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engineering principles and practices

March 1996

Linear Power RFIC Covers 1 MHz to 1 GHz

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June 1995

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INFO/CARD 1

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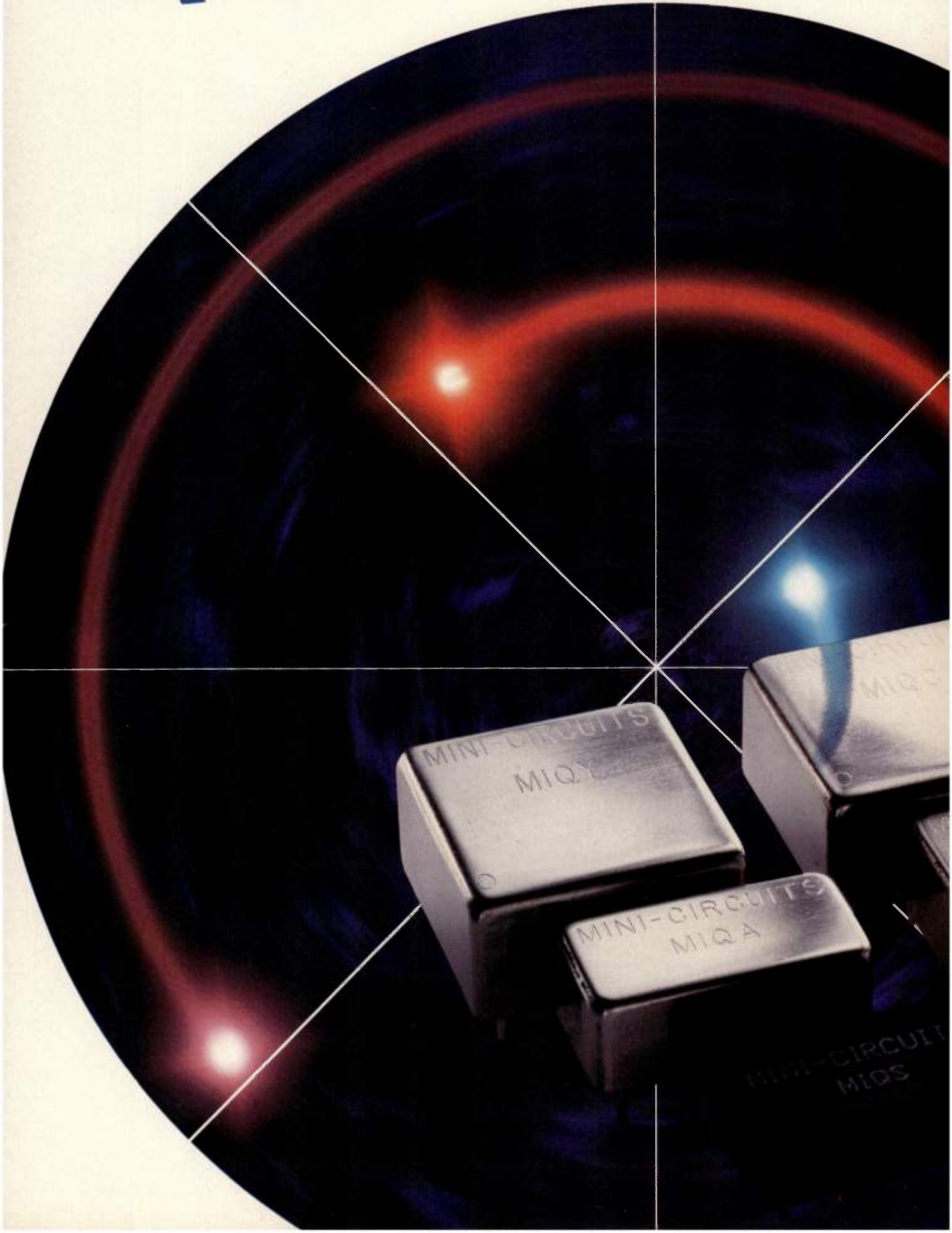
INFO/CARD 2

