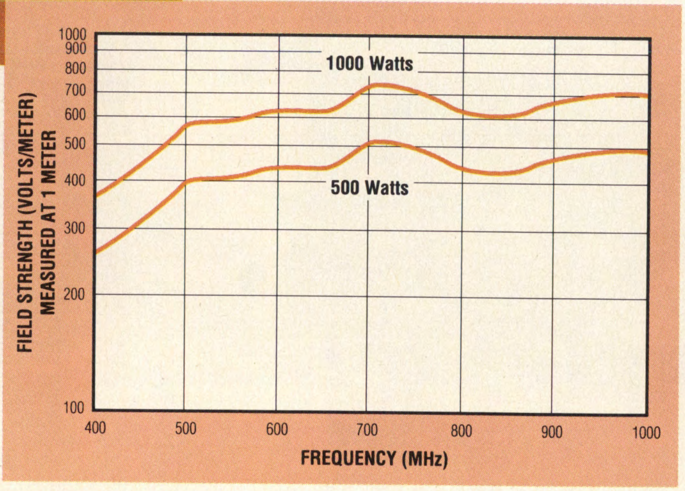


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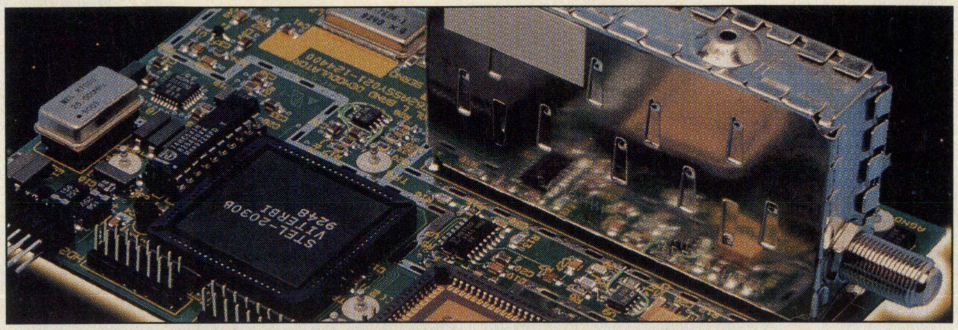
# Manufacturing Considerations for the Design of RF Products

*By Robert L. Barron and William J. Choe*
*Stanford Telecom, MQA Division*

*Current product trends require RF circuits in higher volumes, smaller packages, lighter weights and lower prices than ever before. This article provides designers with an understanding of manufacturing principles to help address product design issues in today's demanding markets.*



*Stanford Telecom's STEL-9236 VSAT receiver module is an example of manufacturing and engineering working together in concert.*

An RF design engineer who has spent his career at defense contractor firms commented, "RF design engineers can not take manufacturing concerns into consideration, as it is somewhat trivial compared to the design task and may compromise the functionality of the design." What this engineer has not recognized is that commercial markets are demanding RF circuits in high volume, with repeatable steps in their manufacturing process. Not only that, digital designs are running at higher speeds and therefore must learn to adapt RF design techniques to support their digital designs. In both cases, successful products demand attention to more than just the ideal circuit design.

These demands necessitate using Surface Mount Technology (SMT) and understanding the manufacturing processes to produce the product. The processes include PCB layout, materials analysis, materials procurement, assembly and test. Although RF production remains plagued by unautomated, manual labor intensive processes, good design practices can minimize them. The purpose of this article is to provide an understanding of the necessary design practices.

While our domestic corporations earnestly realign themselves with current market trends and declining government spending, two markets are clearly positioned for growth: Contract Manufacturing and Telecommunications. Contract manufacturing continues to grow after an explosive surge in the 80s, due to personal computers. Simultaneously, many high-tech firms are focusing on their core business (primarily engineering and marketing), relying on contract manufacturers to handle capital expenditures and workforce fluctuations.

Telecommunications is being heralded as the "market of the 90s". Spread spectrum applications such as CDMA telephones and GPS navigation systems have recently emerged from the military world to offer a major advance in personal mobile communications. These technologies, coupled with relaxed FCC regulations, have produced an explosion of new communication products.

In the 80s, new businesses arose in personal computer markets. Today, exciting new businesses are being established to address wireless markets such as wireless LAN, wireless PBX, vehicular tracking, cordless phones, Intelligent Vehicle Highway Systems, and utility meter reading. It is essential for these new firms to select competent contract manufacturers to support their success in both concurrent engineering and production.

## Design Considerations

Designing and manufacturing products for RF and high speed digital technologies is best accomplished by possessing an awareness of sound design practices and manufacturing experience. Sound design practices include repeatable performance characteristics, the elimination of adjustable components, and proper selection of component tolerances. RF designs present manufacturing issues not apparent in lower speed digital and analog assemblies. Some of the major issues that should be considered during the design phase are listed in Table 1.

In contrast to the past, RF assemblies are becoming much more constrained by size and weight. This added dimension for RF designers has necessitated learning the proper utilization of SMT components in their designs. An appreciation for SMT utilization on PCBs was gained by the digital designers during the last decade. Table 2 describes issues related to SMT manufacturing processes. Simultaneously, RF designers must learn digital design techniques, and digital designers have to understand RF design techniques to successfully use SMT high speed digital circuits. Manufacturers will have to combine RF and digital designs on a single PCB as both technologies will be used in emerging products.

While this article presents SMT as a

- PCB layout and substrate selection
- Grounding and shielding techniques
- Component specifications for compatibility with automated assembly equipment
- Thermal management problems due to power dissipation especially in high frequency applications
- Cleaning processes

**Table 1. Manufacturing issues related to RF designs.**

# FILTERS

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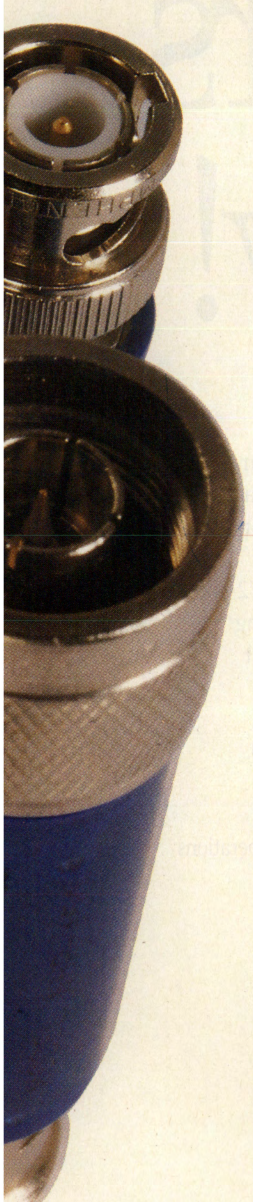
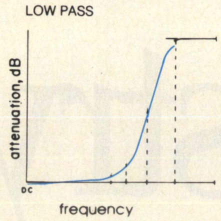
## lowpass, highpass, bandpass

• less than 1dB insertion loss • greater than 40dB stopband rejection • surface-mount • BNC, Type N, SMA available • 5-section, 30dB/octave rolloff • VSWR less than 1.7 (typ) • rugged hermetically-sealed pin models • constant phase • meets MIL-STD-202 tests • over 100 off-the-shelf models • immediate delivery



### low pass, Plug-in, dc to 1200MHz




| Model No. | Passband MHz loss < 1dB | Stopband, MHz loss > 20dB | Stopband, MHz loss > 40dB | Model No. | Passband MHz loss < 1dB | Stopband, MHz loss > 20dB | Stopband, MHz loss > 40dB |
|---|---|---|---|---|---|---|---|
| ★LP-5 | DC-5 | 8-10 | 10-200 | ★LP-250 | DC-225 | 320-400 | 400-1200 |
| ★LP-10.7 | DC-11 | 19-24 | 24-200 | ★LP-300 | DC-270 | 410-550 | 550-1200 |
| ★LP-21.4 | DC-22 | 32-41 | 41-200 | ★LP-450 | DC-400 | 580-750 | 750-1800 |
| ★LP-30 | DC-32 | 47-61 | 61-200 | ★LP-550 | DC-520 | 750-920 | 920-2000 |
| ★LP-50 | DC-48 | 70-90 | 90-200 | ★LP-600 | DC-680 | 840-1120 | 1120-2000 |
| ★LP-70 | DC-60 | 90-117 | 117-300 | ★LP-750 | DC-700 | 1000-1300 | 1300-2000 |
| ★P-90 | DC-81 | 121-137 | 167-400 | ★LP-800 | DC-720 | 1080-1400 | 1400-2000 |
| ★LP-100 | DC-98 | 146-189 | 189-400 | ★LP-850 | DC-760 | 1100-1400 | 1400-2000 |
| ★LP-150 | DC-140 | 210-300 | 300-600 | ★LP-1000 | DC-900 | 1340-1750 | 1750-2000 |
| ★LP-200 | DC-190 | 290-390 | 390-800 | ★LP-1200 | DC-1000 | 1620-2100 | 2100-2500 |

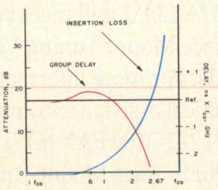
Price, (1-9 qty), all models: plug-in $14.95, BNC $32.95, SMA $34.95, Type N $35.95

#### Surface-mount, dc to 570MHz

| | | | | | | | |
|---|---|---|---|---|---|---|---|
| SCLF-21.4 | DC-22 | 32-41 | 41-200 | SCLF-190 | DC-190 | 290-390 | 390-800 |
| SCLF-30 | DC-30 | 47-61 | 61-200 | SCLF-380 | DC-380 | 580-750 | 750-1800 |
| SCLF-45 | DC-45 | 70-90 | 90-200 | SCLF-420 | DC-420 | 750-920 | 920-2000 |
| SCLF-135 | DC-135 | 210-300 | 300-600 | | | | |

Price, (1-9 qty), all models: $11.45

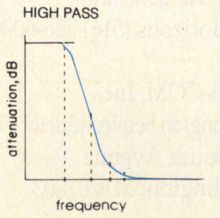
#### Flat Time Delay, dc to 1870MHz




| Model No. | Passband MHz loss < 1.2dB | Stopband MHz loss > 10dB | Stopband MHz loss > 20dB | VSWR Freq. Range, DC thru 0.2fco X̄ | VSWR Freq. Range, DC thru 0.6fco X̄ | Group Delay Variations, ns Freq. Range, DC thru fco X̄ | 2fco X̄ | 2.67fco X̄ |
|---|---|---|---|---|---|---|---|---|
| ★BLP-39 | DC-23 | 78-117 | 117 | 1.3:1 | 2.3:1 | 0.7 | 4.0 | 5.0 |
| ★BLP-117 | DC-65 | 234-312 | 312 | 1.3:1 | 2.4:1 | 0.35 | 1.4 | 1.9 |
| ★BLP-156 | DC-94 | 312-416 | 416 | 0.3:1 | 1.1:1 | 0.3 | 1.1 | 1.5 |
| ★BLP-200 | DC-120 | 400-534 | 534 | 1.6:1 | 1.9:1 | 0.4 | 1.3 | 1.6 |
| ★BLP-300 | DC-180 | 600-801 | 801 | 1.25:1 | 2.2:1 | 0.2 | 0.6 | 0.8 |
| ★BLP-467 | DC-280 | 934-1246 | 1246 | 1.25:1 | 2.2:1 | 0.15 | 0.4 | 0.55 |
| ▲BLP-933 | DC-560 | 1866-2490 | 2490 | 1.3:1 | 2.2:1 | 0.09 | 0.2 | 0.28 |
| ▲BLP-1870 | DC-850 | 3740-6000 | 5000 | 1.45:1 | 2.9:1 | 0.05 | 0.1 | 0.15 |

Price, (1-9 qty), all models: plug-in $19.95, BNC $36.95, SMA $38.95, Type N $39.95
NOTE: ▲: -933 and -1870 only with connectors, at additional $2 above other connector models.

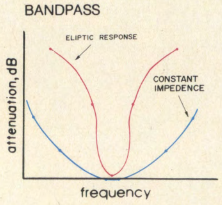
### high pass, Plug-in, 27.5 to 2200MHz




| Model No. | Stopband MHz loss < 40dB | Stopband MHz loss < 20dB | Passband, MHz loss < 1dB | VSWR Pass-band Typ. | Model No. | Stopband MHz loss < 40dB | Stopband MHz loss < 20dB | Passband, MHz loss < 1dB | VSWR Pass-band Typ. |
|---|---|---|---|---|---|---|---|---|---|
| ★HP-25 | DC-13 | 13-19 | 27.5-200 | 1.8:1 | ★HP-400 | DC-210 | 210-290 | 395-1600 | 1.7:1 |
| ★HP-50 | DC-20 | 20-26 | 41-200 | 1.5:1 | ★HP-500 | DC-280 | 280-365 | 500-1600 | 1.8:1 |
| ★HP-100 | DC-40 | 40-55 | 90-400 | 1.8:1 | ★HP-600 | DC-350 | 350-440 | 600-1600 | 2.0:1 |
| ★HP-150 | DC-70 | 70-95 | 133-600 | 1.8:1 | ★HP-700 | DC-400 | 400-520 | 700-1800 | 1.6:1 |
| ★HP-175 | DC-70 | 70-105 | 160-800 | 1.5:1 | ★HP-800 | DC-445 | 445-570 | 780-2000 | 2.1:1 |
| ★HP-200 | DC-90 | 90-116 | 185-800 | 1.6:1 | ★HP-900 | DC-520 | 520-660 | 910-2100 | 1.8:1 |
| ★HP-250 | DC-100 | 100-150 | 225-1200 | 1.3:1 | ★HP-1000 | DC-550 | 550-720 | 1000-2200 | 1.9:1 |
| ★HP-300 | DC-145 | 145-170 | 290-1200 | 1.7:1 | | | | | |

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

### bandpass, Elliptic Response, 10.7 to 70MHz




| Model No. | Center Freq. (MHz) | Passband I.L. 1.5 dB Max. (MHz) | 3 dB Bandwidth Typ. (MHz) | Stopbands I.L. > 20dB at MHz | Stopbands I.L. > 35dB at MHz |
|---|---|---|---|---|---|
| ★BP-10.7 | 10.7 | 9.6-11.5 | 8.9-12.7 | 7.5 & 15 | 0.6 & 50-1000 |
| ★BP-21.4 | 21.4 | 19.2-23.6 | 17.9-25.3 | 15.5 & 29 | 3.0 & 80-1000 |
| ★BP-30 | 30.0 | 27.0-33.0 | 25-35 | 22 & 40 | 3.2 & 99-1000 |
| ★BP-60 | 60.0 | 55.0-67.0 | 49.5-70.5 | 44 & 79 | 4.6 & 190-1000 |
| ★BP-70 | 70.0 | 63.0-77.0 | 68.0-82.0 | 51 & 94 | 6.0 & 193-1000 |

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

#### Constant Impedance, 21.4 to 70MHz

| Model No. | Center Freq. MHz | Passband MHz loss < 1dB | Stopband loss > 20dB at MHz | VSWR 1.3:1 Total Band MHz |
|---|---|---|---|---|
| ★IF-21.4 | 21.4 | 18-25 | 1.3 & 150 | DC-220 |
| ★IF-30 | 30 | 25-35 | 1.9 & 210 | DC-330 |
| ★IF-40 | 42 | 35-49 | 2.6 & 300 | DC-400 |
| ★IF-50 | 50 | 41-58 | 3.1 & 350 | DC-440 |
| ★IF-60 | 60 | 50-70 | 3.8 & 400 | DC-500 |
| ★IF-70 | 70 | 58-82 | 4.4 & 490 | DC-550 |

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

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

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necessity, it must be understood that SMT is a more precise science than conventional through-hole technology. SMT defects are more difficult to observe and repair. Production and Q.C. departments must closely control processes and perform audits to reduce rejects and rework. In addition, careful matching of stress related design criteria, particularly for ceramic components, is required when implementing SMT. Coefficients of thermal expansion of components and substrates, plus PCB mechanical bending must be considered when utilizing SMT components.

### Component Selection

Component selection is key to achieving low-cost and process efficient manufacturing. It is always desirable to have a minimum parts count. However this task is more easily attainable with digital and mechanical parts than with RF components. RF component selection becomes more a task of reliability, availability, and ease of manufacturing. Since 80 percent of rework can easily be caused by 20 percent of the components, rigorous statistical analysis of the components must be performed during the design phase. For example, performing pareto analysis to identify major problems coupled with vibration testing during pre-production facilitates proper RF component selection.

In the pursuit of optimum circuit performance, designers may inadvertently compromise cost, time-to-market, and production yields by specifying stringent component tolerances. Engineers can optimize a nominal response for performance, but to increase the manufacturing yield the engineer must intuitively replace parts or change component tolerances. Designers often try to meet specifications by using tighter toleranced components, so the manufactured response falls symmetrically about the nominal design. In this way the true statistics of the problem are ignored, and the resulting design may not be cost effective. Cobler [1] points out that it is the statistical response, not the nominal response, which determines the final success of a design, since it represents the actual manufactured performance.

An alarming attribute of component selection is the availability of the part. In a desire to have the optimum state-of-the-art circuit or to have a SMT device, the designer often chooses a part that is provided by a salesperson. Yet when the engineer's product is released and purchasing goes to buy the part, it is discovered that the part is not in production. A painful example of this occured when our firm, MQA, received a contract to build several thousand RF receivers in time for the Christmas season. When the MQA buyer tried to purchase the customer-approved filter, the buyer was given a 26 week leadtime. The part had not yet been scheduled for production. In this case, both MQA and its customer lost the order since the "Christmas window" was missed.

- Thermal Coefficient of Expansion between the component and substrate
- Reflow process controls for thermal shock and proper solderability
- Automated parts placement techniques
- PCB layout principles for proper land patterns and sizes
- Process controls for solder paste and adhesive applications

**Table 2. Principal issues for SMT manufacturing.**

- Temperature profile for the reflow process
- Land patterns and pad dimensions for layout on the PCB
- Packaging specifications for allowing automatic feed to the pick & place machine
- Mounting specifications when components may have RFI/EMI leakage
- Mechanical part specifications for compatibility with pick & place machines

**Table 3. Component-related issues for SMT manufacturing.**

Assuredly, extensive use of surface-mount components drastically reduces assembly time while maximizing board density and providing the RF performance required. SMT devices can be automatically placed and either reflow or wave soldered with the rest of the RF components. In specifying SMT devices, the design engineer should work with manufacturing to determine various

| Substrate | Dielectric Constant | Dissipation Factor | CTE (ppm/°C) x-y | CTE (ppm/°C) z |
|---|---|---|---|---|
| FR4 | 4.8 | 0.022 | 12-16 | 80 |
| Polymide/glass | 4.5 | 0.01 | 12-14 | 60 |
| Alumina | 9.6 | 0.0001 | 6.2 | 6.2 |
| PTFE (GX-woven glass filled) | 2.4 ±0.04 | 0.0019 | 15 | 200 |
| RT/duroid 5880 | 2.2 | — | 2-3 | 28.3 |
| RT/duroid 6002 | 2.94 ±0.04 | 0.0012 | 16 | 24 |

**Table 4. Properties of common substrates.**

parameters as illustrated in Table 3.

Generally, the critical components in a SMT assembly are the chip capacitors and resistors. Other passive and active components, although surface mountable, generally have leads or electrodes which are compliant. Ceramic chips are leadless, and have a poor ability to withstand thermal shock. Thermal shock may result during soldering or baking processes. Ceramic chip failures resulting from excessive thermal shock may include: micro cracks in the ceramic, electrical short, and insulation resistance degradation in accelerated life test conditions. In addition, the recommended soldering process time-temperature profile for components should be followed. The reliability of any surface-mount attachment is directly dependent on the board type, solder type, and mounting procedure utilized.

### PCB Considerations

Multilayer microwave PCBs are emerging rapidly to meet the needs of large volume, low cost, and reliable RF and microwave systems. Advancements in PCB fabrication processes have made complex high frequency designs more practical. By adhering to some basic design guidelines for high frequency PCBs, design, the evolution of substrates have provided opportunities for improving system reliabiliers are able to lay out a more manufacturable and reliable printed circuit assembly. Benefits of a well designed RF PCB include improved electrical performance and mechanical reliability, and smaller PCBs at reduced costs.

In the past, microwave board designs have been limited to single microstrip and stripline fabrications. The resulting hardware has often been heavy and bulky because of the need for multiple RF and digital circuits. New laminates have opened up new design options for true RF multilayer designs including the integration of digital functions.

These exotic substrates have favorable mechanical and electrical properties, making demanding RF/microwave multilayer PCBs more feasible. In the past, ceramic or glass reinforced polytetraflourethylene (PTFE, also known by DuPont's TEFLON® brand name) substrates were used to meet stringent electrical requirements. Although these materials have relatively low dielectric constants, they have poor mechanical properties. The coefficient of thermal expansion (CTE) of PTFE is especially high in the z-axis or the thickness plane. If the CTE of the package material (i.e. ceramic chip caps) is not closely matched with the board's, then solder joint reliability problems such as fatigue and creep may occur. Table 4 illustrates the properties of some frequently specified materials.

The RT/duroid® microwave laminates from Rogers have superior mechanical and electrical properties compared to conventional digital multilayer materials. Material technology has provided a variety of substrate options that can be cost effectively selected for unique mechanical and electrical characteristics.

Proper layout of a PCB requires a thorough understanding of your board fabricator's capabilities. Communicating innovative ideas in the preliminary stages can prevent fabrication problems. While the PCB fabrication industry is constantly improving its etching and drilling processes, specifying realistic trace spaces and hole sizes will avoid unnecessary costs. It is worth noting that many board fabricators are etching 0.005-inch traces and spaces and drilling 0.015-inch holes at no extra cost.

RF and microwave PCB's have inherent requirements that need to be addressed to assure a reliable and functional circuit in both ambient and extreme environments. For RF circuits demanding higher current levels, a heavier copper weight needs to be specified. Typically, copper is available in 2, 1, or 1/2 ounces per square foot. These weights correspond to a thickness of .0028, .0014, and .0007 inch, respectively. There are tables which illustrate the minimum trace widths recommended for current densities at different temperatures [4]. The Institute for Interconnecting and Packaging Elec-

## DFM Case Study: a VSAT Receiver

*By Hatch Graham*
*ASIC & Custom Products Division, Stanford Telecom*

An excellent example of an industry limited by production costs due to manufacturing and integration is the VSAT (Very Small Aperture Terminal) market. VSATs provide data, voice and video over satellite transponders without the need for land based wiring infrastructures. A typical application today might include one-way transmission of background music to retail stores, hotels and other enterprises. Or, retail stores can transfer point-of-sale information in real time to central processing sites for inventory management, check and credit card verification, and financial reporting. Finally, VSAT technology is seeing growing markets in third world areas that are without communication infrastructures.

Prior to the rapid expansion of applications such as VSAT, satellite equipment pricing was secondary to performance because the users were government or internationally sponsored, with large budgets. Currently, the installed VSAT market is 160,000 sites with an annual growth of 30 percent. VSAT is a good example where the fundamentals of manufacturing and DFM can achieve both reliability and cost effectiveness.

Traditionally, satellite receivers have required discrete analog and digital circuits on a number of boards. Difficulty in maintaining low cost and high reliability was a direct result of wide variations in component manufacturers, package profiles, interconnect complexity, and even non-tangibles such as varying RF interference characteristics.

An example of recent VSAT equipment advancement is the STEL-9236 L-Band PSK receiver developed by Stanford Telecom. The product serves as the heart of a VSAT system by receiving an L-Band signal and converting, demodulating and decoding the signal to provide a stream of recognizable data at the output.

Stanford Telecom relied heavily on VLSI technology to reduce parts count and implement most of the receiver digitally, minimizing the number of analog components. As a result of considering manufacturing technology during design, built-in-test ports were included for digital automated testing. By designing the VLSI circuits in conjunction with the boards, pinouts were chosen to reduce the number of trace overlaps, reducing the PCB layers required, and enabling test vias to be placed with little impact on the consumed area.

Isolation of the RF analog components was achieved by using a fully self-contained L-Band to IF down converter. Surface mount technology was maximized throughout the design. Furthermore, the receiver was designed as a single sided assembly. The result was a low cost, high performance receiver that achieves its advantage through reduced parts count and increased manufacturability.

As the shift of priorities moves from solely technical to one split between technical and financial, concurrent development rises to achieve both objectives. Without this new focus on product development, manufacturers would not be able to offer sufficiently reduced pricing to stimulate the market volume requirements.

tronic Circuits (IPC) publishes various standards including *IPC -D-275, Design Standard for Rigid Printed Boards and Rigid Printed Board Assemblies,* which describe accepted procedures for trace routings and protection.

Thermal management is a major concern for the manufacturability of high frequency printed circuit assemblies. One prime consideration is to keep thermal masses balanced when routing traces to lands. This provides relatively equal reflow times for each solder joint. Vias are often tied to internal ground planes which simulate a large heat sink. This is enough to a cause a thermal imbalance during reflow. To avoid this, it's helpful to use thermal relief pads where appropriate. Also, high power components should be strategically distributed to avoid focused heat damage, improving the reliability of sensitive circuits.

### Cleaning

Prior to selecting and laying out the RF components, the designer needs to evaluate the current cleaning process for any possible restrictions or limitations. Some RF components do not lend themselves to proper cleaning due to inaccessible areas and obstruction of cleaning solvent flow. All components need to be strategically laid out in order to allow proper cleaning underneath and around the packages. Moreover, many RF components can't withstand chlorofluorocarbon (CFC) based solvents and conventional cleaning processes. In this case, hand cleaning of the assemblies must be employed.