I&Q MODULATORS



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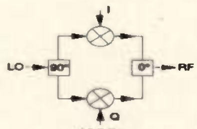
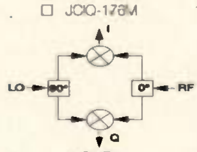


I/Q MODULATORS

MODEL	FREQ. (MHz) f_L	f_U	CONV. LOSS (dB) $\bar{x}$	σ	CARRIER REJ. (-dBc) Typ	SIDEBAND REJ (-dBc) Typ	HARM SUPPRESS (-dBc) Typ. 3xI/Q	5xI/Q	PRICE $ Qty. (1-9)
MIQA-10M	9	11	5.8	0.20	41	40	58	68	49.95
MIQA-21M	20	23	6.2	0.14	50	40	48	65	39.95
MIQA-70M	66	73	6.2	0.10	38	38	48	58	39.95
MIQA-70ML	66	73	5.7	0.10	38	38	48	58	49.95
MIQA-91M	86	95	5.5	0.10	38	38	48	58	49.95
MIQA-100M	95	105	5.5	0.10	38	38	48	58	49.95
MIQA-108M	103	113	5.5	0.10	38	38	48	58	49.95
MIQA-195M	185	205	5.6	0.10	38	38	48	58	49.95
MIQC-38M	34	38	5.5	0.10	48	37	54	65	49.95
MIQC-88M	52	88	5.7	0.10	41	34	52	66	49.95
MIQC-176M	104	176	5.5	0.10	38	36	47	70	54.95
MIQC-895M	868	895	8.0	0.10	40	40	52	58	99.95
MIQC-1785M	1710	1785	9.0	0.30	35	35	40	65	99.95
MIQC-1880M	1805	1880	9.0	0.30	35	35	40	65	99.95
□ MIQY-70M	67	73	5.8	0.20	40	36	47	60	19.95
□ MIQY-140M	137	143	5.8	0.20	34	36	45	60	19.95
Surface Mount Models									
□ JCIQ-88M	52	88	5.6	0.1	40	35	45	65	49.95
□ JCIQ-176M	104	176	5.6	0.1	35	35	45	65	54.95



I/Q DEMODULATORS

MODEL	FREQ. (MHz) f_L	f_U	CONV. LOSS (dB) $\bar{x}$	σ	AMP. UNBAL. (dB) Typ.	PHASE UNBAL. (Deg.) Typ.	HARM. SUPPRESS (-dBc) Typ. 3xI/Q	5xI/Q	PRICE $ Qty. (1-9)
MIQA-10D	9	11	6.0	0.10	0.15	1.0	50	65	49.95
MIQA-21D	20	23	6.1	0.15	0.15	0.7	64	67	49.95
MIQA-70D	66	73	6.2	0.10	0.15	0.7	56	58	49.95
MIQC-38D	34	38	5.5	0.10	0.10	0.5	60	65	49.95
MIQC-60WD	20	60	5.3	0.10	0.15	1.0	55	67	79.95
MIQC-895D	868	895	8.0	0.20	0.15	1.5	40	55	99.95
□ MIQY-1.25D	1.15	1.35	5.0	0.10	0.15	1.0	59	67	29.95
□ MIQY-70D	67	73	5.5	0.25	0.10	0.5	52	66	19.95
□ MIQY-140D	137	143	5.5	0.25	0.10	0.5	47	70	19.95
Surface Mount Models									
□ JCIQ-176D	104	176	5.5	0.1	0.15	2	52	65	54.95
□ JCIQ-895D	868	895	8.6	0.1	0.2	1	45	65	99.95
□ JCIQ-1785D	1710	1785	8	0.2	0.2	2	50	65	99.95
□ JCIQ-1880D	1805	1880	8	0.2	0.2	2	50	65	99.95