With environmental hazards and pending government bans of CFCs, manufacturers are seeking to adapt aqueous or semi-aqueous cleaning solutions. Both an aqueous system with a saponifier (a detergent) or a semi-aqueous system typically provide the power of CFC based systems. However with either alternative, they require a post cleaning and drying operation, a significant process addition from CFC cleaning. The key is to match up a flux which will provide comparable solderability attributes and can be cleaned thoroughly. Rosin and water-soluble fluxes for aqueous or semi-aqueous cleaning systems are examples of comparable substitutes.

### Design For Manufacturability

A systematic Design for Manufacturability (DFM) approach provides for an opportunity to achieve cost and development benefits and to reduce the time to production. Often times, there is a conflict between circuit optimization and assembly productivity. For example, the designer, in an attempt to minimize crosstalk, may choose to lay out components in a specific pattern that may compromise the efficiency of the pick and place machines. While these conflicts need to be considered, there remain standard DFM guidelines. These guidelines generally apply to both the RF and digital side of a system, but additional considerations need to be taken specific to high frequency assemblies. Table 5 describes some of the major points.

Power dissipation, especially in high frequency systems, poses thermal management problems. Strategic layout of high power components will minimize failures due to progressive heat damage. High power RF components also produce spurious signals which in turn introduce isolation and production challenges. RF subsystems may require shielding and buffering from other circuits and the environment in order to meet acceptable emission specifications. Each RF sub-system can be designed with surrounding ground planes or casting modules. The use of ground vias to connect the component packages to the PCB insures a good RF ground connection, improved isolation, and reduces the possibility of oscillations in amplifiers. Concurrently, the designer needs to be cognizant of thermal management issues to prevent a heat sink effect, as previously mentioned. Environmental Stress Screening (ESS) provides an opportunity to test systems in variable environments, thus increasing the reliability in the field.

RF/microwave systems commonly demand mixed technology assemblies

- Place packages to minimize crosstalk and impedance problems caused by adjacent traces and components
- Place parts to optimize power and ground distribution, and conduct a thermal energy survey
- Develop an ESS procedure/Qualification tests
- Minimize the number of different parts and improve yield and reliability of the key RF components
- Eliminate multiple solder and cleaning steps

**Table 5. DFM guidelines.**

due to component package limitations and functional requirements. Although single sided SMT assemblies are the most efficient and cost effective, sometimes complex designs necessitate double sided SMT and through-hole part placement. Many high frequency components do not lend themselves to automation. These delicate parts have to be hand soldered to minimize damage and to assure reliability. As a general rule, minimizing soldering and cleaning steps will improve assembly yields and reduce manufacturing costs.

## Conclusion

New markets are requiring RF designs in smaller packages, lighter weights, higher volumes and lower cost, in capacities previously known mainly in the digital world. Many new product innovations are merging RF with digital circuits on the same PCB; combining RF and digital circuitry is a must for the emerging wireless communication markets.

Many companies, especially start-up companies are turning to contract manufacturers to procure, assemble and test their products. Since contract manufacturing's popularity is rooted in digital circuit card assembly, many are unprepared to address the differences in RF assembly. For example, many contract manufacturers may not have RF procurement specialists, comprehend proper shielding or grounding techniques, or own the necessary test equipment required for RF circuits. In selecting a contract manufacturer to build your products, it is essential to address all these issues. *RF*


## Acknowledgement

The authors would like to thank Jack Foley Jr. for his extensive research and contributions.


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### About the Authors

Robert L. Barron received his BS in Industrial Engineering from California Polytechnic State University, San Luis Obispo and an MBA from Santa Clara University. Currently, Mr. Barron is the Vice-President of MQA a division of Stanford Telecom.

William J. Choe received his BS in Industrial Engineering from California Polytechnic State University, San Luis Obispo . Currently, Mr. Choe is a Manufacturing Engineer at MQA. The authors can be reached at MQA, 480 Java Drive, Sunnyvale, CA; tel. (408) 745-2670.

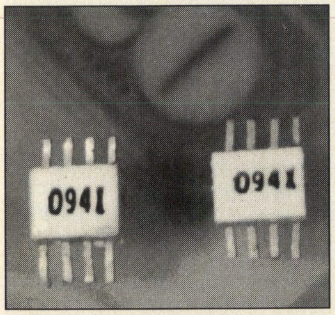
## Arbitrary Waveform Generator


Wavetek has announced a 100 MHz arbitrary waveform generator for bench and ATE use that is less than half the price of other 100 MHz units on the market, while providing the functionality of seven different types of signal generators including synthesized arbitrary waveform, synthesized pulse, synthesized function, noise, sweep, trigger, and amplitude modulation. Priced at $3395, the Model 395 generates simple or complex waveforms up to 10 Vpp and provides a set of synthesized standard waveforms including sine waves to 40 MHz and square waves to 50 MHz. It provides user-defined waveforms from 1 MHz to 100 MHz clock rate with 12 bits of vertical resolution. At clock rates of 50 MHz and below, it has direct digital synthesis capability. Waveforms can be created and modified via the front panel using point-by-point coordinate entry, waveform insert and line editing, or downloaded through the included RS-232 or optional GPIB interface. Up to four waveforms can be linked, and waveforms can be looped. The optional Direct DSO Upload allows uploading waveforms captured with a digital storage oscilloscope directly into the Model 395. Other options include 256k extended memory and rack mount kit.

**Wavetek Corp.**
**INFO/CARD #250**

## SMT, 1W, 1 GHz Amplifier

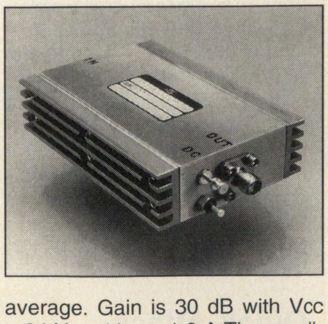
RF Products announces the first in a family of 1000 MHz, 12.5 V, class A MOSFET transistors specifically designed for surface mount, power amplifier applications. The WRLS0941 provides 1 W of class A output power with 12.5 dB of gain from a 12.5 V supply. The device will withstand a 30:1 load mismatch at rated power at 950 MHz. The WRLS0941 is packaged in a special economical, high dissipation, ceramic SO-8 surface mount package. This eight-lead package combines low inductance and high current handling capabilities with cost effectiveness and the advantages of surface mount assembly. The combination of performance, power handling capabilities and economical pricing make the WRLS0941 well suited for cellular and PCS amplifier designs. Pricing at 100 pieces is $19.95, with delivery from stock.




**RF Products, Inc.**
**INFO/CARD #249**

## Multi-Channel, Linear Amplifier

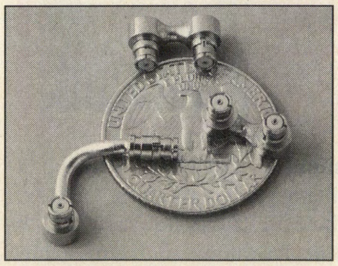
A multi-channel, feed-forward linear amplifier from AML operates in the 1735-1845 MHz PCN environment. The feed-forward amplifier achieves linearity, and therefore very low intermodulation characteristics, by employing a main and an error amplifier. These are arranged in two loops. The first loop seperates the distortion products from the main signal. In the second loop, these products are amplified by the error amplifier and fed forward to the main amplifier 180° out of phase. Using this architecture, the Model PA180003007C provideds intermodulation products of -45 dBc at 7 watt PEP output. This intermodulation performance is maitained over a 6 dB dynamic range.Typical operation will support any number of carriers with up to a total output of 700 mW average. Gain is 30 dB with Vcc = 24 V and Icc = 1.9 A.The amplifier measures 6 × 3 × 1.5 inches, and SMA female connectors are standard.



**AML, Inc.**
**INFO/CARD #248**

## Push-On Cable Assemblies

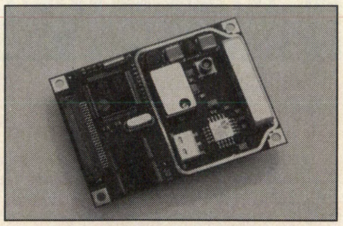
Precision miniature semi-rigid and flexible cable assemblies, feature GPO™ connectors which are 1/5 the size and weight of comparable SMA connectors. The GPO snap-together interconnect system is unique in its ability to mate and operate with up to ± .010" radial misalignment and .010" axial displacement. This push-on connector eliminates threaded couplings and provides excellent vibration performance. Semi-rigid cable assemblies operate through 40 GHz with diameters from .034" to .085". Low density dielectric semi-rigid cable options deliver improved phase stability versus temperature, lower attenuation and a wider operating temperature rage of –65° to +165°C. Flexible GPO assemblies operate through 18 GHz. A quadraform four-ply shield design delivers improved isolation and lower signal attenuation than conventional round braided cables. Operating temperature range is –47° to +125°C. GPO is a registered trademark of Gilbert Engineering.



**Storm Products Co.**
**INFO/CARD #247**

## Small, Low-Power GPS Receiver

Rockwell International has announced a global positioning system (GPS) engine called the NavCore® MicroTracker™. The new, credit-card size, five-channel GPS receiver measures only 2.0 × 2.8 × 0.53 inches and weighs two ounces. A power management option can reduce average power usage to as low as 670 mW. The MicroTracker is designed to operate with an inexpensive passive antenna in most applications. RTCM SC-104 differential GPS compatibility is an option on the new engine. The MicroTracker has the same interface, software and general performance features of the NavCore V GPS receiver, introduced in 1991. Standard features of the MicroTracker include a time-to-first fix of 20 to 30 seconds (from warm start), a normal operating range of +30 to +75 degrees C; and dynamic tracking, both in foliage and urban environments and under conditions where severe vibration and shock are present. MicroTracker will sell for $480 in quantities of 200, with production quantities available in the fourth quarter of 1993.

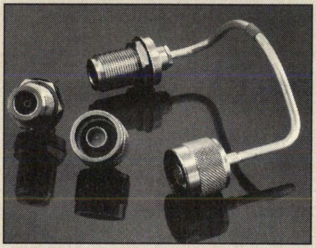


**Rockwell International**
**INFO/CARD #246**

# Product Spotlight: Connectors

## Type N Plugs and Jacks

Ideal for fast production 0.085 inch and 0.141 inch semi-rigid cable assembly requirements, the 4000 series of Type N plugs and jacks feature captivated contacts which plug on to the cable center connector, thereby eliminating contact soldering and gapping. The devices are constructed



of 303 stainless steel, with PTFE insulators and beryllium/copper contacts. VSWR performance for these devices is rated at 1.20:1 max. to 18 GHz using 0.141 inch semi-rigid cable.

**Applied Engineering Products**
**INFO/CARD #245**

## Corrugated Cable Connectors

RFI announces the introduction of its corruflex cable connectors designed to fit all popularly available 1/2 and 7/8 inch foam dielectric corrugated cables. All fittings are machined from solid brass and silver plated, with PTFE dielectrics and gold-plated beryllium contacts. All are water resistant. All connectors can be installed in 3 to 5 minutes, using no special tools. The connectors exhibit over 250 lbs. of pull strength and VSWR less than 1.2 up to 2.5 GHz and beyond.

**RF Industries, Ltd.**
**INFO/CARD #244**

## Quick-Attach Connectors

Andrew Corp. announces the availability of the C41-series quick attach connectors for its line of 1/4 inch super-flexible HELIAX$^R$ coaxial cable. The connectors employ a unique "collet compression" design that makes attachment quick and easy while providing high retention against pull-off. The connectors can be field-installed in less than three minutes. The type N and SMA connectors have been optimized for low VSWR up to 6.5 GHz.

**Andrew Corp.**
**INFO/CARD #243**

## Mixed Layout Connectors

Coaxial-Inserts 1.0 and 2.3 from Huber+Suhner comply with DIN 41626/2 and are designed for the insertion of mixed layout connectors DIN 41612 (pattern M). The coaxial inserts allow the transmission of low frequency signals and high frequency signals in one unit. Fast and easy assembling, as well as high reproducibility due to the Suhner full-crimp cable attachment, are typical features of this series.

**Huber+Suhner AG**
**INFO/CARD #242**

## SC Coaxial/EIA Adapters

Custom manufactured SC coaxial to 1 5/8 inch EIA adapters are being produced by Tru-Connector. The adapters feature pressure sealed construction and are manufactured in straight-through and right-angle designs that meet MIL-C-39012 specs.

**Tru-Connector Corp.**
**INFO/CARD #241**

## BNC Filtered Connector

Metuchen Capacitors now distributes Oxley's BNC filter connector line. The Oxley BNC connector with integral noise filtering is useful where a coaxial cable screen is not to be grounded directly to the equipment chassis. The connectors provide AC coupling between a coaxial cable's screen and the equipment chassis via an integral multilayer ceramic capacitor.

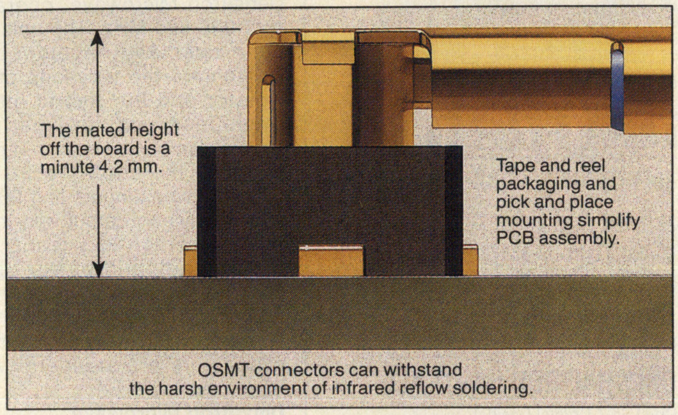
**Metuchen Capacitors Inc.**
**INFO/CARD #240**

# Until now, no one has produced surface mount connectors with greater frequency.

Now all of the advantages of surface mount technology are available for applications that require microwave coaxial connectors with frequencies of up to 6 GHz.

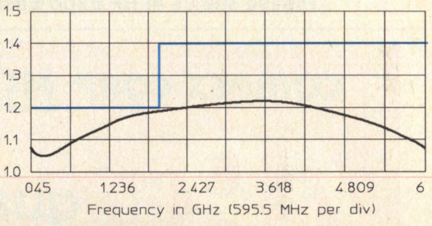
Surface mount technology offers a number of advantages over traditional through-hole connectors. It reduces the required amount of PCB real estate, allowing for greater component density. The smaller, lighter connectors resist shock and vibration better. And surface mount technology gives OEMs the benefit of improving quality and producing a smaller end product.

But until now, true surface mount connector technology did not cover the broad spectrum of microwave frequencies. Now we have an interconnect solution for every application, in every market.




Their mated height off the board is a minute 4.2 millimeters. And they use substantially less PCB real estate than standard through-hole connectors, which allows for denser packaging and results in smaller, lighter PCBs.

Despite all of their technological advantages, OSMT connectors don't require costly new equipment or placement procedures. In fact, they can be installed using standard surface mount processes. The connectors are packaged to be ESD safe, are available in tape and reel packaging, and are pick and place mountable. If you're equipped for SMT, OSMT won't increase your expenses at all. In fact, the smaller size, higher quality and improved performance of these connectors will ultimately lower your installed costs.

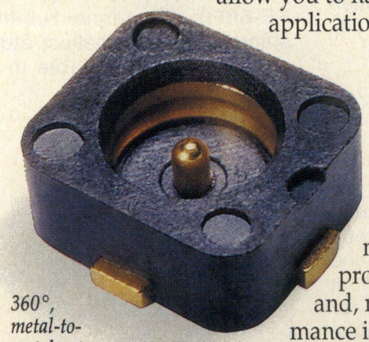


TYPICAL VSWR-OSMT MATED PAIR

## The interconnect system that eliminates obsolescence.

M/A-COM now offers true surface mount coaxial connectors that are rated from DC to 6 GHz. They'll allow you to handle the widest range of applications—now and in the future.

With M/A-COM Surface Mount Technology's interconnect system (OSMT), you can count on the smaller PCBs, lower profiles, higher reliability, and increased quality yields of SMT for all of your microwave applications. The products are readily available and, most important, their performance is always superior to that of traditional through-hole connectors.



*360°, metal-to-metal outer contacts give you superior RF transmission.*

## OSMT microwave coaxial connectors have a number of advantages.

OSMT coaxial connectors are the first and only ones to be rated all the way from DC to 6 GHz with a VSWR rating of 1.2 @ 2 GHz Max; 1.4 @ 6 GHz Max.

Our connectors are rated to 100 mating cycles. And their full circle metal-to-metal outer contacts give you complete, reliable RF transmission. They're also durable enough to withstand the high temperatures and harsh environment of infrared reflow soldering.

## Being first is nothing new.

M/A-COM developed and produced the first SMA and blind mate connectors for the volume market. And now M/A-COM is in the vanguard again, with the first microwave coaxial connectors that are truly surface mounted.

At M/A-COM, we've made research and development efforts a top priority, so we can continue leading the way in developing the products you need. And you can count on our manufacturing capabilities to produce the quantity and the quality you require, no matter how innovative the product or demanding the application.

Whenever your application requires RF or microwave components, make sure they come from the world leader. Make sure they come from M/A-COM.

For more information on the OSMT surface mount interconnect system, write to M/A-COM, 140 Fourth Avenue, Waltham, MA 02254. Or call 617-890-4750. In Europe: +44 (0344) 869 595. In Asia: +81 (03) 3226 1671.

M/A-COM

INFO/CARD 62

Please see us at RF Expo East '93, Booth #604, 606

## SIGNAL PROCESSING COMPONENTS

### Isolator/Hybrid

The GSM ISOHYBRID, from Densitron Microwave, operates over 925-960 MHz. The unit has been specially developed to incorporate the functions of an isolator and 90° hybrid into a single, compact, low-cost unit for cellular base station applications. Isolation is 50 dB (min.) between inputs and 26 dB (min.) between output and input ports; insertion loss is 0.6 dB, excluding hybrid power split. Forward and reverse power handling is 60 W. The device measures 75 × 75 × 26mm. N type connectors are standard, with others available.
**Densitron Microwave**
**INFO/CARD #239**

### 10 W Digitally Tuned Filter

Pole/Zero announces the introduction of the POWER/POLE™ digitally tuned filter with power handling capability of 10 W. The filters cover 10 to 400 MHz in four separate bands: 10-30 MHz, 30-90 MHz, 90-200 MHz and 225-400 MHz. They use PIN diode technology, yielding fast 10 µs tune times and IP3 performance on the order of +50 dBm. The filters measure 3 × 3 × 2 inches and quantities of 1-4 are between $1500 and $2500 with deep discounts for higher quantities.
**Pole/Zero Corp.**
**INFO/CARD #238**

### Video Filters

KR Electronics' video filters are available in three performance levels. The filters feature group delay equalization, flat passband response and small PCB mount packaging. Sin(x)/x shaping for post D/A conversion is available. The filters come in both luminance and chrominance bandwidths. A brochure containing specifications, plots and outline drawings is available.
**KR Electronics, Inc.**
**INFO/CARD #237**

### Multi-Position Failsafe Switch

Model STR-10, from RLS Electronics is a 7-10 position, terminated, multi-position, failsafe, coaxial switch. These switches have extremely low insertion loss and VSWR and high isolation over the DC-18 GHz range. These failsafe switches are available in remote DC operation with a 25 ms maximum switching time. Prices start at $1325 for the 7-position switch in unit quantities.
**RLC Electronics, Inc.**
**INFO/CARD #236**

## TEST EQUIPMENT

### Network Analyzer Calibrators

Maury Microwave announces the release of the 8770C series of precision K (2.92mm) calibration kits. All 8770C series calibration kits contain a full complement of both female and male calibration standards including fixed shorts, open circuits, and both fixed and sliding terminations for high accuracy calibrations from 40 MHz to 40 GHz. kits are available for use with a variety of network analyzers including the HP8510A/B/C, HP8719A/C, HP8720A/B/C, HP8722A, and the Wiltron 360.
**Maury Microwave Corp.**
**INFO/CARD #235**

### Miniature Rubidium Standard

Frequency Electronics' model FE-5650 commercial rubidium standard is the smallest atomic standard now available in the

marketplace. The FE-5650 is contained in an incredibly small 3 × 3 × 1.4 inch package and features power consumption < 7.5 W, warm-up time < 4 minutes, low phase noise, outstanding accuracy, low spurious and excellent harmonics. It can be

factory set at any frequency from 10 kHz to 20 MHz, with a setting resolution of $2 \times 10^{-12}$.
**Frequency Electronics, Inc.**
**INFO/CARD #234**

## VXIbus Synthesizer

EIP Microwave has introduced the model 1140A, a synthesized signal generator combining 0.01 to 20 GHz performance with 1 Hz resolution, and outstanding spectral purity. In addition to amplitude and phase modulation capability, the 1140A offers unique IF modulation, providing the means to mix complex digital signals or complex jamming scenarios onto the microwave output signal. Spurious signals are below -60 dBc, and phase noise is -80 dBc/Hz at 10 kHz offset. The EIP 1140A is a 3-wide, C-size VXIbus module.
**EIP Microwave, Inc.**
**INFO/CARD #233**

# SIGNAL SOURCES

## 10 MHz Oscillator

Wenzel Associates introduces a 10 MHz At-cut crystal oscillator featuring excellent phase noise, –170 dBc/Hz at 10 kHz offset, and temperature stability of $\pm 2\times10^{-8}$ over 0° to +50° C. This dual output, shielded oscillator is specially mounted for vibration isolation. A card edge connector is featured for ease and high reliability without solder. A solder pin version is also available. Electrical and/or mechanical tuning is provided.
**Wenzel Associates, Inc.**
**INFO/CARD #232**

## ECL X-tal Oscillators

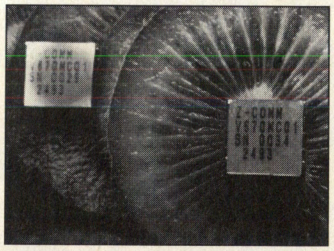
Connor-Winfield introduces a series of ECL crystal oscillators which are drop-in replacements for SAW oscillators. These units are pin compatible and feature a quartz based circuit for applications requiring frequency stabilities as tight as 25 ppm. The E72 series is available to operate over temperature ranges as wide as –40° to +85° C. Currently, frequencies are available to 500 MHz.
**Connor-Winfield Corp.**
**INFO/CARD #231**

## VCXOs

Designed specifically for phase lock loop (PLL) circuits, the 322 series is a TTL/CMOS output VCXO. The series is offered over a frequency range of 1.5 to 60 MHz and can be ordered at common PLL frequencies. Stability can be specified as tight as ±25 ppm over –40° to +85° C, with deviation options as wide as ±200 ppm. The 322 comes in an all metal hermetically sealed 14-pin DIP package. Price for a typical 44 MHz unit is $27.000 at 500 pieces.
**Reeves-Hoffman**
**INFO/CARD #230**

## Surface Mount VCO

Z-Communications announces the V670MC01 VCO with a frequency range of 1.68 to 2.58 GHz. Intended for satellite receiver applications, the VCO generates a 6.5 ±2 dBm signal into a 50 ohm load and pulls less than 9 MHz with a 14 dB return loss and will push no more than 3 MHz.



Phase noise at 10 kHz from the carrier is specified at –90 dBc/Hz (typ.). The V67MC01 draws less than 40 mA at 15 V. The V670MC01 is available in the MINI surface mount package, measuring only 0.5 × 0.5 × 0.2 inches.
**Z-Communications, Inc.**
**INFO/CARD #229**

# SEMI-CONDUCTORS

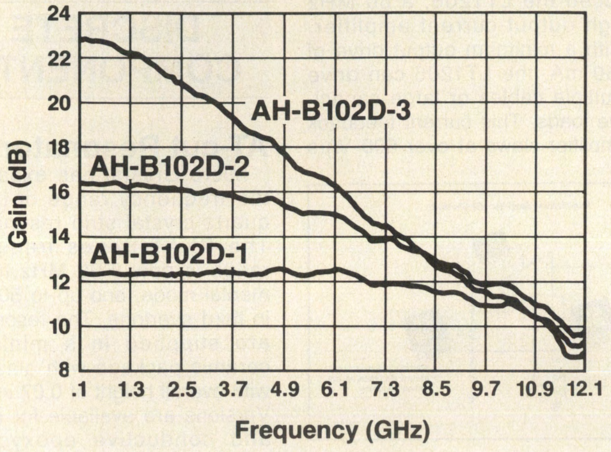
## Dual, Power Op-Amp

Burr-Brown's OPA2662 combines two operational transconductance amplifiers (OTAs), each with ±75 mA drive capability, or doubled drive capability when the two are paralleled. Current slew rate is 58 mA/ns, and TTL-compatible switching stages provide 30 ns/200 ns enable/disable times. The device has stable operation with capacitive and inductive loads and 370 MHz

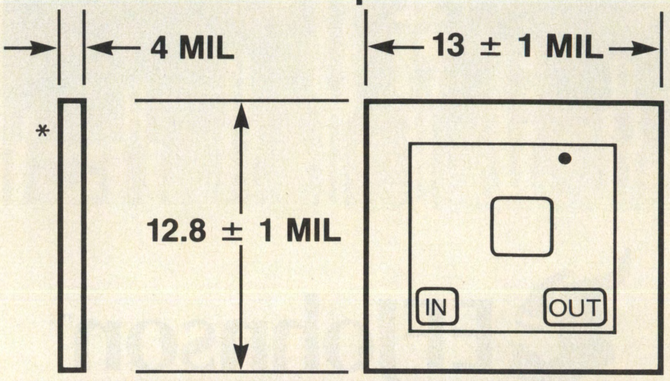
# Introducing . . .

# A New Low Cost MMIC Darlington Gain Block Amplifier

### AH-B102D-1,-2.-3 Gain vs Frequency




### AH-B102D-1 Chip




*Back of chip is ground

Note: Single dot indicates -1 part

Although the circuit diagram is the classical Darlington amplifier, it has been implemented as a MMIC (Microwave Monolithic Integrated Circuit) using HBT (Heterojunction Bipolar Transistor) technology on GaAs (Gallium Arsenide). This change from conventional Silicon Bipolar technology results in a greatly increased Gain-Bandwidth product. Since GaAs material is insulating, all of the circuit connections are on the surface. This amplifier uses a via to connect the circuit ground to the back of the chip (rather than the back contact being the common collector point).

All that is required for use is to die attach to ground and connect the input and output. The positive bias is connected to the amplifier through the output terminal as in a conventional Darlington amplifier.

The gain is determined by feedback resistors that are part of the chip; therefore, there are different dash numbers for the various gain versions. These devices are also available in standard 70 mil, 4-lead packages.

**Typical Performance Characteristics (Ta = 25°C)**

| AH-B102D-1 (-70C)* | |
|---|---|
| Gain (dB) | 12 |
| Bandwidth (± 0.5 dB, GHz) | 7 |
| Bandwidth (3 dB, GHz) | 10 |
| P1 dB (dBm) | 13 |
| Reverse Insertion (dB) | 17 |
| VSWR, Input ( ≤ 7 GHz) | 2.0:1 |
| VSWR, Output ( ≤ 7 GHz) | 2.0:1 |
| Noise Figure (dB) ( ≤ 7 GHz) | 6.5 |

* Commercial ceramic package data

Prices in quantities of 1000 are: $10, each for chips and $14.50, each in package form. Delivery for either version is stock to 30 days. For Data Sheets contact:

FEI Microwave, Inc., 825 Stewart Drive, Sunnyvale, California, 94086. Telephone: (408) 732-0880. FAX (408) 730-1622.

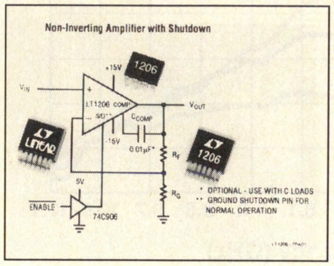
**FEI Microwave, Inc.**
A SUBSIDIARY OF FREQUENCY ELECTRONICS, INC.

INFO/CARD 68

bandwidth. Priced from $11.00 in 100s, delivery is from stock.
**Burr-Brown Corp.**
**INFO/CARD #228**

## Current Feedback Amplifier

Linear Technology has introduced the LT1206, a 60 MHz high output-current amplifier. With a minimum output drive of 250 mA, the LT1206 can drive multiple cables or large capacitive loads. This current feedback amplifier slews at over 600 V/µs




and operates on all supplies from ±5 V to ±15 V. Operating from a ±5 V supply, the LT1206 has flat response to 18 MHz, differential gain error of 0.07%, and differential phase error of 0.12°. Pricing in quantities >100 is $3.45 in 8-pin DIP, $3.95 in 8-lead SOIC, and $4.45 in 7-lead TO-220 or SMT DD packages.
**Linear Technology Corp.**
**INFO/CARD #227**

## High-Speed Standard Cells

Raytheon Semiconductor has added over 35 cells to its RSC4000 CBiCMOS mixed signal cell library. Additions to the library of low-cost, 2 micron technology cells include the A0600B, 6-bit, flash A/D converter, and the D0800B, 8-bit video DAC. The A0600B features a 50 MSPS minimum conversion time, while the D0800B features 50 MSPS minimum conversion time, single supply operation and a 37.5 or 75 ohm load.
**Raytheon Semiconductor**
**INFO/CARD #226**

## Sampling Amplifier

Analog Devices' AD9101 Samplifier™ comprises a track-and-hold amplifier (THA) followed by a gain-of-four amplifier. The AD9101's THA architecture enables acquisition time to 8 bits (0.4%) of 5 ns; to 10 bits (0.1%) is 7 ns; and to 12 bits (0.01%) is only 11 ns. The AD9101 sampling amplifier is available in a 20-pin SOIC or 20-contact ceramic LCC package. Price is $33.00 in 100s.
**Analog Devices**
**INFO/CARD #225**

# DISCRETE COMPONENTS

## AT-cut Resonators

Micro Crystal has expanded the frequency range of its AT quartz crystal strip resonators. The CXAT Series frequency range is now 8-32 MHz fundamental mode, and up to 50 MHz in third overtone. The resonators are supplied in a miniature ceramic package with glass lid, with overall height of 0.07 inches. Versions are available for reflow and conductive epoxy SMT attachment and for through-hole mounting.
**Micro Crystal/Div. of SMH**
**INFO/CARD #224**

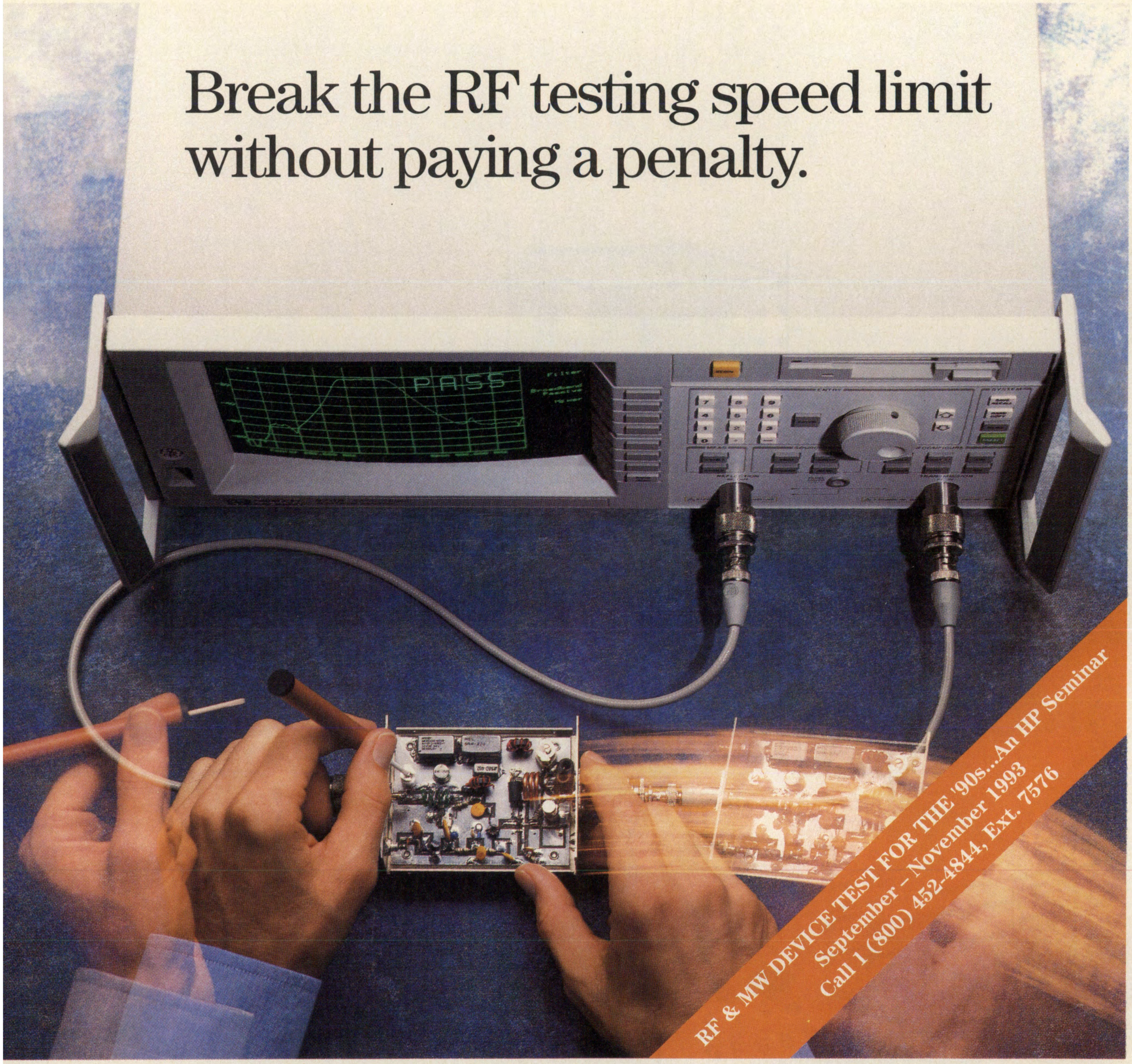
## Cellular/Paging Crystals

TeleQuarz USA is pleased to announce the development of the SMD-008 series of AT-cut crystals. The series designed for use in the LO IF section of cell/mobile phones and paging systems. SMD-008 crystals are available over 5 to 300 MHz, with temperature stabilities down to ±2.5ppm over −10 to +50° C. Resistance-weld housings as small as 6.5 × 5.1 × 1.5mm are available. Tape and reel packaging is standard.
**TeleQuarz USA**
**INFO/CARD #223**

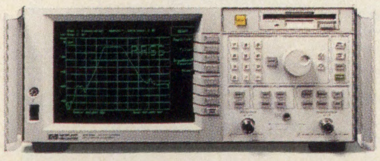
## Bypass Capacitors

AVX MAXI-SLC microwave chip capacitors show virtually no resonance from 40 MHz to 18 GHz. The capacitors exhibit high temperature stability, (±15% capacitance change over -55 to +125° C for a 400 pF, 0.025 × 0.025 inch device). The MAXI-SLCs are terminated with sputtered gold over titanium/tungsten. Gold/nickel is also available. Capacitance ranges from 68 pF to 6300 pF.
**AVX Corp.**
**INFO/CARD #222**

# Break the RF testing speed limit without paying a penalty.





**The new HP 8711A makes faster RF testing affordable.**

To beat the competition in RF manufacturing, you have to get products tested and out the door faster, while keeping costs down. And with the new HP 8711A network analyzer you can do just that.

The HP 8711A brings fast trace update and synthesized accuracy together for the first time. So you can tune in "real-time" from 300 kHz to 1300 MHz—without frequency drift.

Selectable broadband/narrow-band measurement modes let you test conversion loss of frequency translators and mixers. And make high dynamic range (90dB) measurements on filters and switches. All with the same instrument. You don't even need a computer. Built-in automation lets you race through tests without one.

As for cost control, the HP 8711A is just $13,500.* At that price, the only penalty is not having one.

So, move fast. If you'd like more information on the HP 8711A, call **1-800-452-4844.** Ask for **Ext. 2518**, and we'll send you a free video and brochure that show how you can afford to go a lot faster.

There is a better way.

HEWLETT PACKARD

*U.S. Price only.

INFO/CARD 71

**Please see us at RF Expo East '93, Booth #404, 406, 503, 505**

# Notes on Power Supply Decoupling

By Gary A. Breed
Editor

*To avoid unwanted oscillations, interactions and radiation, RF circuits must by properly isolated from power supply circuitry. This short tutorial reviews the reasons why various forms of decoupling are used, and the methods typically employed for their implementation.*

Before specifically looking at RF, let's go over the reasons for decoupling at DC. Primarily, the goals are to prevent centrally-located disturbances from affecting individual circuit sections, and preventing localized disturbances from traveling through the power bus to other circuit sections. Typically, these tasks are accomplished three different ways (Figure 1).

First and most common, is energy storage using a large capacitor. Fluctuations in the DC voltage are absorbed by the "sink" effect of the capacitance. Excess charge due to increased voltage is stored, or charge can be released to make up for a drop in supply voltage. The second method is an extension of this one; adding a resistor to increase the time constant of the decoupling network. This configuration introduces a voltage drop across the resistor, but it effectively smooths out variations from the power supply.

The most effective DC decoupling is distributed voltage regulation. A voltage regulator is placed at each individual circuit section, rather than having one large capacity regulator at the central supply. These small sections can be easily analyzed for localized interactions, while remaining well isolated from the main power bus.

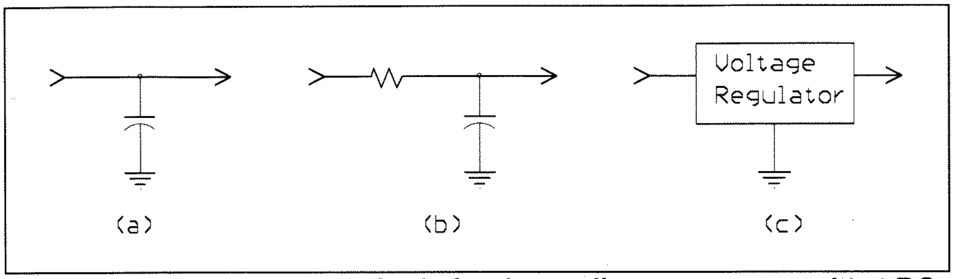



**Figure 1. Three common methods for decoupling a power supply at DC.**

## RF Decoupling

At radio frequencies, the task of decoupling has the same task of preventing unwanted disturbances to and from the power bus. The DC power supply still needs to be isolated from the local circuit, as described above, but the decoupling circuit needs to be effective at high frequencies, not just DC. High frequency noise that may be present on the power bus must be kept out of the RF circuit, and RF energy needs to be kept from the power interconnections. Finally, the power supply decoupling must represent a low impedance at RF in order to avoid becoming an unwanted part of the circuit it is supposed to help. This last requirement is often overlooked until the circuit fails to operate properly.

Several "rules of thumb" are commonly used for RF bypassing/decoupling, providing adequate performance in most cases. However, using simple rules without understanding why they were developed can get you into trouble when you use them improperly.

*Single bypass capacitor* — The most commonly used RF decoupling method is a single capacitor located at the point where the power supply is connected to the active device. This is usually in a transistor drain or collector circuit, on the power supply side of any tuned circuits or impedance matching components. See Figure 2(a). The purpose of this capacitor is simply to look like a short circuit at RF, while allowing DC to pass unchanged. The rule of thumb is to use a capacitor with only a couple ohms reactance at the operating frequency, while avoiding resonance due to the inductance of the component leads and circuit wiring. As an example, a .01 µF capacitor has an impedance of 16 ohms at 1 MHz and 0.53 ohms at 30 MHz; and a 1000 pF capacitor is 5.3 ohms at 30 MHz and 1.6 ohms at 100 MHz.

*Multiple bypass capacitors* — In broadband circuits, and where extra DC or low frequency decoupling is desired, it is common to use two, three or more capacitors for bypassing. A typical example might be a 1 µF tantalum electrolytic for low frequencies, 0.1 µF for medium frequencies, and 0.01 µF for high frequencies, as in Figure 2(b). Although this is a common practice, some engineers feel that the combination can actually result in unwanted resonances and reduced effectiveness.

*Lowpass networks* — RC and LC combinations are often employed to achieve the additional isolation afforded by a multi-pole network. In this case, the idea is not just to create a short circuit, but to build a filter which prevents the escape (or entrance) of RF energy, as shown in Figure 2(c). Engineers are cautioned to observe one rule when using such a configuration: start with a single bypass capacitor that is of the proper value, then use the additional components to extend its effective decoupling to a lower frequency. This method is primarily used to reject noise at frequencies lower than the circuit's normal operation, not to increase bypassing effectiveness at the operating frequency.

*Ferrite beads* — These products represent lossy inductors at RF, and can be used to present a high impedance in series with the power supply lead, rather than a short circuit across it, like a bypass capacitor. Used alone, they present a high impedance to the RF side of the circuit, which may be very undesirable. Most often, ferrite beads are used as the inductive element in LC decoupling networks, as shown in Figure 2(d).

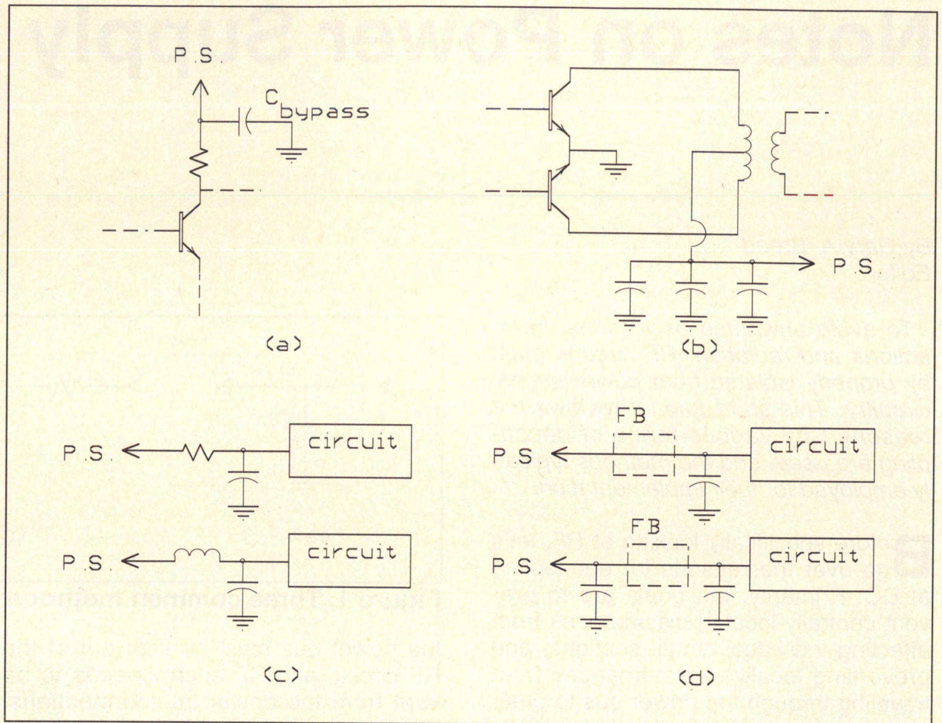



**Figure 2. (a) Single-capacitor bypassing; (b) Multiple-capacitor bypassing; (c) RC and LC networks for extra low-frequency rejection, and; (d) LC (top) and PI network (bottom) decoupling using ferrite beads.**

## Impedance Effects

All of the above methods include cautions against presenting a high impedance to the RF circuit, or avoiding resonances that can upset circuit operation. Decoupling is certainly prone to the old saying, "At RF, components are R, L and C at the same time." The resistance and reactance of the decoupling circuit must be accounted for.

A one-inch length of #24 AWG wire has an inductance of about 20 nH. For a typical through-hole printed circuit board, capacitor leads can be as short as 1/10 inch, or 2 nH. If the capacitor is .01 µF, the series resonance will be around 35 MHz. That value of capacitor will be increasingly ineffective at frequencies above that, since its impedance rises beyond the point of series resonance.

The lowest possible inductance is obtained with surface mounted devices. A chip capacitor has no lead inductance, only the inductance due to the capacitor electrodes and internal construction. While a 0.1 µF capacitor with wire leads might be useful to 11 MHz, a 0.1 µF chip capacitor may have one-tenth the inductance, making it useful to 35 MHz. In the situation where multiple bypass capacitors are being considered, a single larger-value chip capacitor may be a better choice than two leaded devices.

When decoupling with a two-element RC or LC network, the capacitor must be on the RF circuit side, to place a low impedance capacitor immediately adjacent to the circuit. The high-imedance series resistor or inductor should be on the power supply side.

## Conclusion

Decoupling RF circuits from their power supplies is essential for proper operation. In nearly every case, the goal is to have the power connection look like a short circuit at RF. An acceptable target impedance might be 1/20 of the circuit impedance or 10 ohms, whichever is lower.

Additional considerations might include filtering low frequency noise and suppressing higher frequency harmonics, but these tasks must be accomplished without compromisinglow impedance at the operating frequency range.

Simple bypass capacitors will usually provide sufficient decoupling performance, but occasionally, more complex networks will be needed to reject a wider frequency range. The primary caution is to avoid high impedances caused by excessive lead length or by incorrect use of RC or LC decoupling circuits. *RF*

# Combless Generator Tests Radar Warning Receivers

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*The conventional method for testing wideband or multiband microwave receivers is to utilize a combination of stable VHF or UHF signal source and a frequency multiplier or "comb generator" to produce harmonics in the bands of interest. The technique presented here eliminates costly harmonic multiplier components, employing nonlinearities in the receiver under test itself to generate the required test frequencies. Here is the author's account of this technique's development.*

As a consulting engineer, I am frequently employed by clients who have a clear idea of how they want to implement a given RF or microwave function, but require outside circuit design expertise. Generally, I try to give my clients exactly what they pay for. Every now and then, however, an unconventional solution presents itself which is so exciting that an enlightened client will dispense with his or her preconceived notions and try something new.

My client [1] had already secured a patent on RadaRanger™, a product for testing multiband police radar detectors. I was retained to finalize, perfect and package the required microwave circuitry. Three months into the project I had one of those "Aha!" insights for which all engineers pray: that the job can be done better, cheaper and more elegantly in the RF spectrum. My client was progressive enough to embrace the breakthrough. The results have been a new patent application, a new product line, and a new approach which other RF designers may find appealing.

## Prior Art

There's nothing new about testing microwave receivers with lower-frequency oscillators and harmonic generators; I remember first seeing the technique in the MIT RadLab Series, circa 1945. All that's needed is a stable RF source and a non-linear circuit to generate harmonics, as depicted in Figure 1. If multiple microwave output frequencies are required, then an unfiltered comb of frequencies may be employed. The only constraint is that all the bands tested must share a common integer subharmonic.

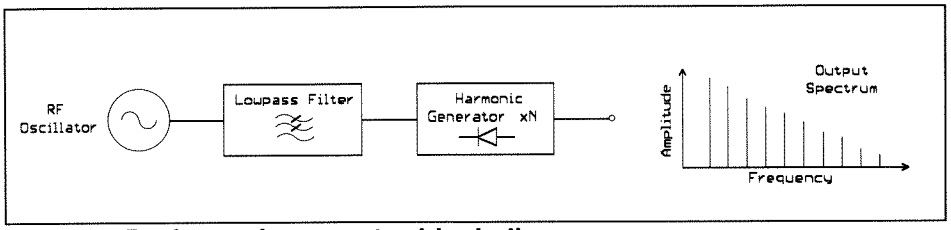


**Figure 1. Basic comb generator block diagram.**

Radio amateurs, whose original bands were all harmonically related, once used a 3.5 MHz crystal oscillator along with a diode comb generator to produce test signals in the 3.5, 7, 14, 21, 28, 56 and 112 MHz bands [2]. More recently, microwave hams have found that one "magic" frequency, 1152 MHz, is a subharmonic of calling frequencies at 2304, 3456, 5760, 10368 and 24192 MHz. An oscillator at 1152 MHz, followed by a broadband comb generator is often used as a "weak signal source" for testing microwave receivers in all five bands.

Now, how to generate the required harmonics? Step recovery diodes have been the traditional favorite [3,4], but at a recent Microwave Update conference two papers were presented which utilized the nonlinearities on MMICs [7,8]. Ward [7] started with a 96 MHz crystal controlled oscillator, then employed a rather expensive silicon bipolar MMIC to generate useful harmonics past 10 GHz. Wade [8] instead started with an 80 MHz TTL oscillator. Its harmonic-rich square wave output drives a much less costly MMIC to useful output in the 5 GHz region.

The original RadaRanger circuit, as envisioned by designers Robert Brocia and Marie Dagata, started with a sinusoidal oscillator at 1507 MHz, driving a similar MMIC comb generator circuit. The idea was for the oscillator's seventh, sixteenth and twenty-third harmonics to fall nicely within the X, K and wideband Ka-band police radar frequency allocations. The numbers all worked out fine. But, there were problems in generating adequate signal power at such high integer multiples, which is where I was called into the project.

## A Conventional Solution

Bipolar MMICs are linear circuits, designed to produce sinusoidal, undistorted outputs in normal use. What is required for harmonic generation, however is a high degree of non-linearity. Fortunately, an overdriven bipolar junction transistor can be readily forced into saturation and cutoff, producing a first order approximation of a square wave whose Fourier series is rich in odd harmonics. This requires an input signal power on par with the amplifier's saturated output power. The output amplitude at a given odd harmonic is approximated by:

$$P_n \approx P_{sat} \div n \qquad (1)$$

provided the desired frequency component is an odd harmonic of the input frequency, and is at or below the transistor's transition frequency, that is:

$$f_n \leq f_T$$

But what if, as is the case in the present application, even harmonics are required as well as odd? Since MMICs are typically biased for midpoint conduction, they tend to be driven symmetrically into saturation and cutoff on alternate half-cycles. Such symmetrical clipping produces a square wave, rich only in odd harmonics. The key to even harmonic generation is to clip the sinusoidal waveform asymmetrically. This is

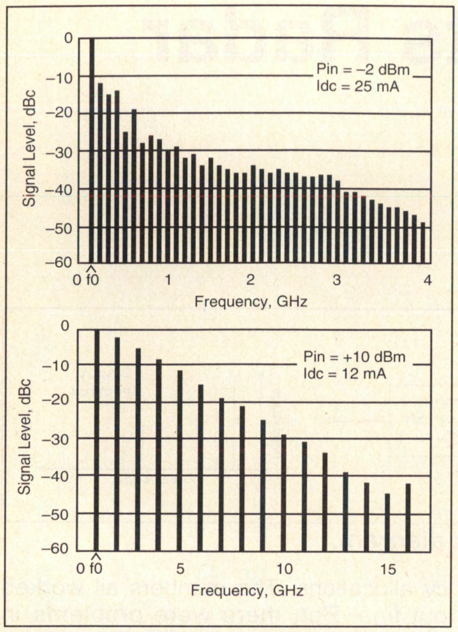


**Figure 2. Harmonic generation vs. frequency for the Avantek INA-03170: (a) $f_0$ = 100 MHz. (b) $f_0$ = 1 GHz.**

accomplished by moving the quiescent DC bias point away from the middle of the load line. To maximize comb generation while enhancing efficiency, simply drop the quiescent collector current of a single-stage bipolar MMIC in half, by increasing the external collector resistor value.