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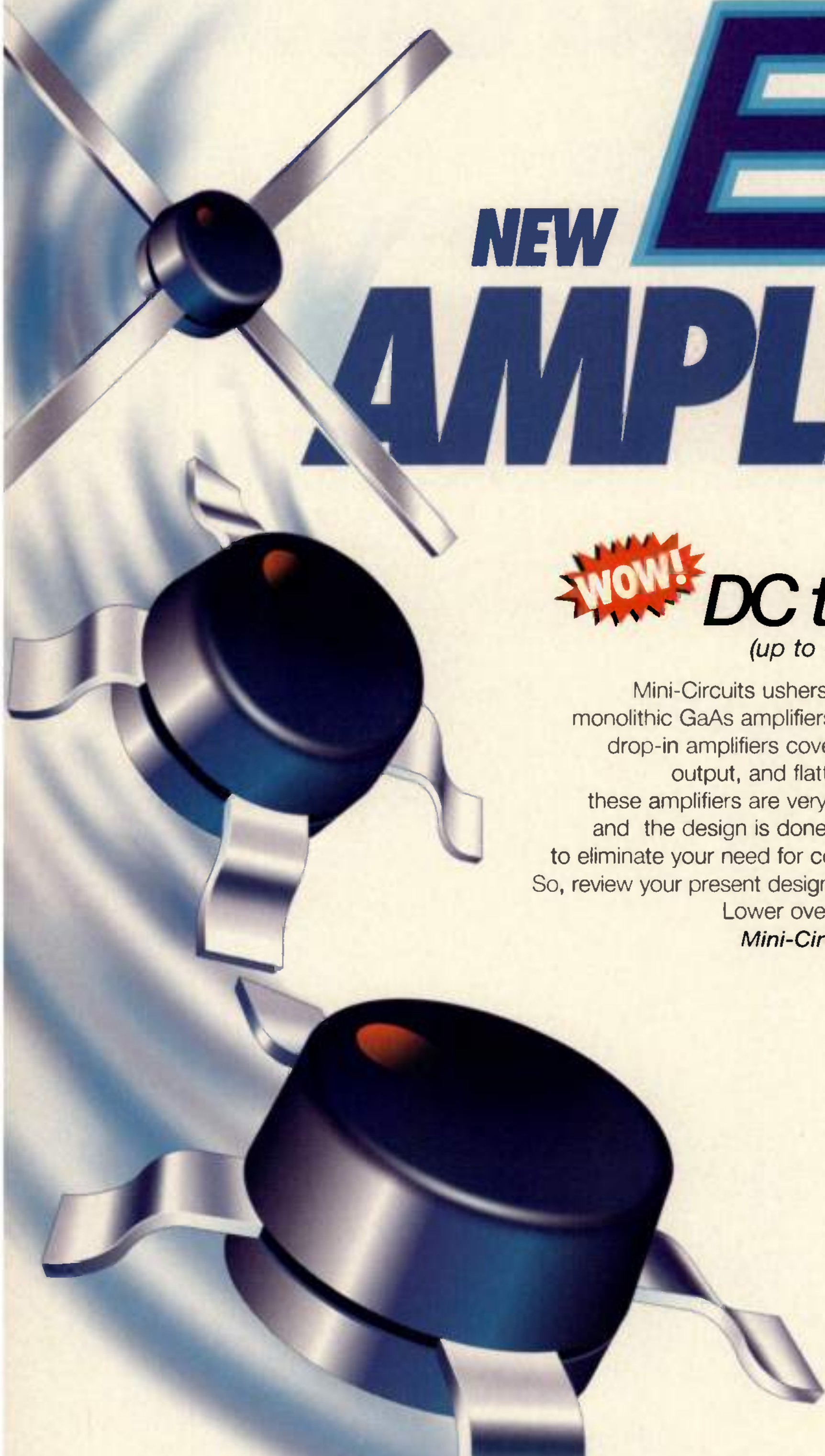

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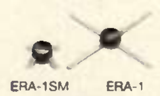
	ERA-1 / ERA-1SM		ERA-2 / ERA-2SM		ERA-3 / ERA-3SM	
*Frequency (MHz) f_L-f_U	DC-8000 / DC-8000		DC-6000 / DC-6000		DC-3000 / DC-3000	
Gain, (dB)	11.6 / 11.0		14.9 / 13.1		20.2 / 19.4	
Max. Power Out (dBm, @1dB comp.)	13 / 13		14 / 13		11 / 11	
Dynamic Range	NF	IP3	NF	IP3	NF	IP3
	7dB	26dBm	6dB	27dBm	4.5dB	23dBm
	7dB	26dBm	6dB	27dBm	4.5dB	23dBm
(1)Price ($ ea., Qty.10)	1.80 / 1.85		1.95 / 2.00		2.10 / 2.15	

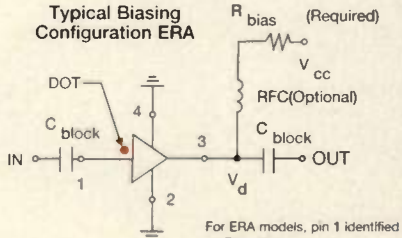
Note: All specifications typical at 2GHz, 25°C.
*Low frequency cutoff determined by external coupling capacitors.
(1)Price (ea.) Qty. 1000: ERA-1 $1.16, ERA-2 $1.31, ERA-3 $1.46.
Add $.05 for SM option.

Designer's Amplifier Kits
K1-ERA: 10 of each ERA model (30 pieces) only $49.95



ACTUAL SIZE

Typical Biasing Configuration ERA



For ERA models, pin 1 identified by Red dot.

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RF design

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- Active Filter Tutorial
- Product Focus: Crystal Oscillators
- More Engineer's Notebook Ideas!

RF editorial

A Preview of The Next RF Design Seminars

By Gary A. Breed
Editor

April 23-26, 1996 should be marked on your calendars! Not only is *RF Design* sponsoring our next seminar event at that time, but we and our sister publications in the cellular, PCS and mobile radio industries are hosting the giant International Wireless Communications Expo (IWCE).

Last year, some 7000 people came to IWCE, and that was before *RF Design* added its influence! Here's what you can find in Las Vegas at our big wireless show:

- Our highly acclaimed short courses on Wireless & RF Engineering, Filter Design, Oscillator Design, Power Amplifier Design, and Digital Modulation/Spread Spectrum
- Hundreds of exhibiting companies, including many RF component, instrument and software providers, plus companies providing equipment to the mobile radio industries — base stations, handsets, antennas, towers, test equipment, and more.
- Business and technical presentations on all aspects of wireless communications, including panel discussions of all the latest technologies — it's a great place to find out what's hot in communications!

RF Design's contribution to the informational meetings is a session on RF engineering, to be held on Thursday afternoon, April 25. Our speakers are busily preparing their papers on these topics:

"An Overview of Competing and Supplementary Wireless Services" by Roohollah Hajbandeh of DSC Communications Corp.

"High Speed Wireless Telemetry" by Roger Bracht, Larry Birkbigler, Ted Crawford and Paul Lewis of Los Alamos National Laboratory.

"Design & Development of a New Polyester Laminate for Wireless Communications Markets," by Chuck Ludwig of Glasteel Industrial Laminates.

"Probabilistic Models Allow More Accurate Estimates of Time to do Reliability or Qualification Testing of Electronics," by John Hahn of the Naval Air Warfare Center.

These papers on RF design and manufacturing topics are an excellent addition to the many other sessions on such topics as Digital Cellular, PCS, LEO satellite systems. Other IWCE sessions cover all key technologies, regulations and markets for wireless communications. Panels include industry leaders from government, associations, and companies.