Figure 2, taken from [5], depicts the output spectrum of a three-stage silicon bipolar MMIC biased for harmonic generation. Notice that higher frequency spectral components are enhanced by driving the MMIC with a relatively high input frequency. The only problem with such an approach in the present application is that we require output components at 24 and 34 GHz, and silicon bipolar devices seem to run out of GaAs (pun intended) at around 18 GHz.

To produce comb elements in K and Ka bands, I suppose we're going to have to utilize GaAs FET technology. Figure 3, from [6], depicts the disappointing result. While a bipolar device brought us out to the eleventh harmonic before output amplitude dropped to –30 dBc, the GaAs MMIC shows a –30 dBc

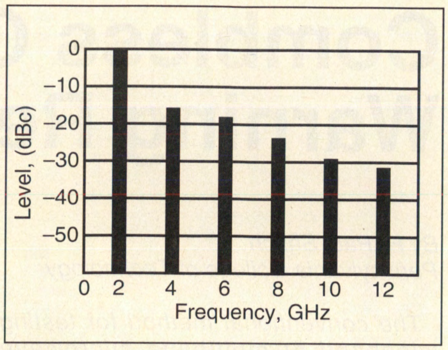


**Figure 3. Typical GaAs MMIC comb generator output spectrum (Avantek MGA64135, $f_0$ = 2 GHz).**

level at only the sixth harmonic. GaAs FETs are by nature highly linear devices. Although their operating frequency exceeds that of their bipolar counterparts, the linearity "advantage" makes it more difficult for them to generate substantial amounts of power at the higher harmonics. Back to the drawing board!

Step recovery diodes have long been utilized to produce harmonic-rich out-

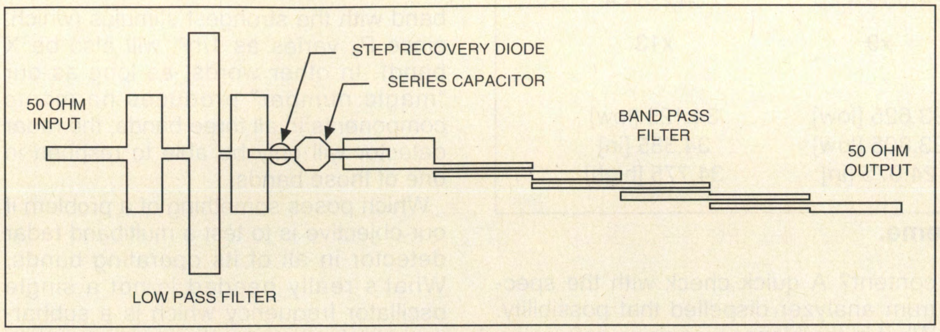


**Figure 4. Typical SRD frequency multiplier.**

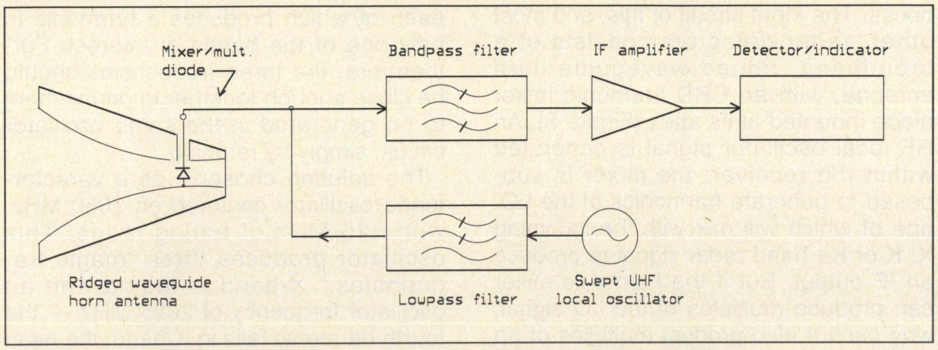


**Figure 5. Typical multiband police doppler radar detector.**

puts well into the microwave spectrum, their only drawback (relative to MMICs) being their lack of gain. However, they seemed a good compromise, in view of the frequency limitations of bipolar devices and the excessive linearity of GaAs FETs. A rule of thumb often cited for the harmonic output power of SRD comb generators is:

$$P_n = P_{in}(1/n) \quad (2)$$

Equation 2 seems to hold relatively well for tuned harmonic generators, such as that depicted in Figure 4 (from [4]). The output bandpass filter enhances specific spectral component, and 1/n power performance is readily achieved. However, it doesn't take long to see in the laboratory that an untuned output comb generator doesn't even come close to satisfying Equation 2. If we derive an equation for the output spectrum of a 1/n comb generator, the reason becomes painfully apparent:

$$\sum_{n=1}^{N}(P_n) = f_2 @ (P_{in} \div 2) + f_3 @ (P_{in} \div 3) + f_4 @ (P_{in} \div 4) + \cdots + f_N @ (P_{in} \div N) \quad (3)$$

If Equation 3 were true, then a SRD comb generator producing the second through fourth harmonics of its input signal would have a total output power of (1/2)+(1/3)+(1/4) = 1.08 times $P_{in}$, obviously a violation of the principle of conservation of energy!

A more realistic estimate of output power from an untuned SRD multiplier might be:

$$P_n = P_{in}(1/n^2) \quad (4)$$

which gives us a total power relationship of:

$$P_t = \sum_{n=2,3,4...}(P_{in} \div n^2) \quad (5)$$

which asymptotically approaches unity. But note that the higher order harmonics drop off quite rapidly in amplitude. Thus, for example, in order to utilize the SRD combline generator's 23rd harmonic for testing Ka band radar detectors we need to drive it at an amplitude of ($23^2$) = 529 times the required output amplitude. Conversely, if we have a given input power available to the SRD, its 23rd harmonic will be, at best, 529 times weaker. For the present project, we anticipated driving the SRD multiplier at about 1 mW input. Thus, its anticipated Ka band output power was on the order of 2 uW, just barely within the sensitivity specs of the receivers being tested.

But another problem arose — an attempt to breadboard a comb generator with a surface-mount packaged SRD was a dismal failure. Output was ample at X band, sub-marginal at Ka band and

| Radar Band | X | K | Ka |
|---|---|---|---|
| Allocation (GHz) | 10.475-10.575 | 24.05-24.25 | 34.2-34.4 |
| n (harmonic) | x4 | x9 | x13 |
| Oscillator Freq. | | | |
| 2625 MHz | 10.5 [in] | 23.625 [low] | 34.125 [low] |
| 2645 MHz | 10.58 [high] | 23.805 [low] | 34.385 [in] |
| 2675 MHz | 10.7 [high] | 24.075 [in] | 34.775 [high] |

**Table 1. RadaRanger™ frequency scheme.**

nonexistent at Ka band! It turns out that the package transconductance of the SOT device is on the order of 2 uH [9], making this package entirely unmatchable at the higher microwave frequencies. The recommended solution: a bare chip SRD, thin-film fabrication, and attendant prohibitive fabrication costs, which nearly scuttled the RadaRanger project.

## The Breakthrough

And then (just as depicted in the famous Sidney Harris cartoon of mathematicians at work), a miracle occurs. While playing around with the oscillator portion of the RadaRanger breadboard (comb generator now sadly abandoned), an X band radar detector which just happened to be turned on, and just happened to be on the bench, beckoned loudly. Curious, thought I, are there police cruising the neighborhood? Are they hot on the trail of some industrial spy, out to steal all my secret circuits? As I turned off the oscillator, the detector silenced. A few flicks of the power supply switch convinced me that the radar detector was somehow responding to my RF oscillator. Was its output frequency coincidentally on the superhetrodyne receiver's intermediate frequency? The spec sheet said it wasn't. Was my oscillator somehow rich in harmonic content? A quick check with the spectrum analyzer dispelled that possibility. What in the blazes was going on?

It hit me like the proverbial ton of bricks. The input circuit of this, and most other, radar detectors consists of a broadband, ridged waveguide horn antenna, with an SRD harmonic mixer diode mounted at its apex (Figure 5). An RF local oscillator signal is generated within the receiver; the mixer is supposed to generate harmonics of the LO, one of which will mix with the incoming X, K or Ka band radar signal to produce an IF output. But if the harmonic mixer can produce multiples of the LO signal, why can't it also product multiples of an incoming 1.5 GHz test signal? It can, and it did. Here was the serendipitous solution to our design dilemma.

## The Next Step

If a 1.5 GHz oscillator can generate harmonics in the input diode of a police radar detector, and if one of those harmonics can trigger the X-band input of a multiband radar detector, can higher harmonics trigger the same detector's K and Ka band modes? Theory said yes, but practice indicated otherwise. When simultaneously excited at multiple bands, most radar detectors either default to indicating a single band threat (typically X band), or respond to the band with the strongest stimulus (which, since $P_n$ varies as $1/n^2$, will also be X band). In other words, as long as our "magic number" produces harmonic components in all three bands, the radar detector will only be able to respond in one of those bands.

Which poses something of a problem if our objective is to test a multiband radar detector in all of its operating bands. What's really needed is not a single oscillator frequency which is a subharmonic of all three bands, but rather three separate oscillator frequencies, each of which produces a harmonic in only one of the bands of interest. Furthermore, the three frequencies should be close enough together to permit them to be generated in the same oscillator circuit, simply by retuning.

The solution chosen was a varactor-tuned oscillator centered on 2650 MHz, with ±25 MHz of tuning range. This oscillator produces three "magic frequencies." X-band is tested with an oscillator frequency of 2625 MHz — the fourth harmonic falls in X-band, the ninth harmonic falls just below the K band allocation, and the thirteenth harmonic is just below the Ka band allocation. K-band is tested by tuning the oscillator to 2675 MHz — the fourth harmonic is now too high for X band, the ninth harmonic falls within the K band allocation, and the thirteenth harmonic is just above the Ka band allocation. And, Ka band is tested with an oscillator frequency of 2645 MHz — the fourth harmonic is too high for X band, the ninth harmonic is just below the K band allocation and the thirteenth harmonic is just within the police radar Ka band allocation. These numbers are summarized in Table 1.

We now come to the problem of circuit implementation. Figure 6 shows the final schematic. Notice that the final product has three push buttons, one to activate each band. Tuning is accomplished by adjusting a potentiometer to properly bias a varactor for each of the three frequencies. And since the etched microstrip antenna need only radiate signals in the rather narrow 2.625-2.675 GHz range, not their harmonics, its bandwidth is not a problem.

Of course, the input circuit of the radar receiver is a waveguide beyond cutoff, as far as the test signal is concerned. Its loss at 2.6 GHz can be predicted, and compensated for in the link budget. Path loss is fortunately minimal, since: 1) we expect the RadaRanger to be held close to the input of the receiver, and 2) it is an S band signal, rather than X, K or Ka

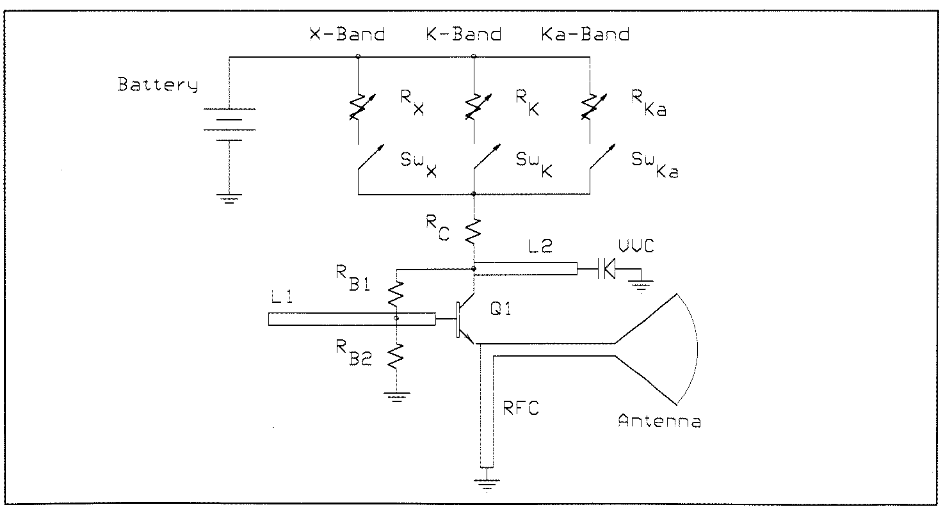


**Figure 6. The RadaRanger™ multiband radar detector tester.**

band signals, which must be radiated. Link analysis and bench testing confirm that a 1 mW oscillator near 2.65 GHz will generate internal to the receiver, at a range of 12 cm, ample fourth, ninth and thirteenth harmonics to readily trigger a typical multiband radar detector in all three bands.

The chief problem associated with the original radiated comb solution to multiband radar receiver testing is its spectral inefficiency. In order to test a receiver at three discrete frequencies, it was necessary to generate a comb of not less than 23 separate frequency components. Obviously, FCC Part 15 radiation testing was something of a problem. But by generating a single, spectrally pure RF signal and generating harmonics in the input circuit of the receiver under test, FCC radiation compliance is virtually assured.

By the time you read this, the design concept presented in this article will be available for commercial licensing. Please contact the author or the patent assignee, listed in Reference 1. *RF*

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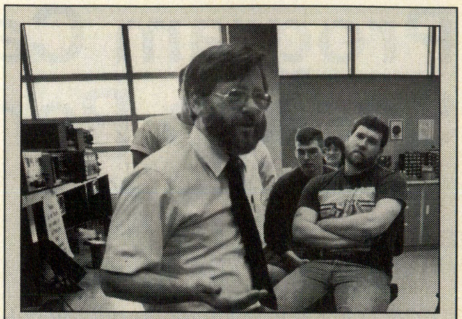


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# Program Calculates ECM System Performance

*By Ronald G. Day*
*ITT Avionics Division*

*ECMTOOLS was written to quickly calculate and plot the pertinent factors which define airborne ECM system performance against airborne or ground based radar-directed weapons systems. The calculations involved with performance predictions are well known and are often folded into a more complex program. However, such programs can be tedious to set up and are unnecessarily complex for a simple "heads-up" consideration of candidate designs during the system conceptual phase. This simple evaluation is the goal of the program described here.*

Two of the key factors in the design of an airborne ECM system are the radar RF signal level at the jammer receiver, and the Jamming-to-Signal (J/S) ratio at the threat radar receiver. In general, the ECM system receiver must be sensitive enough to see the threat radar signal well beyond the lethal range of the associated weapons system. In addition, the jammer must have enough power to provide a sufficiently high J/S ratio to effectively disrupt the radar operation over the full lethal range of the weapons system.

Using ECMTOOLS, nine separate charts can be plotted in semi-log format which show J/S for self-protect, escort, and stand-off (transponder/saturated) jammers, J/S for linear jammers, radar signal level at the jammer receiver, jammer signal level at the radar receiver, reflected signal level from the aircraft at the radar receiver, path attenuation, and target gain.

All plots provide auto-ranging on both the X and Y axes. Range units can be in nautical miles, kilometers, kiloyards, statute miles or kilofeet. Power units can be either in dBW or dBm. Up to five variable values can be selected. Context specific help screens are provided for each of the nine plots as well as an introductory help screen and a graphics help screen.

## The Radar Range Equation

ECMTOOLS uses variations of the familiar radar range equation [1]. First, the jammer signal level at the radar receiver is:

$$J = \frac{Pj\,Gj\,Gt\,\lambda^2}{(4\pi)^2 R^2} = Pj\,Gj\left[\frac{\lambda^2}{(4\pi)^2 R^2}\right]Gt$$

where,

J= Jammer power at radar receiver (watts)
Pj = Jammer power (watts)
Gj = Jammer transmit gain (ratio)
PjGj = Jammer ERP
λ= wavelength (meters)
R = Range from jammer to radar (meters)

$$\left[\frac{\lambda^2}{(4\pi)^2 R^2}\right] = \text{Path Loss}$$

Gt = Radar antenna gain (ratio)

Converting to a log expression (with dBm and nmi):

$$J(dBm) = PjGj(dBm) - 20\log R(nmi) - 20\log F(GHz) + Gt(dB) - 97.801$$

Next, the reflected signal level at the radar receiver is:

$$S = \frac{Pt\,Gt^2\,RCS\,\lambda^2}{(4\pi)^3 R^4}$$

$$= Pt\,Gt\left[\frac{RCS\cdot 4\pi}{\lambda^2}\right]\left[\frac{\lambda^2}{(4\pi)^2R^2}\right]\left[\frac{\lambda^2}{(4\pi)^2R^2}\right]Gt$$

where,

S = Reflected signal level at radar receiver (watts)
Pt = Radar power (watts)
Gt = Radar antenna gain (ratio)
RCS = Radar Cross Section of aircraft (sq. meters)

$$\left[\frac{RCS\cdot 4\pi}{\lambda^2}\right] = \text{Target Gain}$$

λ = wavelength (meters)
R = Range from jammer to radar (meters)
Note *two-way* path loss

Expressed in log form:

$$S(dBm) = PtGt(dBm) + RCS(dBsm) - 40\log R(nmi) - 20\log F(GHz) + Gt(dB) - 174.146$$

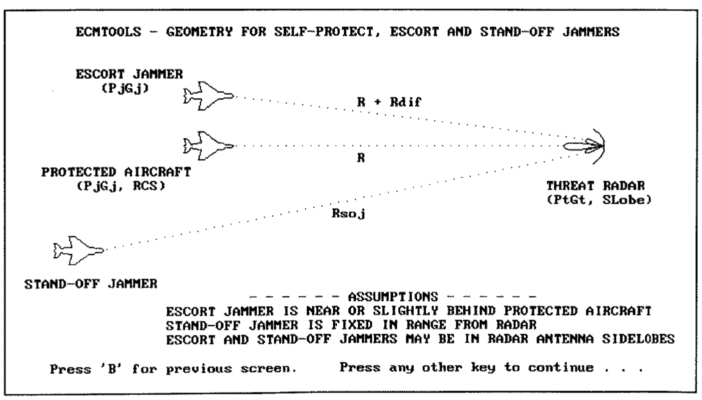



**Figure 1. Geometry for the various jammer types.**

MAIN MENU

J/S CALCULATIONS
1. J/S (SELF PROTECT JAMMER)
2. J/S (ESCORT JAMMER)
3. J/S (STAND-OFF JAMMER)
4. J/S (LINEAR REPEATER)

RF LEVEL CALCULATIONS
5. JAMMER SIGNAL LEVEL AT RADAR (J)
6. A/C SKIN RETURN AT RADAR (S)
7. RADAR SIGNAL LEVEL AT JAMMER (Sr)

MISCELLANEOUS CALCULATIONS
8. PATH ATTENUATION
9. TARGET GAIN

ENTER SELECTION: ?

H = HELP B = BACK

**Figure 2. Main menu screen.**

```
J/S at Radar
   (Self-Protect Jammer)

PtGt     =   100    dBm        Power Units: dB<m>, dB<W>
PjGj     =   60     dBm
RCS      =   Var.   dBsm
Rmin     =   .2     nmi        Range Units: <n>mi, <k>m, k<y>d, <s>mi, k<f>t
Rmax     =   80     nmi

No. Var. =   3      <1-5>
RCS(1)   =   0      dBsm
RCS(2)   =   10     dBsm
RCS(3)   =   20     dBsm

          PRESS C TO CONTINUE

          H = Help    V = Select New Variable    X = Deselect Variable
```

**Figure 3. Data Entry Screen for J/S (Self-Protect Jammer)**

J/S can be derived by simply subtracting S(dBm) from J(dBm):

$$J/S(dB) = PjGj(dBm) - PtGt(dBm) + 20\log R(nmi) - RCS(dBsm) + 76.345$$

Similarly, the received power from the radar at the jammer receiver can be calculated as follows:

$$Sr(dBm) = PtGt(dBm) - 20\log R(nmi) - 20\log F(GHz) + Gj(dB) - 97.801$$

where Gj is the combined jammer receive antenna/distribution gain.

The balance of the equations used for the various plots are all derived from these basic relationships. The specific equations are shown at the bottom of each of the data output screens. The variables are defined in each of the associated help screens.

## The Program

ECMTOOLS is written in Quick Basic 4.5 which also runs on QBasic supplied with DOS 5.0. The program should run on any MS-DOS computer. In order to plot the graphical data, a CGA, EGA or VGA display is required. Also, an appropriate 'Print Screen' driver such as GRAPHICS.COM must be installed prior to running the program. You will need to check for proper operation of the DOS GRAPHICS.COM, display mode and printer type; operation may be different for the various combinations.

The program runs in a straightforward manner. When first started, the program checks to see if an EGA or CGA monitor is present. The result is displayed on the introductory screen. The second screen is a graphical display showing the geometries for self-protect, escort and stand-off jammers. The next screen is the main menu which allows selection of the particular calculation to be performed. Next is the selected data entry screen. The final screen is the graphics screen which can be printed using the 'Print Screen' key.

The screen shown in Figure 1 depicts the geometry for self-protect, escort and stand-off jammers. Note that the escort and stand-off jammers may or may not be in the sidelobe region of the radar antenna pattern. Pressing any key (except 'B') accesses the main menu.

The main menu has nine selections as shown in Figure 2. The desired selection is made by typing the appropriate number followed by 'Enter'. Typing 'H' in lieu of a number accesses a general help screen while 'B' backs up to the previous screen.

If '1' is typed at the main menu, the data entry screen for J/S (Self-Protect Jammer) is displayed as shown in Figure 3. Note that representative default values are shown which are readily edited. Data is entered one line at a time and the cursor can be moved up, down and sideways using the arrow keys, backspace key and the enter key. In this example, RCS is selected as a variable and three variable values are shown as the bottom three values. Up to 5 values of the selected variable specified. PtGt or PjGj can be selected as the variable in lieu of RCS by placing the cursor on that line and pressing 'V'. All variables can be deselected by entering 'X'.

Note that the power and range units can be changed with the cursor at any position by typing 'W' for dBw, 'M' for dBm, 'N' for nautical miles, 'K' for kilometers, 'Y' for kiloyards, 'S' for statute miles and 'F' for kilofeet. Letters can be in upper or lower case. Typing 'H' accesses an application specific help screen.

Pressing 'C' or scrolling down to the bottom of the screen causes an 'Addi-

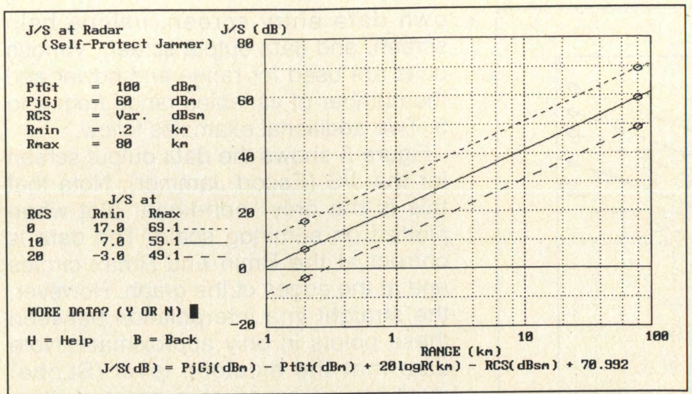



**Figure 4. Output screen for Self-Protect Jammer.**

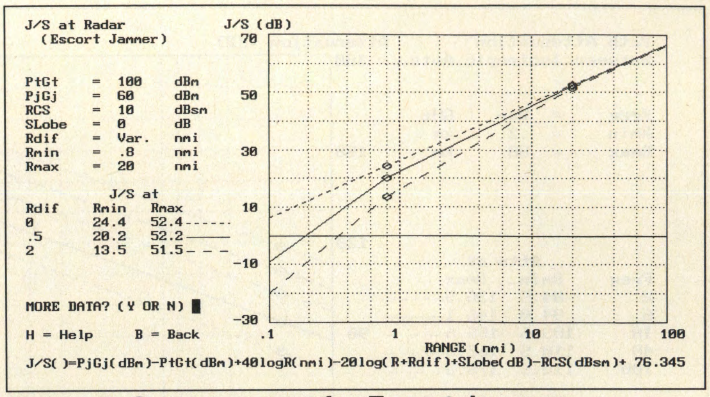



**Figure 5. Output screen for Escort Jammer.**

tional Changes? (Y or N)' to be displayed. Pressing 'Y' followed by 'Enter' returns the cursor to the top of the data entry screen. Pressing 'N' accesses the data output screen. 'H' accesses the application specific help screen and 'B' backs up to the previous screen.

Figure 4 is the data output screen for the J/S (Self-Protect Jammer) selection. The data and variables are tabulated on the left side of the screen and the plotted data is on the right in semi-log format. The X-Axis is plotted to the nearest full log cycle and the Rmin and Rmax variables are shown as circles on each variable line. The Y-Axis is also auto-ranged rounded to the nearest 10 dB. The equation used for the calculations is shown at the bottom of the screen and reflects the particular power and range units selected on the data entry screen.

If the appropriate GRAPHICS.COM (or equivalent) printer driver is installed, the graph can be printed simply by pressing 'Print Screen' (or 'SHIFT-PRT SCRN' depending on your computer). If you have an EGA or VGA monitor, the default display is in color. However, some print drivers work better with a monochrome display (particularly DOS 5.0). Thus prior to printing, you can change the screen mode to mono-EGA by pressing 'M' or to CGA by pressing 'C' or return to color EGA by pressing 'E'. The graphics help screen can be accessed by pressing 'H'.