While at IWCE, the RF Technology Pavilion on the trade show floor will be filled with the RF companies you need to see. From capacitors, inductors and transistors, to frequency counters, spectrum analyzers, and the latest analysis software. More than 70 exhibit booths are devoted to the RF engineer's needs.

So, join us in Las Vegas! For more information, look over the advertisement on pages 81-84, and the brochure included after page 68, then call and reserve your spot at IWCE. This is the only event where components through finished products are all presented in one place.

See you there!

Kalmus: A Tradition of High Performance, Value-Packed RF Amplifiers

"When Kalmus began designing and manufacturing RF amplifiers in 1971, the vision was to produce the best high performance product possible for the best possible price."

"As we address the needs of the engineering community in the 21st century, we never lose sight of this tradition. At the same time, however, we are taking our operations to an unprecedented level of engineering and manufacturing excellence."

"For example, our engineering and design capability enables us to respond rapidly to customer requirements. And with improved order handling and production capability, we are continually reducing delivery times."

"With the resources of Thermo Voltek behind us, Kalmus is committed to producing technically advanced RF amplifiers upon which customers can depend, and to expanding its leadership position."

Colin I. W. Baxter, President
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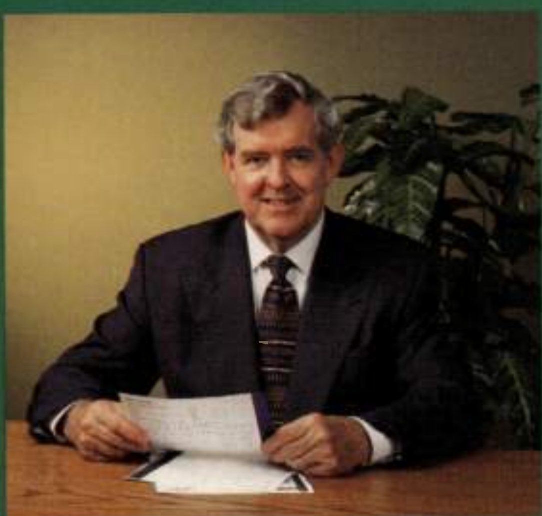

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Kalmus offers over 120 ruggedly built RF amplifiers for use in EMC testing, medical, communications and general laboratory applications. A comprehensive line of RF amplifier modules and custom amplifier options are also available.

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- Covering the frequency spectrum from 10KHz to 1GHz

Communications Amplifiers
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- Covering communications bands to 1GHz

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INFO/CARD 6

RF design

Established 197

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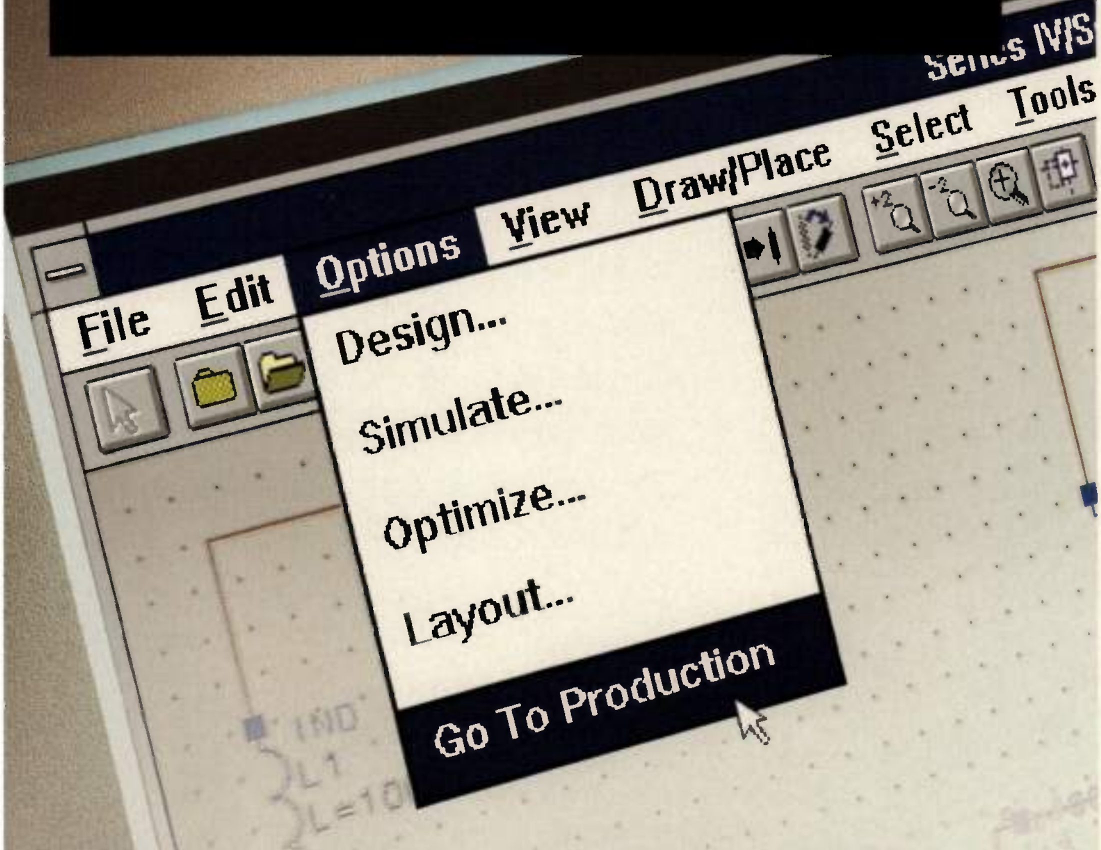

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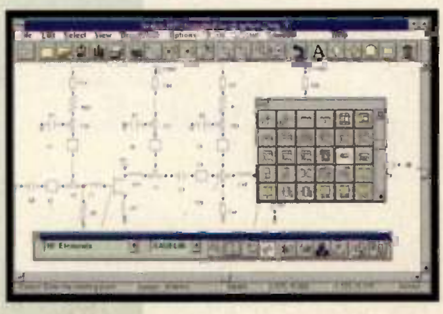
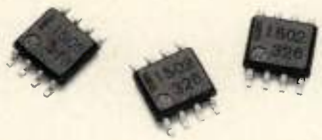


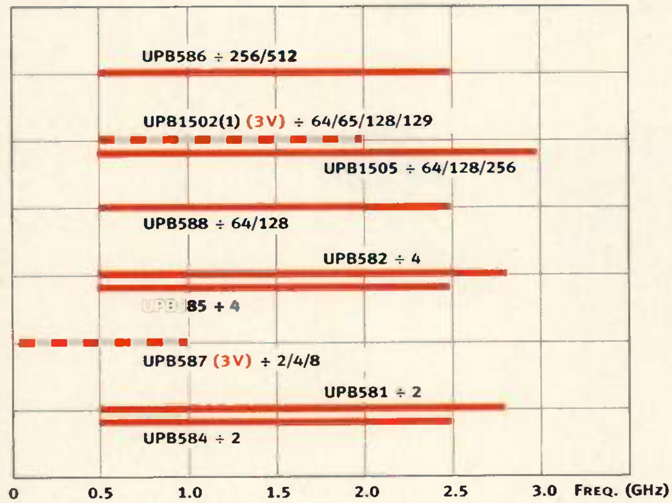
INFO/CARD 8

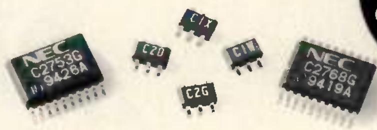
Booth 610 Wireless Show

CTIA Show Booth 1514



3V and 5V Prescalers—from $1.30





3V Frequency Converters—from 99¢

Part	RF Frequency (MHz)	I_{CC} (mA)	Conversion Gain (dB)	Output IP_3 (dBm)
UPC2756T[1]	100 - 2000	5.9	14	0
UPC2757T[1]	100 - 2000	5.6	13	0
UPC2758T[1]	100 - 2000	11	17	+6
UPC2753GR[1]	DC - 400	6.9	79	-17
UPC2768GR[1]	10 - 450	7	80	-17
UPC8106T[2]	100 - 2000	9	9	+1
UPC8109T[2]	100 - 2000	5	4	-4