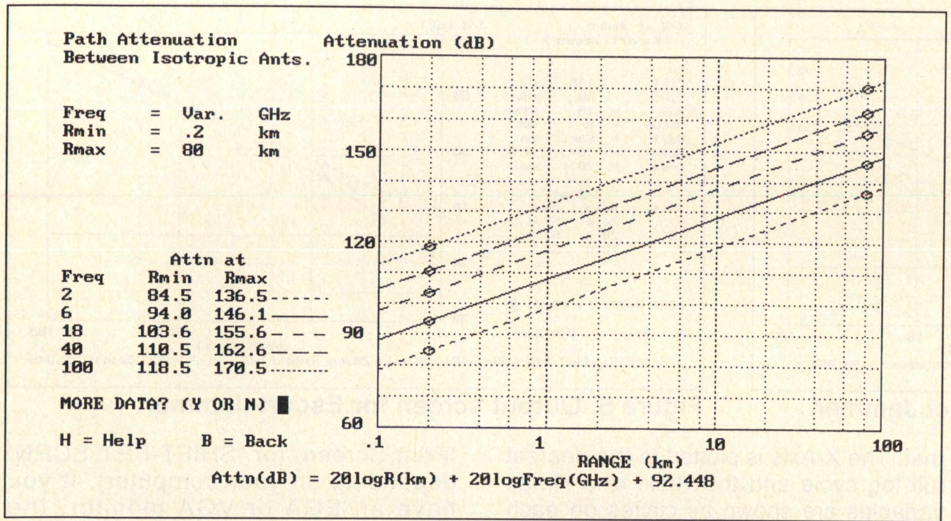



**Figure 6. Data output screen for path attenuation calculations.**

After the data is plotted, there is a 'MORE DATA? (Y OR N)' prompt displayed. By pressing 'N', the program ends after a confirmation prompt. By pressing 'Y', you are returned to the previous data entry screen. Note that variable data previous entered remains so when you want to do a series of iterations, you only have to enter the new data.

The balance of the main menu selections access screens that are similar to that for the J/S (Self-Protect Jammer) described above. Each selection has its own data entry screen, unique help screen, and data output screen. Various units are used for range and power and the number of variables range from 0 to 5. Two additional examples follow.

Figure 5 shows the data output screen for the J/S (Escort Jammer). Note that this is the only non-linear plot when plotted on semi-log scale. The data is correct at the Rmin and Rmax circles and at the edges of the graph. However, the straight line interpolation between these points in only approximate. Note also that the sidelobe level (SLobe) must be expressed as a negative number (dB below the main beam). If it is entered incorrectly as a positive number, it is converted to a negative number at the 'ADDITIONAL CHANGES? (Y OR N)' prompt. Also, Rdif is the distance the escort jammer is behind the penetrating aircraft. If a negative number is entered for Rdif, it also is converted to the correct sign before plotting. Figure 6 shows the output screen for path attenuation between isotropic antennas. Note that only frequency can be selected as a variable.

## Programming Note

A mention should be made regarding the graphics modes used in the program. The monitor autoselect routine first selects the CGA mode (SCREEN 2) and then the EGA mode (SCREEN 9). If there is an error when SCREEN 9 is selected, the program reverts back to the CGA mode. The text mode SCREEN 0 is used for all text screens. The overall hardware requirement then for graphics output is for a monitor with CGA or EGA graphics modes. Most VGA graphics adapters support EGA although one was found that caused the system to crash at the SCREEN 9 command. It did work satisfactorily in the CGA mode, however.

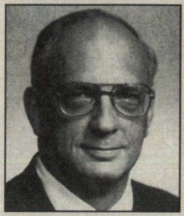
In order to accommodate the occasional incompatible system, all the graphics default settings are at the beginning of the program. Thus by modifying several basic statements the autoselect code can be disabled and the mode fixed at CGA, EGA, color or no color, or any other graphics mode which supports an 80x25 text display. VGA (SCREEN 12) can not be used since it has a 80x30 text display.

*This program is available from the RF Design Software Service; see page 118 for ordering information.* **RF**

## Reference


1. Leroy B. Van Brunt, Applied ECM, EW Engineering, Inc., Dunn Loring, VA, 1978, Chapter 3.


### About the Author




Ron Day is a Technical Consultant with ITT Avionics Division. His entire career has been spent in RF and microwave EW/ECM system design. He has a BSEE from Purdue University and a MSEE from Newark College of Engineering. He can be reached at ITT Avionics, Dept. 728, 100 Kingsland Rd., Clifton, NJ 07014; tel. (201) 284-4090.

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# A Program for Design and Analysis of Receivers

*By John Donohue*
*Naval Research Laboratory*

*This article describes an easy to use program written to assist engineers in receiver design and analysis. It computes noise figure, gain, gain compression problems, input and output third and second order intercept points, spurious free dynamic range, minimum detectable signal levels and signal to noise ratio including temperature compensation for small signal amplifiers. The program's intended use is to allow the engineer to explore numerous combinations of components and see the results before going beyond the initial design and prototype stages.*

Often, RF Engineers are faced with the task of designing an RF system, choosing the components, building the design and finally testing what they've conceived. In the design process, specification changes and component variations can yield a long and arduous iteration process before finally arriving at an acceptable solution. In the past, many of the tedious, but necessary, computations associated with receiver design would take many hours to complete. This design program reduces time-consuming number crunching and takes virtually no time to learn how to use, allowing the engineer to focus his attention on more important issues.

The ideal receiver would, of course, have 0 dB noise figure, very high intercept points (30 to 50 dBm), and no spurious responses in excess of the thermal noise level in the most narrow available channel bandwidth in the receiver. Physical reality dictates that this is not attainable. In most applications, the most important characteristics of a receiver are sensitivity and dynamic range. Other characteristics such as noise figure, gain and second and third order intercept points are good measures of whether the first two characteristics are meeting specifications. Initial selections of component parameters closest to the front end are then compared to specific RF goals. Stages further into the secondary IF may be chosen and evaluated at a later point. Once a baseline design has been chosen, performance characteristics can be evaluated, changed and re-analyzed repeatedly. This RF receiver program streamlines this task.

## Theory

Several basic RF concepts are covered in the program. The following is a basic tutorial of the theories behind the equations.

Probably the most well known formula addressing noise theory is Friis' cascaded noise figure formula:

$$\text{n.f.} = nf_1 + \frac{nf_2 - 1}{g_1} + \frac{nf_3 - 1}{g_1 g_2} + \cdots + \frac{nf_n - 1}{g_1 g_2 \cdots g_{n-1}} \quad (1)$$

$$nf_i = 10^{(Fi/10)} \quad (2a)$$

$$g_i = 10^{(Gi/10)} \quad (2b)$$

$$F = 10 \log (\text{n.f.})\ \text{dB} \quad (3)$$

In the program, up to 50 devices can be entered in the chain. All values are entered in terms of dB and then converted by the program using eqs. (2a) and (2b). Friis formula in eq. (1) is applied to the cascade and then converted back into decibels using eq. (3).

Temperature fluctuations can greatly affect overall system performance. Although nominal values at room temperature may meet specifications, the same may not be true at elevated temperatures, which becomes important when designing good safety margins into the system's worst case analysis.

The gain of a typical GaAs FET small signal amplifier that does not include special compensating circuitry decreases as operating temperature is increased. Over the temperature range of –55C to +75C, the temperature coefficient of an uncompensated amplifier is approximately –.015 dB/C/stage. The program assumes .015 dB as a worst case. Feedback stabilized GaAs FET amplifiers typically have a temperature

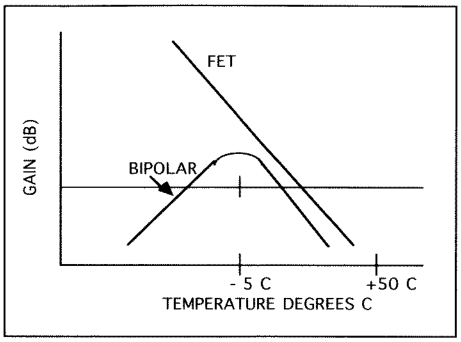



**Figure 1. Gain vs. temperature for amplifiers.**

coefficient of approximately -.005 dB/C. Bipolar amplifiers have a curve characteristic in which gain increases and noise figure decreases from 23C to around –5C and then drops back off again. Above 23C, its temperature coefficient essentially follows the behavior of a GaAs amplifier (see Figure 1).

Generally, the gain variation with temperature works linearly in the other direction as temperature drops below 23C. However, as the temperature approaches –10C and beyond to –55C, noise figure improvement tends to reach an absolute minimum. At the point the program reaches this maximum improvement, it no longer compensates for any further decreases in temperature.

## Sensitivity and Dynamic Range

For computing receiver sensitivity, the MDS (minimum detectable signal) determines the signal level just above the noise level.

$$\text{MDS} = -114\ \text{dBm} + 10\log(\text{NBW}/1\text{MHz}) + \text{NF} \quad (4)$$

where,

NF = cascade's noise figure
NBW = system noise bandwidth in MHz
–114 dBm = kT in 1 MHz BW

The spurious free or true dynamic range of a system is usually defined as the range of input signals over which

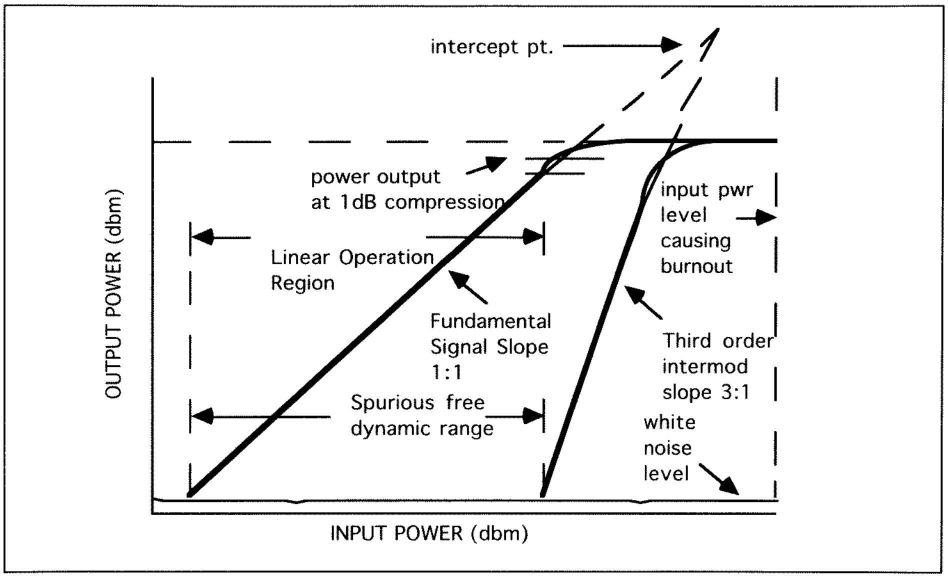



**Figure 2. Linear region and third-order intercept point (IP3).**

spurious outputs are below the noise level of the output (Figure 2). The following equation computes the spurious free dynamic range of the system at the output:

TDF = (2/3) (IP3 – GAIN – MDS) (5)

where,

IP3 = third order output intercept point of the system
GAIN = total gain of the cascade
MDS = minimum signal the cascade can detect.

The typical measure of performance for any order of intermodulation is the intercept point for that order. If we assume that a signal traversing the amplifiers and other components, encounters no phase shift, we may calculate the composite IM performance by assuming in-phase addition of the individual contributions.

The equation for calculating the third order intercept point for a cascade [3], as referenced from the output, is :

$$Ioc^{(3)} = -10\log\left[\sum_{j=1}^{N}\left[Ioc_j^{(3)} g_{j+1}\cdots g_N\right]^{-1}\right] \quad (6a)$$

The formula for calculating the second order intercept point is:

$$Ioc^{(2)} = -20\log\left[\sum_{j=1}^{N}\left[Ioc_j^{(2)} g_{j+1}\cdots g_N\right]^{-1/2}\right] \quad (6b)$$

Note that the superscript terms are not exponents, but indicators of the order of intercept point. Generally, intercept points are given in dBm. All values of intercept points in the program are entered in terms of dBm. All components must have an intercept point entered for them including passive devices in which case 100 dBm or more would be appropriate. Calculations assume a 50 ohm system.

The intercept point can also be calculated as referenced from the input of the cascade. To illustrate this, suppose we have two amplifiers in cascade shown in Figure 3, with their critical parameters of gain, 2nd order intercept point and third order intercept point.

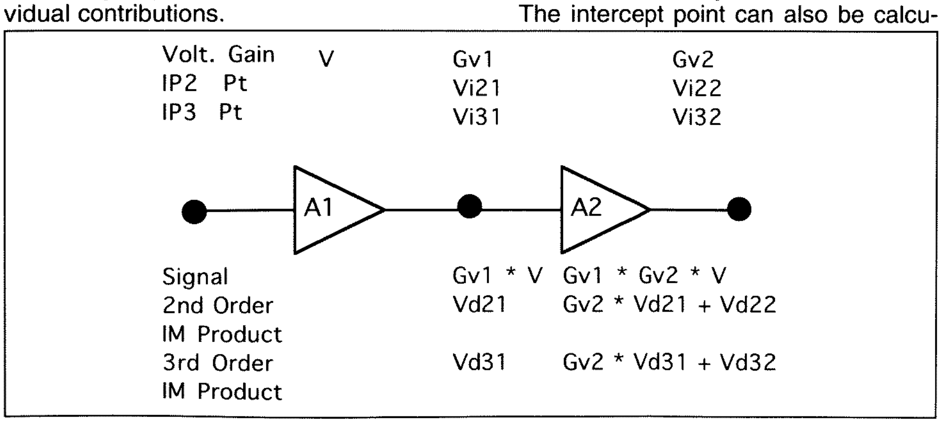



**Figure 3. Example of input intercept calculation.**

Amplifier A1 generates a 2nd order product Vd21 which is present at the input of A2. A2 subsequently generates its own product of Vd22. Therefore if we linearly sum the overall intermodulation product at the output of A2, the total is Gv2 * Vd21 + Vd22 as shown in Figure 3. We can then refer this signal to the input by calculating it relative to the gain of the entire cascade.

$$V_d = \frac{G_{v2} \bullet V_{d21} + V_{d22}}{G_{v1} \bullet G_{v2}} = \frac{V_{d21}}{G_{v1}} + \frac{V_{d22}}{G_{v1} \bullet G_{v2}} \quad (7)$$

Equation (7) then yields what the signal voltage would be if an interfering signal were at the input at the cascade. At the intercept point then, the interfering signal is equal to the input voltage Vi2. The procedure for finding the third order intercept point is similar to that for the second order except that $V_{d3} = V^3/V_{i3}^2$. After collecting terms as above we find that the equation becomes:

$$\frac{1}{V_{i3tot}} = \left[\frac{G_{v1}}{V_{i31}} + \frac{(G_{v1} \bullet G_{v2})}{V_{i32}}\right] \quad (8)$$

Now converting to decibels:

$$IP3\backslash(dB) = 10\log(V_{i3tot}) \quad (9)$$

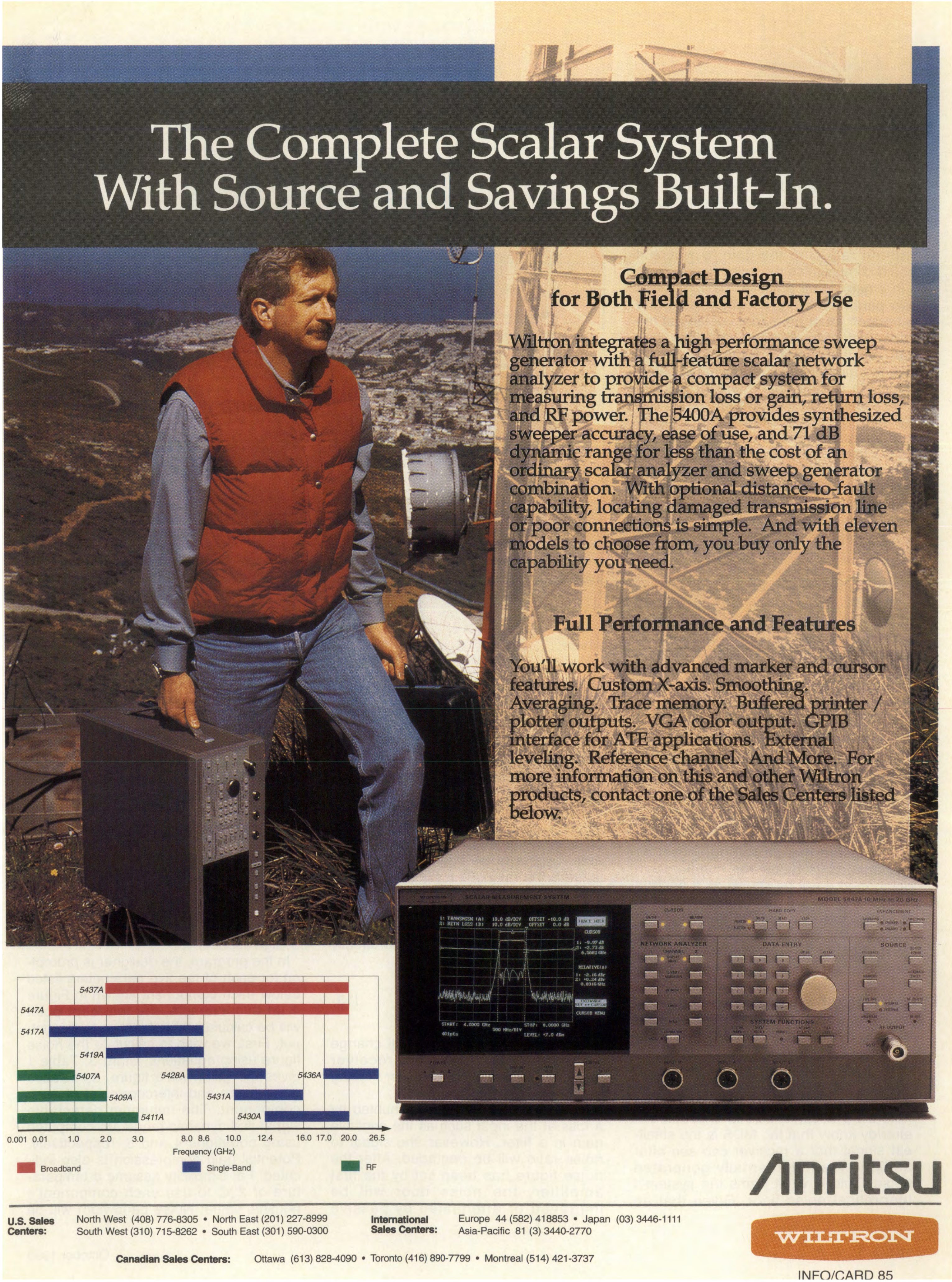
Of course, the output intercept point may also be related to the input simply by:

OIP = IIP + G (10)

where,

OIP = Output intercept point of cascade
IIP = Input intercept point of cascade
G = Total gain of cascade

These equations show that the greater the gain to the indicated point, the more important it is to have a high intercept point. To reduce problems due to intermodulation products, selective filtering should be used as near the front end of the receiver as possible to reduce or eliminate signals likely to cause these products. Often times, however, the designer is restricted by how much insertion loss he can allow while



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achieving the desired filter attenuation. In this instance, both the insertion loss and the first amplifier's gain must be kept to the desired minimum, to avoid raising intermodulation products beyond what can be tolerated by the system, but without reducing the gain to where it could compromise the noise figure.

In small signal amplifier design, these two parameters are directly related and a tradeoff must be made. Looking at the cascaded intercept point equations, one can see that the last stage in the chain has the most influence on what is determined to be the final intercept point of the cascade. Bearing this in mind, it can be said that while choosing an amplifier's intercept point for the first stage is fairly critical, it is also important to carefully select the later stage intercept points. In some instances, 5th order or even 7th order products can become significant much later down the chain of a receiver, depending on the designer's choice of amplifier and filter parameters.

## 1 dB Compression Point

All active devices as well as mixers have what are called $P_{1db}$ compression points. This is defined as the point at which the power gain is down 1 dB from the ideal gain. In order to maintain linearity, operation above the gain compression point must be avoided. Since only active devices and mixers have actual compression points, choice of the specification for passive devices is dictated by how much power the device is rated to withstand, or how much power is realisitically expected to be put through the system.

If a component is in compression, its compression margin will be shown as negative. If its below the compression point, then it will be shown as a positive margin at the program output. A negative compression margin message will be printed if any device is operating in its non-linear region. Before becoming alarmed, be sure that the dynamic range entered coincides with the automatic gain control setting. If this is not the problem, some components may need to be re-evaluated.

## Real-World Noise Performance

When performing system calculations in receivers, the engineer must take nature's pratical limits into account. We already know that the MDS is the smallest signal that a receiver can see after adding external internally generated noise. This is what forms the system's theoretical noise floor. Given that an

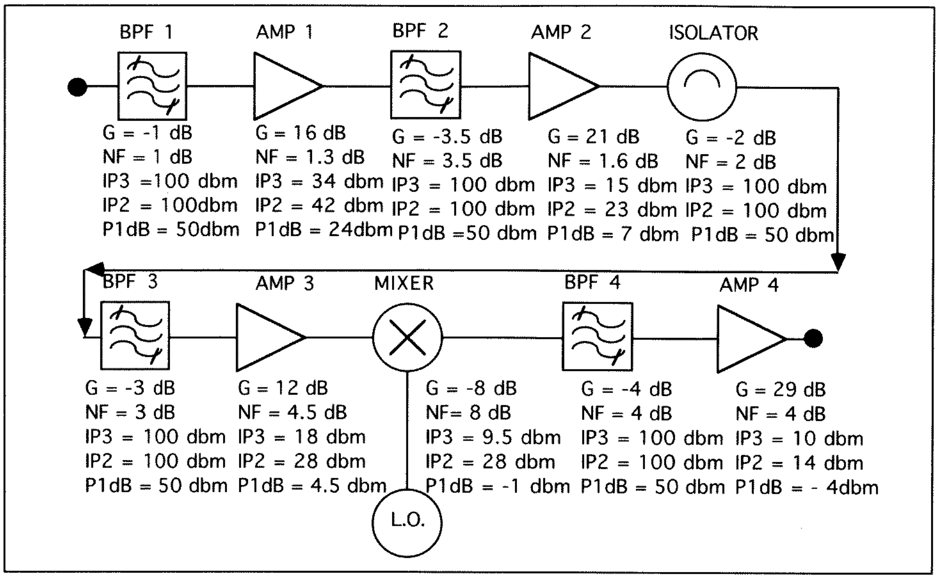


**Figure 4. Receiver design example**

antenna external to the receiver is at some noise temperature $T_a$, we find that the external noise per unit bandwidth is:

$$No = kT_a \qquad (11)$$

where,

$k = 1.38 \times 10^{-23}$ J/K (Boltzman's constant)
$T_a$ = antenna or source noise temperature

Noise power into the receiver then is set by the input preselector bandwidth, $B_n$:

$$N = kT_aB_n \qquad (12)$$

The receiver adds to this noise, however, by the noise figure F. The noise figure is the ratio of the signal to noise in over the signal to noise out. For a single component or a system with uniform bandwidth from input to output, the signal to noise ratio can be found directly from the following formula.

$$F = \frac{(S/N)_i}{(S/N)_o} \qquad (13)$$

However, the noise floor will change as a signal traverses the receiver through narrower and narrower filtering bandwidths.

Noise power cannot be attenuated by a loss at the input such as the negative gain in a filter. However, the signal to noise ratio will be degraded. After the noise figure has been set by the first amplifier, the noise floor will be improved or attenuated by passive devices, but the signal to noise ratio will not improve unless the bandwidth is narrowed. For an active device with gain, the noise floor will worsen (i.e. rise) by approximately the device's gain while the signal to noise should stay relatively constant. In this case, the device noise figure also degrades the noise floor by its value. After, the first amplifier, the effects of subsequent positive gain stages becomes less significant.

One other aspect of computing signal to noise ratio regards mixers. The program assumes an ideal local oscillator, so the mixer noise floor (output) will be attenuated by the loss of the mixer. The signal to noise ratio, however, will remain constant.

## Design Example

The following is an example of a receiver front end cascaded with its first IF stage. It will illustrate the program's design capabilities as well as the critical design aspects.

In the program, the designer is prompted for each of the five parameters listed below each component shown in Figure 4. In this example, the intercept point will be calculated first relative to the output. First, we wish to calculate the noise figure using equations 1 thru 3. Table 1 gives the gain, noise figure, cumulative noise figure and intercept point of each component. The minimum detectable signal and dynamic range are computed using equations 4 and 5, respectively. Potential gain compression is also indicated. For simplicity assume a temperature of 23C to use each component's nominal value. Noise bandwidth will be

| STAGE NAME | NF | MNGN | MXGN | TOTAL NF | IP3 | TOTIP3 | GCOMP |
|---|---|---|---|---|---|---|---|
| | dB | dB | dB | dB | dbm | dbm | dbm |
| 1 filter1 | 1.00 | -1.00 | -1.00 | 1.00 | 100.00 | 100.00 | 50.00 |
| 2 amplifier1 | 1.30 | 16.00 | 17.00 | 2.30 | 34.00 | 34.00 | 26.00 |
| 3 filter2 | 3.50 | -3.50 | -3.5 | 2.40 | 100.00 | 30.50 | 50.00 |
| 4 amplifier2 | 1.60 | 21.00 | 22.00 | 2.48 | 15.00 | 15.00 | 7.00 |
| 5 isolator | 2.00 | -2.00 | -2.00 | 2.48 | 100.00 | 13.00 | 50.00 |
| 6 filter3 | 3.00 | -3.00 | -3.00 | 2.48 | 100.00 | 10.00 | 50.00 |
| 7 amplifier3 | 4.50 | 12.00 | 14.00 | 2.49 | 18.00 | 16.54 | 4.50 |
| 8 mixer | 8.00 | -8.00 | -7.00 | 2.49 | 9.50 | 5.99 | -1.00 |
| 9 filter4 | 4.00 | -4.00 | -4.00 | 2.49 | 100.00 | 1.99 | 50.00 |
| 10 amplifier4 | 4.00 | 29.00 | 31.00 | 2.50 | 10.00 | 9.97 | -4.00 |

COMPUTATION OF INTERCEPT POINTS REFERENCED TO OUTPUT

| TEMP | SYSTEM BW MHz | TOTAL MINGAIN | TOTAL MAXGAIN |
|---|---|---|---|
| 23.00 | 4.00 | 56.50 | 63.50 |

| IP3OUT | IP3IN | IP2OUT | IP2IN | SFDYNRG | MDS |
|---|---|---|---|---|---|
| dbm | dbm | dbm | dbm | dB | dbm |
| 9.97 | -46.53 | 19.07 | -37.43 | 39.30 | -105.48 |

A NEGATIVE GAIN COMPRESSION MARGIN SHOWN BELOW INDICATES A POTENTIAL PROBLEM

| GAIN COMP. MARGIN | | SECOND ORDER INTERCEPT PT | | |
|---|---|---|---|---|
| MXSIG | MNSIG | STAGE | IP2 | TOTIP2 |
| dB | dB | | dbm | dbm |
| 136.00 | 191.00 | 1 | 100.00 | 100.00 |
| 95.00 | 151.00 | 2 | 42.00 | 42.00 |
| 122.50 | 178.50 | 3 | 100.00 | 38.49 |
| 57.50 | 114.50 | 4 | 23.00 | 22.87 |
| 102.50 | 159.50 | 5 | 100.00 | 20.87 |
| 105.50 | 162.50 | 6 | 100.00 | 17.87 |
| 46.00 | 105.00 | 7 | 30.00 | 23.91 |
| 47.50 | 107.50 | 8 | 28.00 | 13.98 |
| 102.50 | 162.50 | 9 | 100.00 | 9.98 |
| 17.50 | 79.50 | 10 | 20.00 | 19.07 |

NO POTENTIAL COMPRESSION PROBLEMS FOUND
NOTE: THIS ONLY CALCULATES COMPRESSION OF IN BAND SIGNALS

**Table 1. Receiver program gain and intercept tabulated output.**

the smallest filter bandwidth in the chain, filter number 4 with a bandwidth of 4 MHz. Filters 1,2 and 3 have bandwidths of 30 MHz. When computing intercept point it is important to understand where the truncation point should be. In this case, amplifier 4 is our last active device and would be the truncation point for determining the intercept point for in-band products within the 4 MHz filter span. If we were interested in what the intercept point is for the filter prior to this, we would truncate the chain at amplifier 3.

Note that 100 dbm has been entered for all passive devices to eliminate their influence in computing the output intercept point.

The program contains a second method of calculating the intercept point except that it is referenced from the input. In this case, the designer need not enter intercept point values for the four bandpass filters or the isolator in the chain because this method compensates for the passive device losses. Not only does the program compute the input intercept point for the entire cascade, but also the input intercept point for each device as the program moves down the chain. The program computes this by assuming that device is the first one in the chain and continuing as before down until the last stage.

With this option, the intercept points are now referenced to the input of each component and the input of the receiver itself. For those stages that do not have a meaningful intercept point (i.e. filters, isolators), their intercept is listed as zero.

The program also has a signal to noise calculation. Antenna noise and sky noise can be specified to obtain an accurate performance characterization.

A tabulation of wide band data results can be created as an added feature to the program. Total SNR given is a combination of wide and narrow band noise. In computing narrowband noise, the noise power is establ;ished by the narrowest bandwidth seen by the system to a selected point. If this narrow bandwidth is equal to the current stage's bandwidth, then this bandwidth is also used to compute the wideband noise as it has been limited by the current component.

## Conclusion

This article describes the critical design parameters of a receiver, namely noise figure, gain, intercept points, dynamic range, minimum detectable signal, gain compression and signal to noise ratio. The basic equations for each parameter have been presented and applied into the receiver program for quick and easy use with an example given. It is hoped that this program will help the designer come relatively close to the desired performance.

One special aspect of the program is that the designer can see the affects of the individual stages with respect to intercept point, noise figure and signal to noise ratio as the signal traverses the system. Also, to help the engineer perform his tasks readily, stage parameters can be individually altered and inserted and/or deleted for quick "what-if" type calculations. Designing RF hardware has always been an iterative process and this program will hopefully reduce that iteration time. Suggestions for enhancing the program are welcome. Good luck!

*This program is available on disk from the RF Design Software Service; see page 118 for ordering information.* ***RF***

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**About the Author**

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# A High Accuracy Phase Shifter Based On A Vector Modulator

By Dominic J. Ciardullo
Brookhaven National Laboratory, AGS RF Group

*Many systems operating at VHF frequencies and below require a method of varying the phase of an RF signal over a full 360 degrees, often requiring broadband operation, as well. While a myriad of phase shifters and modulators exist for this purpose at UHF and microwave frequencies, those available in the LF to VHF range tend to lack range, linear control and accuracy.*

This paper will focus on the concept of the vector modulator, which is capable of providing 360 degrees of unambiguous phase shift over a wide operating bandwidth. In addition, this type of device can provide a linear relationship between the control signal and resulting differential phase shift, while maintaining a constant output amplitude. Although the vector modulator concept is not new, advances in analog signal processing ICs have enabled the RF engineer to apply high speed linear techniques toward construction of phase shifters and modulators for use below UHF.

## Basic Theory

In a linear system, a constant amplitude sinusoidal signal can be represented as:

$$f(t) = Ae^{st} = Ae^{j\omega t} \quad (1)$$

where s is the Laplace transform variable. Since f(t) is a complex function, the amplitude may also be complex;

$$A = Ae^{j\theta} \quad (2)$$

hence

$$f(t) = Ae^{j\theta}e^{j\omega t} = Ae^{j(\omega t+\theta)} \quad (3)$$

We will make use of two basic concepts [1] during the analysis of the vector modulator. The first is that the output of a linear circuit driven with a sinusoid will preserve the frequency of the input signal. The second is that the addition of two equal frequency sinusoids will result in another sinusoid of the same frequency. Combining these ideas, we see that a linear RF circuit (such as the vector modulator) will affect only the amplitude and phase of a sinusoidal input signal; the frequency at any point within the circuit will remain unchanged. In addition, the phase relationship between two signals of equal periodicity remains fixed, irrespective of when it is measured within the RF cycle. Thus for the purpose of analyzing the amplitude and phase of signals within the circuitry of the vector modulator, the $e^{j\omega t}$ term in equation 3 is somewhat redundant. This term can be eliminated by choosing an arbitrary time during the RF cycle with which to make comparisons (say, at t=0). Setting t=0 in equation 3 gives

$$f(t) = Ae^{j\theta} \quad (4)$$

Since the vector modulator is a linear device, the analysis of its operation will use the method of phasors.

The ideal device would provide predictable, unambiguous phase shift of between 0° and 360° via an electrical control signal. In addition, the output amplitude should remain constant, independent of the phase shift. Figure 1 is the block diagram of a device theoretically capable of achieving these goals.

A 90° power splitter is used to decompose the input signal into two equal amplitude quadrature components, I and Q. The amplitude of each component is adjusted by multiplying it by some scaling factor between +1 and −1. The scaled I and Q components are then combined vectorially, resulting in another sinusoid of equal frequency, having, generally, different amplitude and phase characteristics. If the two scaling factors are represented by $\alpha(\theta)$ and $\beta(\theta)$, then the vector sum is:

$$\text{MAG:} \quad \sqrt{\left[\alpha(\theta)\frac{A}{\sqrt{2}}\right]^2 + \left[\beta(\theta)\frac{A}{\sqrt{2}}\right]^2}$$

$$= A\sqrt{\frac{\alpha^2(\theta) + \beta^2(\theta)}{2}} \quad (5)$$

$$\text{PHASE:} \quad \theta = \tan^{-1}\frac{\beta(\theta)}{\alpha(\theta)}$$

By independently adjusting the amplitude of each component, it is possible to obtain a resultant of any phase between 0° and 360°. Since the two signals are

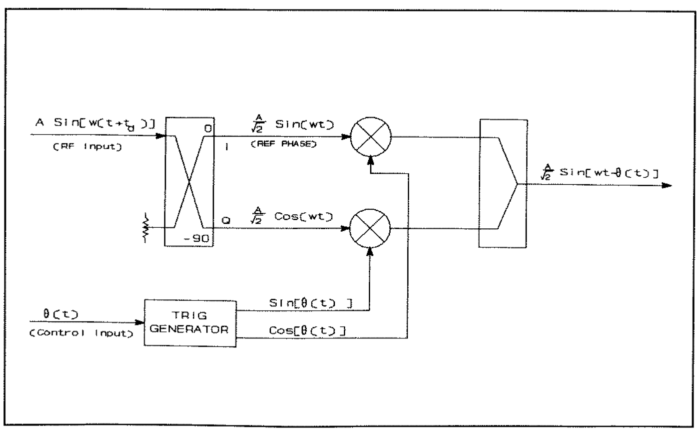


**Figure 1. Basic components of an ideal vector modulator.**

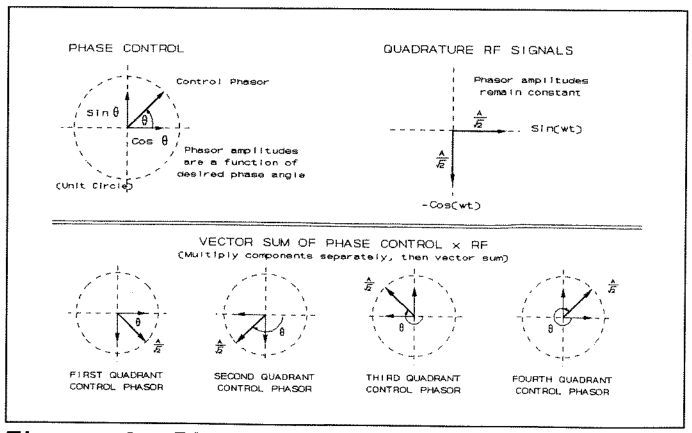


**Figure 2. Phasor representations showing $\cos(\theta)\sin(\omega t) - \sin(\theta)\cos(\omega t) = \sin(\omega t - \theta)$**

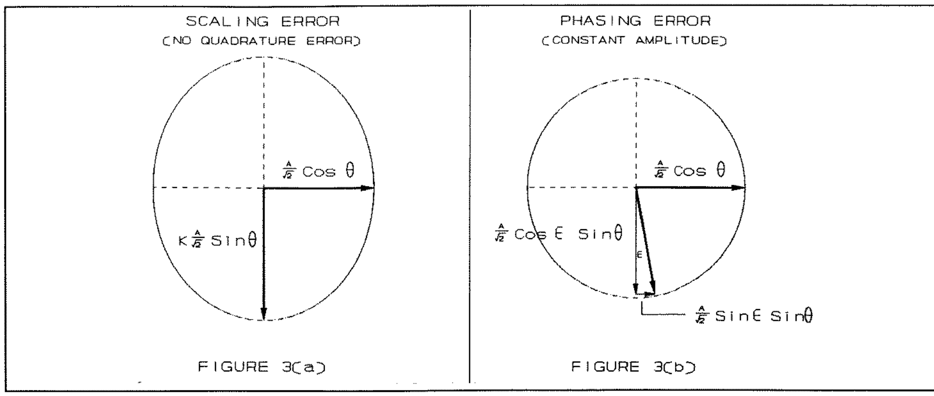


**Figure 3. Phasor diagrams showing the effects of (a) scaling error and (b) phasing error.**

amplitude modulated then vector summed, this phase shifting scheme is often referred to as vector modulation.

To satisfy the requirements of predictable phase shift with constant output amplitude, it is necessary to scale each quadrature RF component in a specific manner. We have effectively fulfilled the first requirement simply by defining $\alpha$ and $\beta$ as functions of the desired phase shift angle. The latter requirement, however, mandates the use of a quadrature pair of scaling functions whose vector sum is a constant magnitude (refer to equation 5). A pair of functions satisfying these criteria may be found from the definition of the unit circle: $\sin^2(\theta) + \cos^2(\theta) = 1$. These two functions are orthogonal, and both depend on the phase shift angle, $\theta$. Substituting $\alpha(\theta)=\cos(\theta)$ and $\beta(\theta)=\sin(\theta)$ into equation 5 we obtain:

$$\text{MAG: } A\sqrt{\frac{\cos^2(\theta)+\sin^2(\theta)}{2}} = \frac{A}{\sqrt{2}} \qquad (6)$$

indicating that the resultant amplitude is independent of the phase shift angle. Simultaneously scaling the amplitude of $\sin(\omega t)$ [the I component] by $\cos(\theta)$, and $\cos(\omega t)$ [the Q component] by $\sin(\theta)$ will therefore result in a constant amplitude sinusoid of predictable phase. Figure 2 illustrates how these scaling factors are used to manipulate the phase of the RF by solving the following trigonometric identity:

$$\cos(\theta)\sin(\omega t) - \sin(\theta)\cos(\omega t) = \sin(\omega t - \theta) \qquad (7)$$

The control phasor, which represents the desired phase shift $\theta$, is decomposed into its quadrature components [$\sin(\theta)$, $\cos(\theta)$]. These control components can now be used to scale the I and Q RF signals, which are also 90° apart in phase. The top half of Figure 2 shows the components of both the phase control signal and the RF input. The quadrature components of both signals are multiplied together, and the resulting components vector-summed to obtain a phase-shifted version of the original RF input signal. Since both the sine and the cosine of the phase shift angle may take on negative values, the output of the vector modulator may have a resultant in any of the four quadrants (0° to –360°). The bottom half of Figure 2 is a sequence of four phasor diagrams illustrating the resultant output for four different phase shift settings. The sequence is from left to right with each diagram representative of the control phasor in a different quadrant. Note that for positive phase shifts the control phasor moves counterclockwise, while the output phasor moves clockwise.

### Practical Limitations

Each individual component shown in Figure 1 makes some contribution to the total amplitude and phase error of the output signal. The trigonometric identity in equation 7 may be used to analyze the extent of these contributions. In words, two quadrature RF signals are scaled as the sine and cosine of the desired phase shift angle, then vector summed. Since we are using phasor analysis, it is natural to separate the total vector modulator output error into two basic categories, scaling (amplitude) inaccuracies and quadrature phase error. We will define scaling error as the cumulative effect of multiplier inaccuracy, lack of trigonometric conformance of the sine and cosine control signals, and any amplitude imbalance from the 90° divider or in-phase combin-

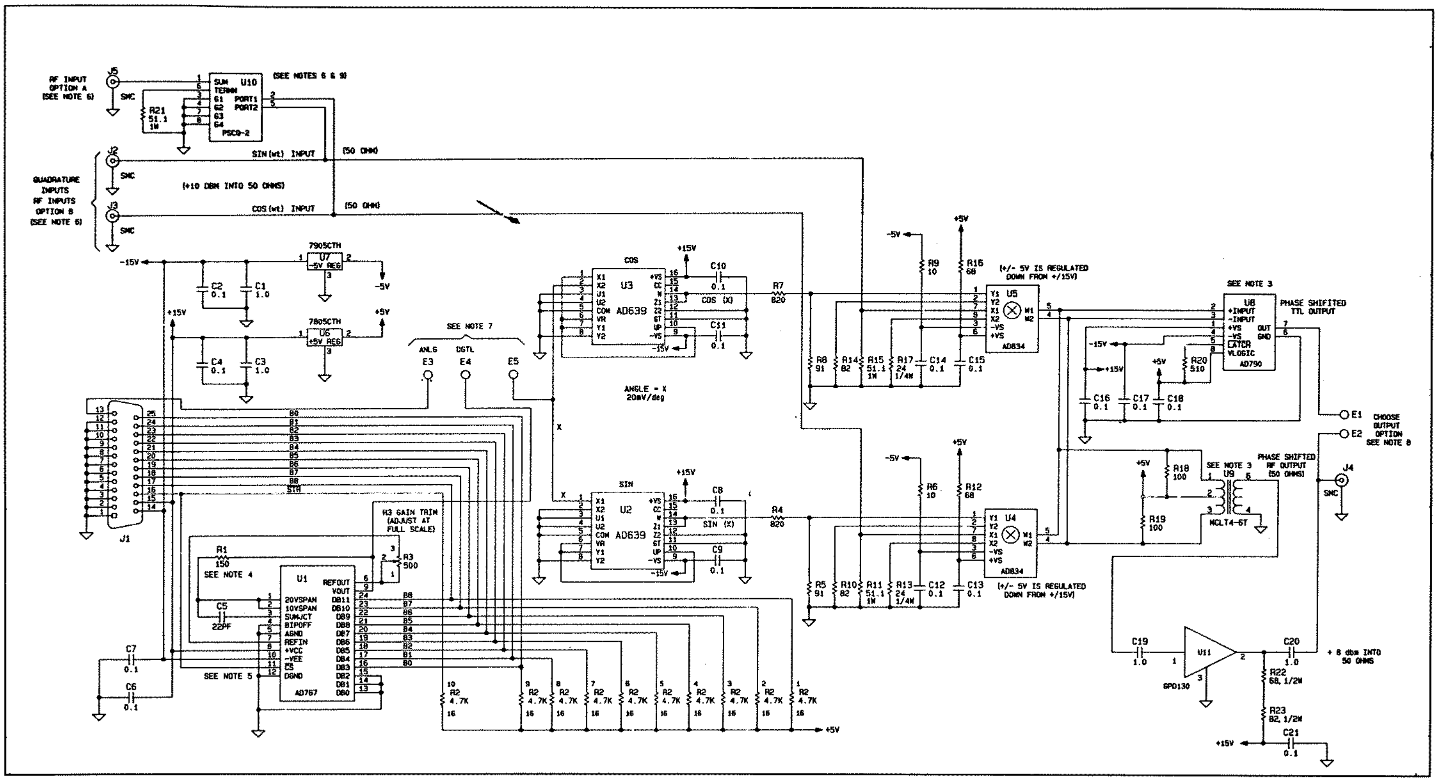

**Figure 4. Schematic of phase shifter prototype.**

er. Quadrature error is primarily due to the 90° power divider, but can also include any phase imbalance of the 0° combiner used to execute the vector-sum. Both categories contribute to the total phase and amplitude inaccuracy of the output signal.

*Effect of Scaling Error* — In principle, the scaling factors of both quadrature RF phasors shown in Figure 2 will vector-sum to a constant amplitude. Each of the major components shown in Figure 1, however, can potentially affect the amplitude of one or both RF signals. The scale factors for each of the phasors is shown in Figure 2:

$$\text{Sin}(\omega t)\ \text{Scale Factor:}\ \frac{A}{\sqrt{2}}\ \text{Cos}[\theta(t)]$$

$$\text{Cos}(\omega t)\ \text{Scale Factor:}\ \frac{A}{\sqrt{2}}\ \text{Sin}[\theta(t)] \quad (8)$$

At this point, the assumption is made that we are predominantly interested in the constancy of the output level and not its absolute amplitude. To this end, Figure 3(a) illustrates a general scaling error, where the ratio of the two phasor amplitudes is given as K. In this case, the two phasors are of unequal amplitude but are assumed to be in perfect quadrature. The effects of cumulative scaling error on the output signal can again be found using vector addition:

$$\text{MAG:}\ A_0[\theta,t,K]$$

$$= \sqrt{\left[\frac{A}{\sqrt{2}}\ \text{Cos}[\theta(t)]\right]^2 + \left[K\frac{A}{\sqrt{2}}\ \text{Sin}[\theta(t)]\right]^2}$$

$$= \frac{A}{\sqrt{2}}\sqrt{\text{Cos}^2[\theta(t)] + K^2\ \text{Sin}^2[\theta(t)]}$$

$$\text{PHASE:}\ \Phi[\theta,t,K] = \tan^{-1}\left[K\ \tan[\theta(t)]\right] \quad (9)$$

where: $A_0$ is the actual output amplitude
$\phi$ is the actual output phase shift
$\theta$ is the desired phase shift

resulting in an output signal $A_0 \sin(\omega t+\phi)$, where both $A_0$ and $\phi$ are assumed to be functions of the desired phase shift, time, and the RF amplitude imbalance. The time dependence is applicable only when the device is used as a phase modulator; For DC phase shifts there is no time dependence, and $\theta(t)$ may be replaced with $\theta$ in the above relations.

If K is chosen to be always greater than unity, then $K(A/\sqrt{2})$ will represent the larger amplitude axis (refer to Figure 3(a)). Equation 9 reveals that the maximum output amplitude error due to this effect occurs at values of $\theta$ which cause the resultant to approach the larger amplitude axis. For example, in the case of Figure 3(a) the maximum error occurs when $\theta$ is either 90° or 270°, since the vertical is the larger of the two axes. The effect of amplitude balance on the output phase error is not quite as obvious, however. If the output phase error $(\phi-\theta)$ is plotted against the requested phase

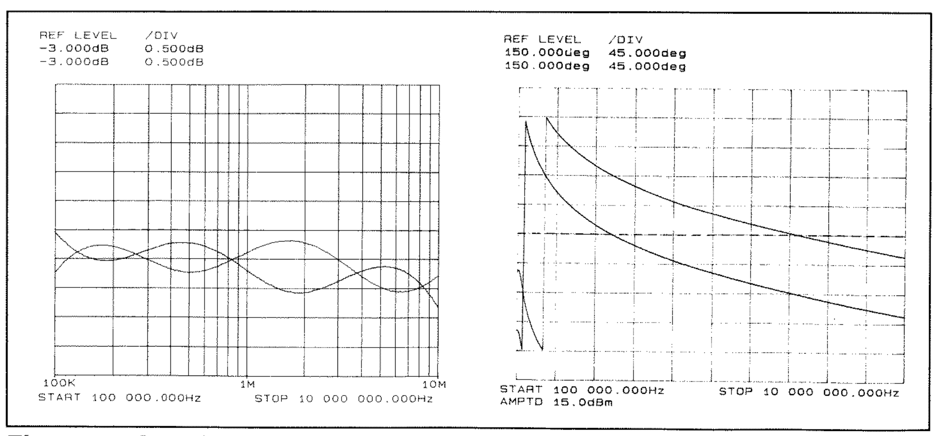


**Figure 5. Amplitude and insertion-phase plots for a Merrimac QH-7-4.9 hybrid power divider.**

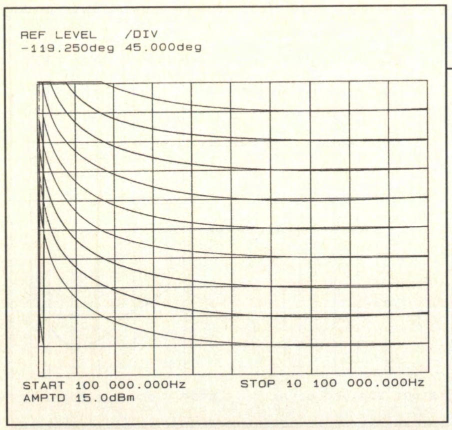


**Figure 6a. Output of phase shifter using QH-7-4.9 hybrid.**

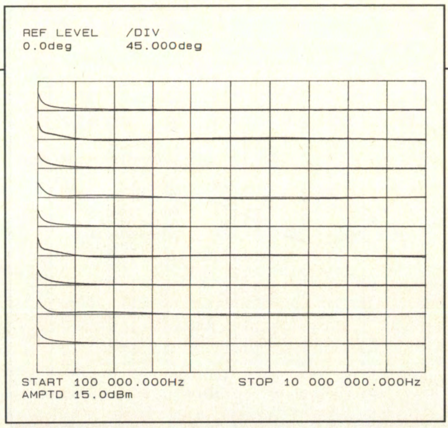


**Figure 6b. Output of phase shifter, normalized to the insertion phase of the QH-7-4.9.**

shift (θ), the maximum absolute phase error will occur near 45° for "small" values of K (i.e., less than 10%).

*Effect of Quadrature Error* — For analysis purposes, phase imbalances from both the 90° divider and the in-phase combiner will be lumped into one effect. Figure 3(b) illustrates a non-quadrature phase division of the vector modulator input signal. The two phasors are assumed to be of equal amplitude and slightly out of quadrature, with ε degrees deviation from 90°. The effects of this type of error on the output signal are:

$$\text{MAG: } A_0[\theta, t, \varepsilon, \omega] = \frac{A}{\sqrt{2}}\sqrt{1+\text{Sin}[2\theta(t)]\ \text{Sin}[\varepsilon(\omega)]}$$

$$\text{PHASE: } \Phi[\theta, t, \varepsilon, \omega] = \text{Tan}^{-1}\left[\frac{\text{Sin}[\theta(t)]\ \text{Cos}[\varepsilon(\theta)]}{\text{Cos}[\theta(t)] + \text{Sin}[\theta(t)]\ \text{Sin}[\varepsilon(\omega)]}\right] \quad (10)$$

where $A_0$, φ and θ are defined as in equation 9. Note that the output signal in this case is a function of the desired phase shift, time, quadrature phase error and frequency. The frequency dependence of ε is included to reflect the non-constant phase vs. frequency characteristics generally associated with broadband combiners and dividers (especially those providing quadrature phase splits). Inspection of equation 10 reveals that the maximum output phase error occurs when θ is at 90°, while the amplitude error is greatest when θ is at multiples of 45°.

### The Constructed Vector Modulator

A schematic for the vector modulator is shown in Figure 4. It was desired to build a general purpose, broadband phase shifter for a variety of RF and instrumentation applications. To achieve this goal, the device was designed with several input, output and control options which provide maximum flexibility while minimizing the amount of specialized hardware required for multiple applications.