1. Downconverter
2. Upconverter

NEC's UPC8104GR Quadrature Modulator with up converter

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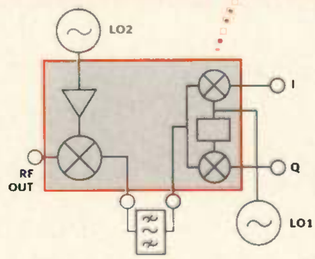

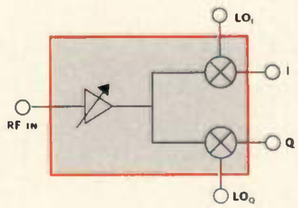

UPC8104GR

Quadrature Modulator w/Upconverter

- 3–5 Volt operation
- Ports for IF filter
- 10 MHz I/Q BW
- 400 MHz IF Bandwidth
- 1.9 GHz RF Output

UPC2766GR

Quadrature Demodulator

- RF & LO —DC to 1 GHz
- IF (IQ)—DC to 100 MHz
- 35 dB typ AGC dynamic range
- 30 dBc typ distortion

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Our power amplifiers know how to survive on the street. In fact, ITT guarantees both stability and ruggedness. And the low cost will surprise even the toughest Wall Street banker.

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INFO/CARD 10

RF letters

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

Comments on IMD Article

Editor:

I read the article by Michael Leffel in your June 1995 issue, "Intermodulation Distortion in a Multi-Signal Environment." I would like to convey my congratulations to Mr. Leffel for his treatment of the subject.

However, I believe it would be interesting to clarify some aspects of the article. First, equations (18), (19), (20) and (22) have some typing mistakes, but the readers can correct them. There are two more important errors. Equation (23) should be:

$$4\left\{\frac{N+1}{2}+\frac{1}{2}\mathrm{mod}\left(\frac{N+1}{2}\right)-1\right\}+\mathrm{mod}\left\{\frac{N}{2}\right\}$$

and the summation term in the set of equations (33) to (37) should be:

$$4\left\{\frac{j}{2}+\frac{1}{2}\mathrm{mod}\left(\frac{j}{2}\right)-1\right\}+\mathrm{mod}\left\{\frac{j+1}{2}\right\}$$

Fortunately, data in Tables 1 and 2 are correct, as would be found from these corrected formulae. *[it appears that all errors are entirely typographical — ed.*].

Another point is to clarify a statement made on page 81. It should read, "This IM product will always exist, even if a single tone is fed into a *non-linear* [instead of *linear*] system." If a single tone, or even multiple tones, are fed into a truly linear system, they will never produce IM products, since K3 = 0 in eq. (9).

Josep Jordana
Spain

Grounding System Notes

Editor:

The article, "Electrolytic Grounding Solves Failures for Alyeska Pipeline, " (December 1995 issue), indicates that the National Electric Code (NEC) requires a ground resistance under 25 ohms. I've worked with the NEC for over 50 years and it does not require a specified ground resistance for a completed system. Article 250, H, "Grounding Electrode System" specifies acceptable grounding systems. For several of these systems the NEC does not require the grounding resistance to be measured at all. Article 250-84 for a single electrode ("...rod, pipe or plate"), states that, if the ground resistance is greater than 25 ohms for the single electrode, one additional electrode shall be installed. But it does not specify the final ground resistance.