The phase shifter can be configured to accept either a single RF input, or separate quadrature inputs. The first option is useful for applications requiring a relatively narrowband, stand-alone phase shifter. The PC board is designed to accommodate standard 90° power dividers packaged in 8-pin relay cans, which are readily available from several RF component distributors. For wider bandwidth applications, the quadrature input option allows the use of an external 90° splitter to provide the quadrature phase relationship of the input signals. In this mode, the device can be used in conjunction with a variety of off the shelf wideband 90° hybrids. This option is also useful when the phase shifter is to be used in a system employing Direct Digital Synthesis (DDS) techniques. Quadrature DDS signals are noted for their superior linear phase characteristics, as well as their excellent amplitude and phase balance. To accommodate both RF input options, the main body of the vector modulator has been designed to operate from 100 kHz to 250 MHz.

Options have been provided for either analog or digital phase shift control. When used in the analog control mode, the device can be used as either a phase shifter or a linear phase modulator with modulation frequencies to 1MHz. The control input may span ±10 Volts, and has the effect of adding (negative control voltage) or subtracting (positive control voltage) 50° of phase shift per volt. When configured for digital

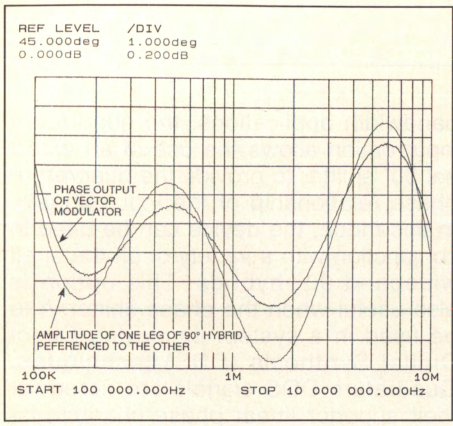


**Figure 7a. Output of phase shifter and relative output balance of the hybrid at 45 degrees.**

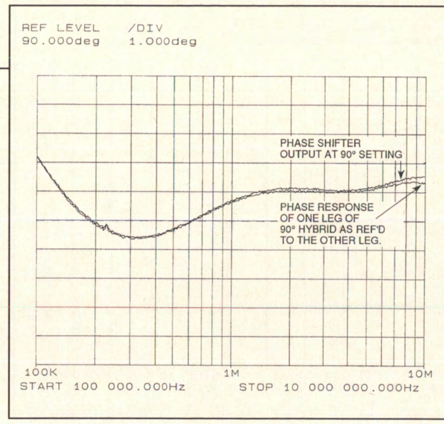


**Figure 7b. Output of phase sifter and relative output balance of hybrid at 90 degrees .**

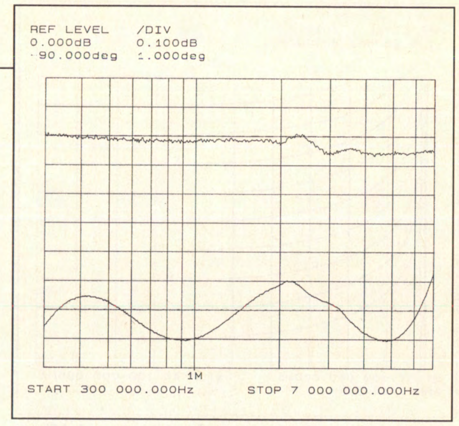


**Figure 8. Amplitude and phase balance of active, all-pass 90 degree phase splitter.**

control, an on-board DAC is used to provide accurate, stable voltage control for phase shifts of 0° (h000) to 511° (h1FF).

The constructed phase shifter also has two output options. The first is the common 50 ohm RF sinusoidal output. The alternative option provides a TTL output for interfacing with digital timing systems. When used in this mode, the main body of the phase shifter operates from DC (when used with DDS quadrature inputs) to >10 MHz.

*Choice of Components* — The circuit element used to decompose the input signal of the vector modulator into equal-amplitude quadrature components depends on several factors. Device characteristics such as amplitude imbalance, phase deviation from 90° and group delay flatness must all be considered before the proper phase splitting element can be selected. The extent to which the two signals are in quadrature affects both the phase accuracy and amplitude of the overall vector modulator (quadrature error). In addition, the amplitude balance of the two quadrature outputs contributes to the overall scaling error, causing additional phase and amplitude inaccuracy.

For applications where flat group delay (deviation from linear phase) is not critical, the most popular circuit element used to phase split the input signal is the 90° power divider. Alternatively, if the phase shifter is supplied with quadrature inputs from a direct digital synthesizer, much higher phase accuracy, phase independent output amplitude, and linear phase can be achieved. For frequencies in the HF band and below, a compromise may be struck by using high speed active all-pass filters to achieve the 90° phase split with excellent amplitude balance and a relatively small amount of phase ripple. The phase shifter constructed has the flexibility of utilizing signals from any of these sources.

As shown in Figure 1, two multiplying elements are needed to amplitude modulate the quadrature RF signals by the sine and cosine of the desired phase shift angle. These devices must be linear over the bandwidth of interest, with minimal DC offset errors. Any gain mismatch between the two multipliers will contribute to the overall scaling error of the vector modulator.

The circuit elements used in the phase shifter are a pair of high speed 4-quadrant multipliers (Analog Devices AD834), chosen for their DC-500 MHz frequency response, high linearity, and ease of use. In addition to their broad bandwidth, these multipliers have two important advantages over the use of current-controlled attenuators [2]. The inputs to the device are voltage driven, eliminating the need for V-I conversion of the sine and cosine modulating signals. Vector summing the outputs of both multipliers is accomplished by simply paralleling their differential current outputs. This current-combining technique helps to extend the bandwidth of the phase shifter, while greatly reducing any phase and amplitude imbalances which might be introduced by using an RF power combiner for the same purpose. The simplified transfer function for the multiplier is $I_O=V_{i1}V_{i2}(4\ mA)$, giving a peak output current of 4 mA for peak input voltages of $V_{i1} = V_{i2} = 1V$. A broadband center-tapped transformer (Mini-Circuits T4-6T, 20 KHz-250 MHz) is used to convert the differential vector sum to a single-ended 50 ohm signal. A 12 dB amplifier (Avantek model GPD-130, TO-39 package, 100 kHz-400 MHz) is used to compensate for the decrease in voltage level affected by the transfer function of the multipliers. This combination of transformer and RF amplifier produces the upper and lower bandwidth constraints for the main body of this phase shifter. These parts can be easily modified, however, for other operating frequency requirements.

When the phase shifter is configured for the TTL output option, the transformer and RF amplifier are not utilized. Instead, a high speed TTL comparator is used as the output driver. The comparator used (Analog Devices AD790) was chosen for its high common mode input voltage, necessary to accommodate the voltages resulting from the quiescent current of the multiplier outputs.

In order to achieve a constant output amplitude signal whose phase is linearly related to the control input, the trigonometric generator block (Figure 1) must provide the sine and cosine of the desired phase shift angle. For the phase shifter described by Figure 4, two high conformance trigonometric function generator ICs (Analog Devices AD639) are used. Given an input voltage V(θ) scaled at 50°/Volt, one IC outputs a voltage which is proportional to sin(θ) and the other a voltage proportional to cos(θ). These ICs offer the flexibility of linear input voltage control, and can be driven via DAC, linear potentiometer or modulating waveform (dependent upon the control option selected). These devices also determine the highest angle modulation rate that can be achieved by the phase shifter.

The digital control option uses a 12-bit DAC with built-in voltage reference. By utilizing only the highest 9 bits, the DAC functions as an accurate, stable voltage input for the trigonometric generator ICs. For "user friendliness" the gain of the DAC is adjusted such that 1 LSB corresponds to 1° of phase shift (DAC output of 20 mV/LSB).



## Results

The prototype phase shifter was configured for digital phase control, quadrature RF inputs (implying the use of an external hybrid or other source of quadrature signals) and 50 ohm RF output. Test results are presented using the external hybrid option to illustrate the sensitivity of the vector modulator to the amplitude and phase balance of its quadrature inputs. To help illustrate this, results will be presented using a commercially available 90° power divider and an active 90° all-pass network. Both hybrids exhibit similar non-linear phase characteristics and phase ripple, but the amplitude balance of the active circuit is an order of magnitude better than that of the passive device. In addition, the active circuit has a slightly smaller overall bandwidth.

The external 90° power divider used for these tests is a Merrimac model QH-7-4.9. Network analyzer plots of the amplitude and insertion phase for this hybrid are presented in Figure 5. Note that the phase response is not linear with frequency. Figure 6(a) shows the output of the phase shifter, using the QH-7-4.9 to supply the quadrature signal inputs. Time delay has been (mathematically) added to the network analyzer to help illustrate the degree of phase linearity (notice the similarity to the phase-frequency response of the hybrid). In Figure 6(b), the output of the phase shifter is normalized to the insertion phase of the hybrid. Once again, a pure delay has been added to the measurement for illustrative purposes. This relative measurement separates the group delay characteristic of the hybrid from that of the main body of the vector modulator. The sharp upward rise at the bottom end of the response occurs because of the low end roll-off of the GPD-130 used to boost the output signal.

Figure 7(a) shows the "worst case" output of the phase shifter for a requested shift of 45°. The relative amplitude balance of the hybrid is overlaid on the same plot to show its effect on the output phase error. Note that any phase imbalance from the hybrid also contributes to the total error at this requested shift, so there is not an exact one-to-one correspondence between the two plots in the figure. Figure 7(b) is a plot of the vector modulator output for a requested phase shift of 90°. The relative phase balance (deviation from quadrature) of the hybrid is overlaid on the same plot to show how closely the output phase error follows that of the 90° divider. These plots, together with equa-

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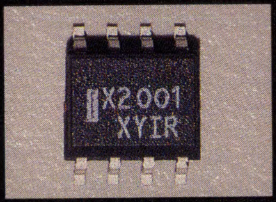


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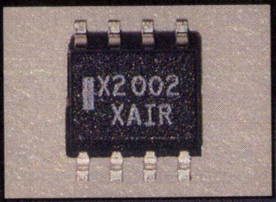


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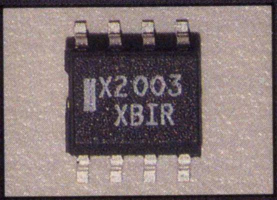


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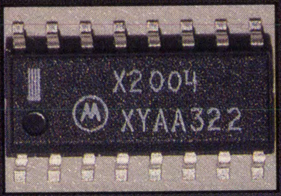


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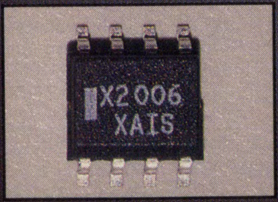


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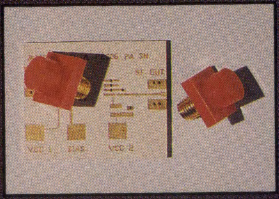


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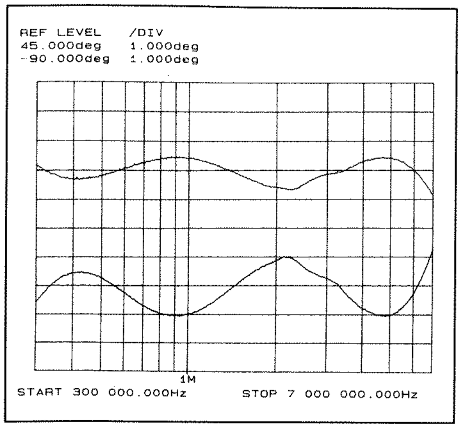


**Figure 9a. Output of phase shifter at 45 degrees, normalized to the insertion phase of the active splitter.**

tions 9 and 10, illustrate how the phase shifter output error is primarily a function of the amplitude and phase balance associated with the 90° hybrid (both plots have been normalized to the insertion phase of the QH-7-4.9).

The second group of plots result from using active all-pass networks to obtain the 90° phase split. The amplitude and phase balance for this circuit (one output normalized to the other) is presented in Figure 8. The amplitude balance is seen to be within 0.05 dB (much less than that of the passive hybrid), and the deviation from quadrature phase is approximately ±1° from 300 kHz to 7 MHz. Figure 9(a) is a plot of the phase shifter output for a requested shift of 45° (top trace), normalized to the insertion phase of the active 90° all-pass network. For convenience, the phase balance is presented on the same plot (bottom trace). Notice that the shape of the output phase is a scaled version of the bottom trace; the relatively tight amplitude balance of the quadrature splitter contributes a negligible amount to the total output phase error. Figure 9(b) is a plot of the output for a requested shift of 90° (top trace) and the phase balance of the active splitter (bottom trace). Note the almost one to one correspondence between the shape factors of both traces.

Figure 10(a) shows the phase shifter output normalized to the insertion phase of the active splitter. The bottom trace in Figure 10(b) is an expanded view (2°/div) for a requested shift of 0°. This plot shows the contribution that the main body makes to the total phase non-linearity of the vector modulator (a delay has been added mathematically to "flatten out" the linear phase portion of the response for illustrative purposes). Again, the rise at the low end of the response results from being close to the low frequency roll-off of the amplifier used (100 kHz). The plot in Figure 10(b) shows the insertion phase of the entire vector modulator, including the active splitter (top trace).

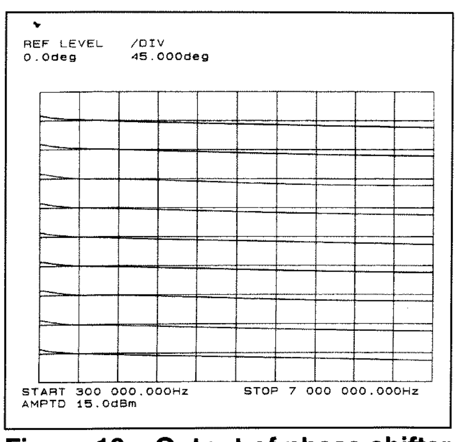


**Figure 10a. Output of phase shifter normalized to the insertion phase of the active splitter.**

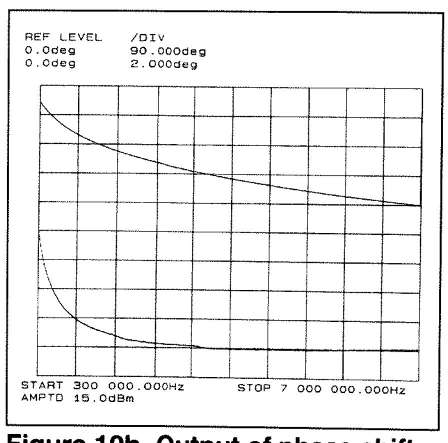


**Figure 10b. Output of phase shifter normalized to the insertion phase of the entire vector modulator.**

## Conclusion

A design for a high accuracy broadband phase shifter based on the concept of the vector modulator has been presented. The design is flexible enough to allow either digital or analog phase shift control, allowing the device to be used as a high linearity phase modulator. Using currently available analog processing components, the design achieves its goal of broadband, linear phase control for shifts well in excess of 360° (digital control input option). The device is also capable of being controlled via analog input voltage (or modulating waveform), resulting in phase excursions of up to +/- 500°. The accuracy of the phase shifter has been shown to rely heavily on the amplitude and phase imbalances associated with the 90° phase splitting device used. In addition, the group delay of the quadrature phase splitter directly affects the delay characteristics of the phase shifter. For linear-phase sensitive applications, the use of either DDS techniques (successfully implemented, but not shown here) or linear phase all-pass structures may be employed to provide the required I and Q inputs to the vector modulator. For applications not requiring flat group delay, many "off the shelf" quadrature power dividers are available to serve the same purpose.

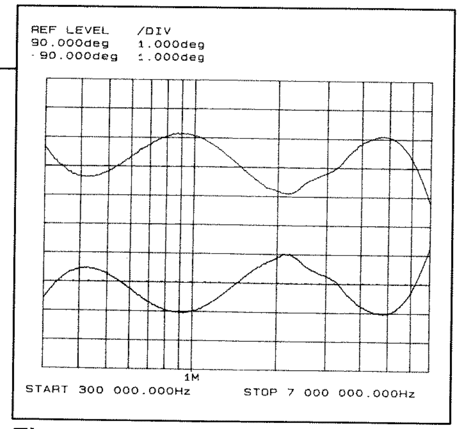


**Figure 9b. Output of phase shifter at 90 degrees and phase balance of active splitter.**

## Acknowledgements

The author is indebted to the engineering team of the AGS RF group for the many fruitful discussions and development ideas leading up to the current design. In addition, the author wishes to express his gratitude to T. Hayes for his review of the manuscript, and to J. Cupolo for his RF prototyping talents. This work was performed under the auspices of the U.S. Department of Energy. ***RF***

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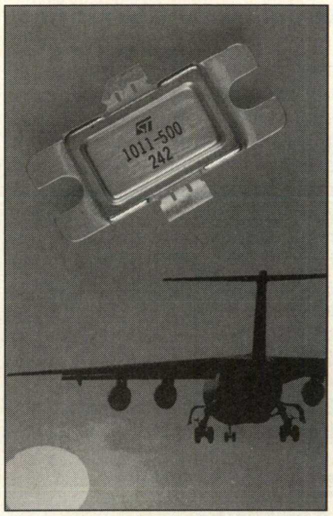
3. Hostetter, G.H., *Fundamentals of Network Analysis*, New York, Harper & Row, Inc., 1980, pp. 337-344.

### About the Author

Dominic J. Ciardullo is a Research Engineer in the Low Level RF group at Brookhaven National Laboratory. He received an Associates Degree in Engineering Science from Nassau Community College, BSEE from Rensselaer Polytechnic Institute and MSEE from the Polytechnic University. He is on the adjunct staff in the Eng./Phy./Tech. department at Nassau Community College. He can be reached at Brookhaven National Laboratory, Building 911B, Upton, NY 11973.

## 500 W Transistor

SGS-Thomson Microelectronics has introduced a high pulsed power, class-C transistor, specifically designed for L-band avionics, TCAS (Traffic and Collision Avoidance Systems) and Mode-S transponder output driver applications. The AM1011-500 offers a minimum output power of 500 W peak. The device has minmum collector effiency of 40% and minimum power gain of 8.5 dB. The AM1011-500 is supplied in the 0.4×0.6 inch SIXPAC™ hermetic metal/ceramic package with integral input/output matching structures.
**SGS-Thomson Microelectronics**
**INFO/CARD #220**




## SAW IF Filters

RF Monolithics will introduce four additions to its line of SAW IF filters for digital radiotelephone applications at RF Expo East. The new filters are the PX1002, PX1003, PX1004 and PX1005. The PX1002 has a center frequency of 86.85 MHz, the PX1003 has a center frequency 150.005 MHz, the PX1004 has a center frequency of 82.20 MHz, and the PX1005 has a center frequency of 86.01 MHz. The PX1003 is designed for CT-2 and PCN IF applications, while the other filters are designed for AMPS, IS-54 (TDMA) and CDPD use.
**RF Monolithics, Inc.**
**INFO/CARD #219**

## Digital Down Converter

The HSP50016 digital down converter is a high speed (75 MHz), monolithic synthesizer, quadrature mixer and lowpass filter device. The device can operate as a single narrow band lowpass filter and receives CW, frequency hopped, or linear FM up or down chirp signals. The HSP50016 is designed to be compatible with common DSPs.
**Harris Semiconductor, Standard Products**
**INFO/CARD #218**

## Feed-Forward Amplifier

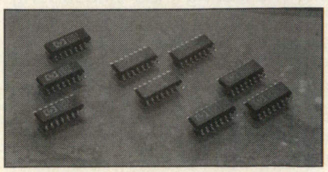
The QBS-101 is a feedforward, high dynamic range amplifier operating from 2 to 70 MHz. The amplifier has greater than 60 dBm 3rd order intercept point and greater than 100 dBm 2nd order intercept point. Gain is 12 dB with less than 4 dB noise. The amplifier operates from 24 VDC, 350 mA.
**Q-bit Corporation**
**INFO/CARD #217**

## RF Simulation

Compact Software introduces the release of SUPER-SPICE™ — an RF, microwave and high-speed frequency/time domain simulator that overcomes many of the limitations of pure time-domain simulators. The schematic-driven software provides for modeling of transmission media and discontinuities and contains and electromagnetics module that can be called from within the package.
**Compact Software, Inc.**
**INFO/CARD #216**

## Vector Modulators

The HPMX-2003, -2004, -2005 are the latest additions to Hewlett-Packard's HPMX family of vector modulators. The new modulators feature internal 90 degree phase shifters and high impedance I and Q inputs that allow high impedance sources to drive them. They operate from a 5V supply, and are housed in SO-16 surface mount packages. The I, Q, and LO ports are capable of being driven single-ended or differentially. The HPMX-2003 provides 900 MHz direct conversion, the HPMX-2004 provides dual conversion and an upconversion mixer up to 2 GHz, the HPMX-2005 is a < 50-250 MHz direct IF modulator.
**H-P Communications Components Div.**
**INFO/CARD #215**



## MRI Amplifier

Erbtec Engineering introduces its all solid-state modular MRI amplifier. The amplifier produces 5 kW pulsed or 500 W continuous output power. The amplifier includes built-in safety, monitoring and control features. The compact unit requires only 29 3/4 vertical inches of mounting space in standard equipment racks.
**Erbtec Engineering, Inc.**
**INFO/CARD #214**

## Spread Spectrum Transmitter

The Tektronix 900 MHz Spread Spectrum Transmitter provides a level of functional integration previously unavailable for use in radio communications. The device is capable of generating +20 dBm into low-impedance loads and is fully compliant with FCC Part 15 requirements for the 902-928 ISM band. Output power can be controlled over 30 dB. The part draws 120 mA at 5 V.
**Tektronix Microelectronics**
**INFO/CARD #213**

## TCXO

The TF69100 is an extremely small, 20 × 20 × 10 mm TCXO which can be designed to provide a frequency stability over temperature of ±3 ppm. It has a voltage-control option and excellent phase-noise. The device is designed for high volume production.
**Time & Frequency Ltd.**
**INFO/CARD #212**

## Wireless Design

The Communication Design Suite thoroughly integrates all the CAE analysis and design-for-manufacturing tools needed to create competitive PCB-based designs. It is the only CAE tool suite specifically developed for designing wireless applications such as analog or digital cellular radio, GPS, collision avoidance systems, wireless LANs and many other new products.
**EEsof Inc.**
**INFO/CARD #211**

## Low Profile TCXO

Piezo Technology has recently introduced the model X03022C, a low profile, high performance TCXO with a 0.20 inch height and a footprint of 1.00 × 1.25 inches. The unit is available over the frequency range of 16 to 75 MHz and offers temperature stability of ±0.5 ppm over −25 to +75 degrees C. Other features include 5V, 3mA operation with sinewave output and rugged construction.
**Piezo Technology, Inc.**
**INFO/CARD #210**

## Ceramic Filters

Integrated Microwave introduces a full line of two- and three-pole ceramic filters for applications requiring a low profile and small footprint. These units are available in two and four percent bandwidths with center frequencies from 600 to 2700 MHz. The parts exhibit good temperature stability and excellent Q. The units are available in surface mount and through-hole configurations.
**Integrated Microwave**
**INFO/CARD #209**

## PLL Evaluation Board

Motorola has released an evaluation kit for the MC14590 and MC14591 1.1 GHz PLL frequency synthesizers. The turnkey kit contains an assembled PC board with PLL and VCO, software for a PC, and printer port cable. Up to three boards may be hung on the cable, which facilitates multi-loop evaluations. Price is $200.
**Motorola, Inc.**
**INFO/CARD #208**

## Switching Sub-Assemblies

JFW Industries announces their product line containing several types of switching sub-assemblies. These sub-assemblies are custom designed to the customer's requirements and can include combinations of RF switches, power dividers, programmable attenuators, RF transformers, etc. JFW has designed and manufactured products that are PC, GPIB, TTL, RS-232 and HPIB controlled.
**JFW Industries, Inc.**
**INFO/CARD #207**

## Synthesizer Line

Giga-tronics has announced the completion of its purchase of a series of RF signal generators

from John Fluke Manufacturing . The 6060 Series uses indirect synthesis to provide the spectral purity, accuracy and stability needed to test communications and navigation equipment to 2.1 GHz. The 6060B and 6061A cover 10 kHz to 1 GHz, while the 6062A covers 100 kHz to 2.1 GHz. U.S. list prices start at $6250.
**Giga-tronics Incorporated**
**INFO/CARD #206**

## Precision Crystals

A 10.000 MHz, 3rd overtone crystal from Frequency Products has Allen Variance better than $1x10^{-11}$ per second. The crystals have Q of 600,000 and aging is less than 0.5 ppm/year. Enclosed in HC-47 housings, the crystals are available 4 weeks ARO, including a 2 week powered burn-in. Frequencies other than 10.000 MHz require 6 weeks.
**Frequency Products, Inc.**
**INFO/CARD #205**

## Susceptibility Probes and Monitor

Probes in the FM2000 E-field monitoring system from Amplifier Research can be connected via 1000 meters of fiber optic cable to a remote monitor. The probes have frequency responses of 10 kHz to 1 GHz and 80 MHz to 40 GHz and are powered by rechargeable batteries. The FM2000 accommodates up to four probes and provides output via IEEE-488, RS-232 and 0-4 VDC outputs.
**Amplifier Research**
**INFO/CARD #204**

## VCXO

The VC-7000 Series VCXOs from Raltron Electronics is designed for use in high-speed board- and box-level modems, instrumentation, imaging, high-end audio and other applications requiring high stability and fast, wide-range phase locked loops. The series is available in frequencies up to 150 MHz and with control voltage sensitivities of ±100ppm/volt. Output drive is specified at 10 TTL loads or 15 pF for HCMOS loads. The VC-700 series is available in custom and semi-custom versions. Pricing is $15.00 each.
**Raltron Electronics Corp.**
**INFO/CARD #203**

## RF Parameter Extraction

The HP 85123A RF parameter extraction test system is the first fully-integrated RF and DC measurement system designed for modeling active devices often used in RF communications applications. When combined with the HP 85190 family of high-frequency IC-CAP software, the total solution provides the capabilities to model BJTs, MOSFETs and MESFETs.
**Hewlett-Packard Co.**
**INFO/CARD #202**

## 900 MHz RF IC Set

Motorola has introduced a 900 MHz RF integrated chip set for personal communications systems. Although designed as a front end for the CT-2 cordless telephone system, these devices are ideal in other 900 MHz systems, such as GSM, ISM and 915 MHz cordless telephones. The MRFIC2001 through MRFIC2006 device series consists of an upconverter, a down-converter, a low-power amplifier, an integrated driver/ramp/inverter chip and an antenna switch. Low volume pricing for these devices range from $2.33 to $3.79.
**Motorola CSPD**
**INFO/CARD #201**

## Signal Processing Components

M/A-COM recently introduced a new family of ISO 9001 qualified RF signal processing components for commercial applications to 4200 MHz. This family, called E-Series, is detailed in a new 120 page catalog offering detailed specifications, outlines and application notes on mixers, I/Q modulators, power splitters/combiners, couplers and transformers. Many surface mount styles are available in tape and reel packaging.
**M/A-COM**
**INFO/CARD #200**

## VXIbus Multiplexer

The Model 1260-51 from Racal-Dana Instruments is a VXIbus multiplexer with 400 MHz bandwidth. Model 1260-51A has eight 1×4 multiplexer sections and is reconfigurable to a single 1x39 multiplexer. Model 1260-51B has 16 1×4 multiplexer sections and can be reconfigured to a single 1×79 multiplxer. Both multiplexers come on a c-sized, message-based VXIbus card.
**Racal-Dana Instruments, Inc.**
**INFO/CARD #199**

## Peak Power Sensor

The Boonton model 56418 peak power sensor is now available for low level peak power applications in GSM, CDMA, and TDMA digital communications schemes, as well as avionics and radar signals. Fully compatible with the Boonton 4400 power meter, the new sensor operates from 500 MHz to 18 GHz and has a peak power range of –34 dBm to +5 dBm and CW power range of –40 dBm to +5 dBm. Price is $1425, with delivery in 2-4 weeks ARO.
**Boonton Electronics Corp.**
**INFO/CARD #198**

## Linear Power Amplifier

ENI's model 525LA power amplifier produces 25 watts of linear class A output over a 1 to 500 MHz frequency range. With a gain of 50 dB, the 525LA features low harmonic and intermodulation distortion; all harmonics are more than 23 dB below the main signal at full power. The amplifier has


unconditional RF stability, +13 dBm overdrive protection, and infinite maximum load VSWR. The 525LA is available for 30 day delivery at a cost of $5625.
**ENI**
**INFO/CARD #197**

## Dual Tracking Trimmer

Voltronics has expanded its line of dual tracking precision trimmer capacitors to work in the GHz range. The V6100, non-rotating piston sapphire-dielectric trimmer, tunes from 0.5 to 3.5 pF, is only 0.48 inches long, and can be used to 2 GHz. The V4025 is a PTFE dielectric, half-turn trimmer measuring 0.33 inches long by 0.22 inches wide by 0.09 inches high. It tunes between 0.2 and 2.5 pF and can be used up to 6 GHz. Prices are from $15 to $8.47 in 1000 piece quantities.
**Voltronics Corp.**
**INFO/CARD #196**

## Dual Directional Coupler

Merrimac Industries has expanded their range of dual directional couplers with the model CGN-30-0.9G, a 500 W, dual directional coupler designed to monitor both forward and reflected power on a cellular radio base station feed. Typical insertion loss is 0.05 dB, and VSWR is better than 1.02:1 on either main line port.
**Merrimac Industries, Inc.**
**INFO/CARD #195**

## Broadband Delay Line

Sawtek announces the development of a wideband delay line with a bandwidth in excess of 600 MHz, center frequency of 1.3 GHz and 2.8 µs of delay. These wideband delay lines are ideally suited for signal delay in EW systems including channelized receivers that detect and identify signals simultaneously present throughout a large bandwidth. RF crosstalk is more than 55 dB below the desired out put and time-domain spurious signals such as triple transit signals are suppressed by more than 65 dB.
**Sawtek, Inc.**
**INFO/CARD #194**

## Dividers and Couplers

Merrimac Industries, Inc., will display a range of medium power four way power dividers for wideband antenna feed and amplifier applications in the frequency range 10 to 800 MHz. Merrimac will also show their new high power four port quadrature hybrids, with models to cover the 100 to 500 MHz frequency band.
**Merrimac Industries, Inc.**
**INFO/CARD #193**

## Breadboard Test Kits

Lorch Electronics introduces Waffleline® breadboard test kits to aid in prototyping and modifying system designs. The kits can be used to configure components housed in relay, TO and DIP style MIC packages. RF connectors are SMA female. DC connectors are EMI filter feedthroughs.
**Lorch Electronics**
**INFO/CARD #130**

# Subtle Changes for RF Brawn

*By Andy Kellett*
*Technical Editor*

Many of the new developments in RF technology are focused on the information content of RF signals. However, the power that carries that information is still essential to many of the new information-based systems, and RF power components are changing to accommodate them. In addition, there are applications where RF power is used for its own sake, and these applications demand their share of attention from manufacturers too.

## Applications

Probably the largest market for RF power components is the cellular base station market. Cellular base stations for analog systems are still being sold as cell sizes shrink to accommodate more users, and new digital cellular systems being introduced will provide even more sales. HDTV is still in developmental stages, but many manufacturers expect that market to become significant. RF susceptibility testing is climbing to higher frequencies as standards become concerned with higher and higher harmonics and high volume RF devices begin to use higher frequencies. .

According to Bruce Murray, Executive Vice-President of Erbtec Engineering, the MRI market is tough in the U.S., but new systems operating with higher magnetic fields, and upgrades to old systems will provide some business. While defense and aerospace have taken a hit in the last few years, they still provide a sizable portion of the business for high power manufacturers. High power RF has also found many narrow markets where RF power is used for its own sake. RF sources are used for metal sputtering in the semiconductor industry, and they find use in metal treating, plastic welding, adhesive curing, lamp exciters and particle accelerators.

## Devices

"We're in that good old continuous improvement mode," says Jerry Levine, Sales and Marketing Manager for MA/COM Power Hybrids. Transistor power capacity, efficiency, linearity and gain are all seeing steady improvement. According to Levine, MA/COM has been improving die especially to improve linearity and gain. "Our devices' high gain is perhaps their best selling point," notes Levine. For cellular applications, the emphasis is increasingly on linearity and gain says Dave Boylan, Product Manager for RF and Microwave Transistors at Philips Semiconductors. "The increasing market demand for higher frequency performance in the range of 1 to 2 GHz has led to optimization of diffusion methods, resulting in new power transistors fulfilling these linearity and gain expectations," noted Boylan.

However, for applications that inherently need high power, power capacity is still an important issue. "Certainly combiner advances help, but ultimately the answer lies in finding ways to get more power per discrete transistor and in combining several transistors into one package," says Carl Lump, Marketing Manager for RF Products at SGS Thomson. In this area, conflicting design goals make advances slow but steady. There is a drive for lower cost packaging, but packaging is important to heat dissipation. Emitter peripheries can be increased to increase power, but at the expense of gain-bandwidth.

"Tubes are an old technology, but still a high-tech technology," notes Seymour Paul, Manager of Industrial and Scientific Markets for Varian. The market for tubes is still strong in the highest powered applications and in the high power applications where transistor/combiner solutions are too costly. Some tube capabilities are completely unmatched by transistor's. "You can pulse a tube well beyond its normal power limit," notes Paul, a feature which makes tubes attractive for high powered, pulsed applications such as MRI and radar. However, tubes do not enjoy the inherent redundancy of combined transistor power modules. For example, if the 20 kW tube in a television transmitter fails, the signal is lost entirely.

For this reason, and because tubes contain so many parts, tube buyers are especially concerned with ruggedness and reliability. "We kept the Eimac brand name when we acquired it in the 1960's because it was so well known for its performance and quality worldwide," notes Varian's Paul. Reliability is also one of the selling points for the tubes sold by Svetlana, a Russian/American owned company that sells Russian-made tubes. "While our tubes have innovative features ahead of anything marketed in the West, there are no startling advances over Western tubes, but because the Russians depend on tubes more, they place great emphasis on tube reliability," says George Badger, Vice-President and Director of Marketing for Svetlana in the U.S.

Makers of passive devices for high power RF are faced with small, narrow markets, meaning they must be particularly responsive to customer's needs. "The high power market is not that large," says Glen Werlau, president of Werlatone Incorporated, "and the market for the broadband combiners and couplers we make is even smaller." "RF Power Components was started to fill a niche. For instance, our competitors were doing fine with 400 to 1000 MHz couplers that only went up to 400 W. We took it a little further and made one that goes up to 800 W," says Tom Passaro, Vice-President and owner of RF Power Components.

Manufacturers of high power RF amplifiers have a unique perspective on RF power because of their position as both buyer and seller of high power devices. Amplifier Research (AR) has found a large market for their products in the automotive susceptibility and defense-electronics susceptibility markets. Jim Maginn, Manager of Product Operations for AR, says their customers request larger bandwidths, higher power, and smaller cabinets, all while keeping price at bay. AR tries to satisfy this demand with amplifiers that require no band-switching. According to Maginn, these amplifier designs center around the transistor selected. "Transistor advances are made pretty continuously, but the developments seem agonizingly slow to us sometimes," says Maginn.

## Summary

Manufacturers of high power RF components and equipment steadily expand their product's capabilities. New, commercial/consumer markets make their presence known, but their effects are felt more subtly in the high power arena; no one expects manufacturers to produce transmitters on high volume production lines. Smaller volume, more specialized applications are still a good source of business for high power RF manufacturers.

# *RF software*

## Conversion Analysis Software

TxRx Designer from Waypoint Engineering provides rapid determination of the most commonly needed conversion-system cascaded element parameters, response plots, and mixer spurious products. An extremely friendly graphical user interface is employed with moveable windows and extensive help screens. Cost is $149.95.

**Waypoint Engineering**
**INFO/CARD #190**

## SIGINT Network Software

Watkins-Johnson has announced a networked control architecture for SIGINT, ELINT and COMINT receiver systems. W-J's NOVA (Networked Open Versatile Architecture) is designed around two computer standards: layered network protocols, in particular TCP/IP; and the Motif, X Window System environment.

**Watkins-Johnson Company**
**INFO/CARD #189**

## Amplifier Simulation

SW.I.F.T. Enterprises announces the latest release of ASP (Amplifier Simulation Program) Version 3.20 for interactive development of weak signal solid-state amplifiers. Various auto- and manual-routines allow the developer to optimize for noise, gain, and output VSWR, while providing input/output matching circuits. Design by Stern's stability factor, gain, and matched conditions are available to the engineer. Cost is $89.95 plus $3 shipping.

**SW.I.F.T. Enterprises**
**INFO/CARD #188**

## ASIC Design Files

Tanner Research and the MOSIS Service are offering updated ASIC technology setup files for use with Tanner's full custom IC layout editor, L-Edit Version 5. These latest technology files support the MOSIS processes for Hewlett-Packard, Orbit Semiconductor and VLSI Technology chip fabrication.

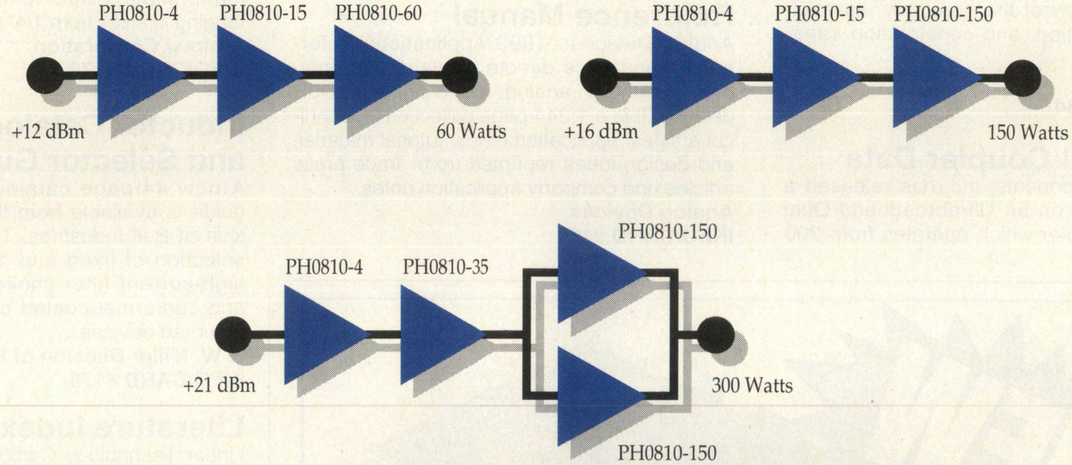
**Tanner Research, Inc.**
**INFO/CARD #187**

## System Simulator

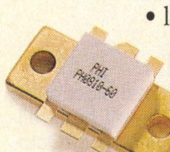
Running under Microsoft Windows, SystemView from Elanix is a high level conceptual design and analysis engine embedded in a visual design environment. SystemView supports multi-rate systems, parallel simultaneous systems, and internal or external data sources and sinks. The software also provides a new approach to analog-digital filter design, discrete time linear system design, and continuous-time Laplace linear system design. A single-user license costs $985.

**ELANIX, Inc.**
**INFO/CARD #186**

# Highest Gain 900 MHz Power Transistors for Base Stations




### Advantages of Higher Gain Transistors




- lower cost lineups
- more margin
- less tuning and assembly time
- lower cost per dB

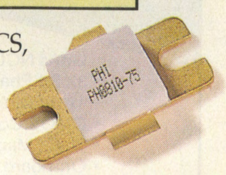
Improve your designs by using M/A-COM transistors.

Call or write for a 1993 catalog, or ask for our catalog on disk. Our transistors are also in stock at Richardson Electronics (800-323-1770) and Nu Horizons (516-226-6000).

| Part No. | P Out | Gain (Min) | $IMD_3$ (Max*) |
|---|---|---|---|
| PH0810-150 | 150W | 10 dB | -28 dBc |
| PH0810-75 | 75W | 10 dB | -28 dBc |
| PH0810-60 | 60W | 10 dB | -30 dBc |
| PH0810-35 | 35W | 10 dB | -30 dBc |
| PH0810-15 | 15W | 12 dB | -30 dBc |
| PH0810-4 | 4W | 14 dB | -30 dBc |

* Below Carrier

Similar lineups available for JDC, PCN, PCS, and INMARSAT frequencies.

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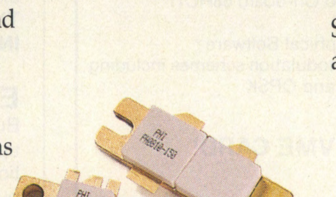


M/A-COM

INFO/CARD 94

**Please see us at RF Expo East '93, Booth #604, 606**

## Test Instruments
Farnell/Wayne Kerr has produced a new RF Test Instrumentation Guide. Included are 2.4GHz and 1GHz Signal Generators, the 1GHz Spectrum Analyzer and Automatic Modulation Meters. Also featured is the EASY 1 Emissions Assessment System, which is a complete Windows-based hardware and software system for EMC precompliance testing.
**Farnell/Wayne Kerr**
**INFO/CARD #185**

## Component Catalog
An updated version of Microflect's Component Catalog, CC793 includes over 1,100 products designed to meet waveguide support and protection, tower accessory, tower hardware and antenna support requirements. Also available from Microflect, is a 24-page overview of the company, its engineering, fabrication, and construction capabilities.
**Microflect**
**INFO/CARD #184**

## Directional Coupler Data
RF Power Components, Inc. has released a new data sheet on an Ultrabroadband Dual Directional Coupler which operates from 200 MHz to 6 GHz. Model DDC-HP-201-602-1, measuring 5.50 × 1.60 × 1.05 inches, is the newest edition to RF Power's line of resistors, termination, 90° hybrid couplers and dual directional couplers.
**RF Power Components, Inc.**
**INFO/CARD #183**

## Spectrum Analyzer Brochure
Anritsu Wiltron Sales Company has released a 16-page, color brochure on its MS2610/ MS2620 series of spectrum analyzers that details all the capabilities of the five spectrum analyzers in the series. The brochure covers many specifications and applications of the series.
**Anritsu Wiltron**
**INFO/CARD #182**

## Reference Manual
Analog Devices' "1993 Applications Reference Manual," is directed towards the engineer working on analog, mixed signal, or DSP designs. It is a 1,344 page collection of technical articles, application notes, tutorial material, and design ideas reprinted from trade press articles and company application notes.
**Analog Devices**
**INFO/CARD #181**

## Broadband Microwave SDLVA Brochure
A 4 page brochure released by Veritech Microwave, Inc. describes the topology, operation and advantages of their Successive Detection Log Video Amplifiers. Full specifications, including graphical presentations of key parameters, are provided for three SDLVA's operating over the .5-2GHz, 2-8GHz and 2-18GHz frequency ranges.
**Veritech Microwave, Inc.**
**INFO/CARD #180**

## Coaxial Cables
Andrew Corporation is offering a 16-page guide to the selection of HELIAX® coaxial cables for cellular, land mobile, paging, microwave, broadcast and military applications. Included are three types of coaxial cables, LDF foam, superflexible foam and air dielectric, ranging in sizes from 1/4" to 5" in diameter.
**Andrew Corporation**
**INFO/CARD #179**

## Inductor Catalog and Selector Guide
A new 44-page catalog and coil selector guide is available from the J. W. Miller Division of Bell Industries. The catalog covers a selection of fixed and adjustable RF coils, high-current filter chokes, toroids, molded and conformal coated chokes, and surface mounted devices.
**J. W. Miller Division of Bell Industries**
**INFO/CARD #178**

## Literature Index
Linear Technology Corporation introduced an index to the Company's applications support literature entitled "Literature Subject Index." The index lists many topics covered in the company's application notes, design ideas and design aids. Topics included are accelerometers, amplifiers and analog-to-digital conversion.
**Linear Technology Corporation**
**INFO/CARD #177**

## Electronic Journal
A biweekly journal of international electronics reasearch, Electronics Letters, will be available as an online journal beginning in October 1993. The journal will be published by the Institution of Electrical Engineers (IEE) and distributed to subscribers by Online Computer Library Centre (OCLC) via Internet and dial-up telecommunications networks.
**The Institution of Electrical Engineers**
**INFO/CARD #176**

## Electronics Guide on disk
Burr-Brown announces the availability of its updated High Performance Electronics Selection Guide disk for IBM PC-compatible computers. The disk contains 1500+ current component models, industry cross-reference section, sales office listings, technical literature, domestic prices, and ordering information.
**Burr-Brown**
**INFO/CARD #175**

# RF design

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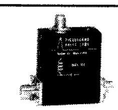
INFO/CARD 103

## BIAS TEES 10 kH - 44 GHz

| MODEL | RISETIME* | BANDWIDTH** | VOLT/AMPS | |
|---|---|---|---|---|
| 5530 | 28 ps | 12 GHz | 200 V | 10 mA |
| 5531 | 45 ps | 8 GHz | 1.5kV | 20 mA |
| 5540 | 8 ps | 44 GHz | 50 V | 100 mA |
| 5550 | 20 ps | 18 GHz | 50 V | 500 mA |
| 5580 | 32 ps | 11 GHz | 50V | 1 Amp |

Other broadband components available from PSPL include: Signal Probes, Risetime Filters, Attenuators, DC Blocks, Transformers, Power Dividers and Amplifiers.



*Risetimes measured with HEWLETT-PACKARD 50 GHz oscilloscope and PSPL 15 ps pulse generator.

**Bandwidths measured with WILTRON 20 GHz network analyzer.

PICOSECOND PULSE LABS Inc

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INFO/CARD 104

# *RF engineering opportunities*

# Advertiser Index

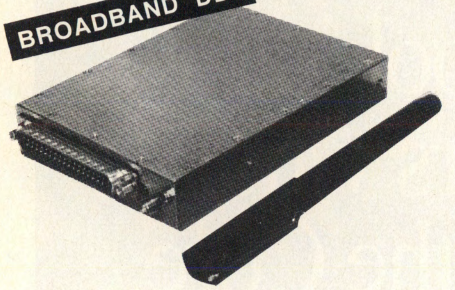


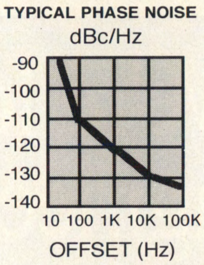
## *RF engineering opportunities*

# FREQUENCY SYNTHESIS

**Bandwidth** ................ **1 to 400 MHz**
Step size ............................................ <1 Hz
Frequency control .................. 30-bits, parallel
Spurs ....................................... <-45 dBc typ
Size ............................ 3.75" x 4.78" x 0.89"
**Output** ................................ **Quadrature**

BROADBAND DDS

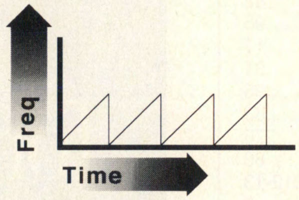


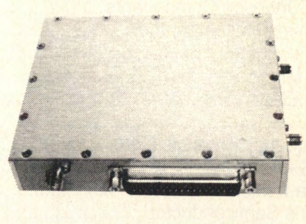



## ADS-431
ADVANCED DDS

**Bandwidth** ................. **1 to 230 MHz**
Step size ......................... <30 Hz granularity
Phase control ..................................... 12-bit
Switching .................... 2 nanosecond update
Spurs ........................... -55 dBc typical (CW)
5.25" chassis or 5" x 7" x 1.125" module

LINEAR FM



## DCP-1
GaAs DDS

This **LOW COST** module is perfect for ATE, simulation, EW, and all other applications where speed and phase continuity are important.

For Synthetic Aperture Radar and other systems requiring state of the art LINEAR FM. Sciteq's DCP-1 provides extended bandwidth, high linearity, phase control, and excellent spectral purity.

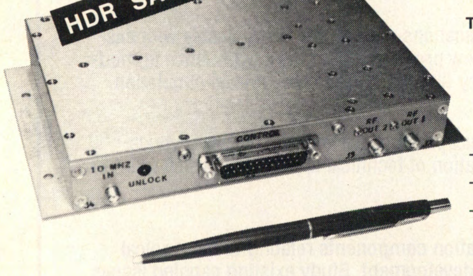
## SCITEQ
specializes in advanced technology
**frequency synthesis**,
using direct digital, phase-locked-loop, Arithmetically Locked Loop, and mix/filter designs, plus unique combination architectures that combine the advantages of multiple underlying technologies.

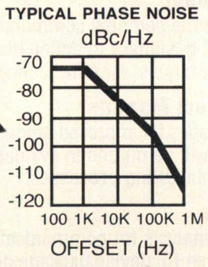
The VDS-6030 is a low cost C-band solution, with excellent phase noise and low power dissipation. It's also available in L-band.

This one square inch waveform generator includes digital phase, frequency, and amplitude control, yet it only dissipates 1.5W at $f_{ck}$ max.

## VDS-6030
ARITHMETICALLY LOCKED LOOP

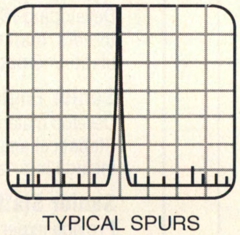
HDR SATCOM




**Bandwidth** ........... **4.65 to 5.25 GHz**
Step size ....................................... 2.5 MHz
Outputs ..................... 2-channels @ +17 dBm
**Total power** ....................................... **<5 W**
BIT .................................... TTL lock indicator
Spurs ............................................... -60 dBc
Reference .......... 10 MHz, ext (internal optional)

## DDS-1
SYNTHESIZER & MODULATOR


TYPICAL SPURS

ONE-CHIP DDS


**Maximum clock** ......... **up to 25 MHz**
Frequency control ................................ 32-bit
Switching speed .............................. <1 µsec
Output ....................................... SINE, 1 VP-P
Phase noise ............................... per the clock
Gating ............ >>100 dB ON/OFF, 1-bit toggle
**Modulation** ...... **digital amplitude, $\phi$, FSK**

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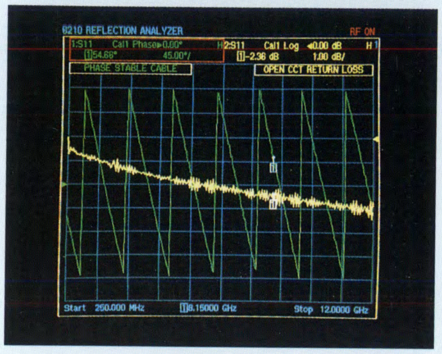
INFO/CARD 96

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

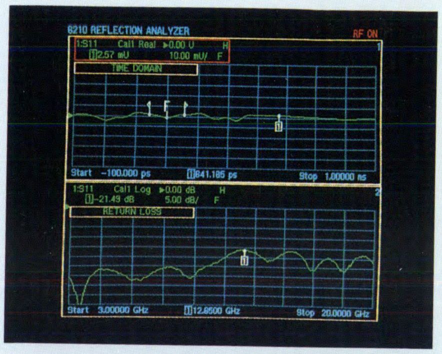



measurements, and time domain measurements. The frequency range of the Model 6210 starts at 250 MHz and extends to 26.5 GHz (or as limited by the host 6200 Series Microwave Test Set).

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:

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Allendale, NJ 07401
1-800-233-2955
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# Microwave Reflection Analysis Made Easy.


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