The article also implies that the completed ground system is rather massive, and that a final ground resistance of less than 15 ohms was achieved. It would be interesting and educational to find out how the ground resistance was measured, as the measurement of large grounding systems is not trivial.

A number of years ago I analyzed several hundred measurements of single 8-foot ground rods installed in more than 30 states, at more or less random locations. At approximately 10 percent of the sites the grounding resistance was greater than 1000 ohms for a single rod.

Also, a number of years ago I was given the task of completing the installation of a grounding plate at an Okinawa communication site. When I arrived, a large hole had been dug and a large copper plate procured. The hole looked like a marble bathtub — the whole site was coral. In accordance with instructions, the plate was installed and the hole filled with an electrolytic solution. The initial resistance (using as reference a ground system associated with a submarine system) was more than 10 kohms. Several months later it was still that high.

In my opinion, a low ground resistance, if achievable at economical cost may be useful. However, the first line of protection should always be adequate bonding so that harmful potentials cannot occur. Always remember that you never saw an airplane with a ground rod!

C.B. Christianson
Sierra Vista, AZ

Filter Follow-up

Editor:

Several readers of my article, "Design Parameters for 1-Through-7-Pole Input-Matched Lowpass/Highpass Filters," (September 1995 *RF Design*) recognized that I had, to my embarrassment, misinterpreted a key reference (Matthaei) and apparently "reinvented the wheel." the values in my Table II are within a few percent of singly-loaded Butterworth values, which Mattaei recommended for branching filters, and are tabulated in other references. A branching filter is obtained by paralleling the unloaded ports of singly-loaded LP and HP filters. The values in my Table III are for doubly-loaded Butterworth filters and, predictably, yield poorly-matched branching filters.

The theoretical input match of several branching filters using singly-loaded Butterworth values (from Matthaei's Table 4.06-1) was found to be somewhat worse than produced by branching filters using the values from my Table II. For example, 7-pole branching filters using singly-loaded Butterworth and my Table II parameters yielded worst-case input return loss of 33 dB and 70 dB, respectively. This interesting result might warrant further study.

In addition, I omitted inductor L44 in shunt with R in Figure 2e. Also, the measured attenuation of the 3-pole prototype filter plotted in Figure 8 at f $= 4f_c$, dB $|S21|$, exceeds the theoretical n = 3 value of –36.12 dB. The excess attenuation is believed to be a result of distributed capacitance associated with the tightly-wound coils used for L1 and L3.

I thank the correspondents for their constructive criticisms and appreciate the opportunity to respond and make these corrections. *RF Design* is a fine magazine — I have saved every issue since the first!

Chase P. Hearn
Williamsburg, VA

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RF calendar

March

18-22 **12th Annual Review of Progress in Applied Computational Electromagnetics**
Monterey, CA
Information: Richard W. Adler, ECE Dept./Code ECAB, Naval Postgraduate School, 833 Dyer Road, Room 437, Monterey, CA 93943. Tel: (408) 649-1111. Fax: (408) 649-0300. E-mail: rwa@mcimail.com

April

16-17 **Minnesota EMC Event**
Bloomington, MN
Information: Judie Anderson, 3001 Habor Lane, Suite 150, Plymouth, MN 55447. Tel: (612) 559-0332; Fax: (612) 553-9326.

23-26 **International Wireless Communications Expo (including the RF Design Seminar Series)**
Las Vegas, NV
Information: Intertec Presentations, 6151 Powers Ferry Road, N.W., Atlanta, GA 30339. Tel.: (800) 828-0420 or (770) 618-0499. Fax: (770) 618-0441.

28-1 **IEEE Vehicular Technology Conference-VTC '96**
Atlanta, GA
Information: Wendy Rochelle, IEEE/VTC'96 Registrar, P.O. Box: 1331, Piscataway, NJ 08855-1331. Tel: (908) 562-3870; Fax: (908) 981-1769. E-mail VTC96@IEEE.com

June

5-7 **Frequency Control Symposium**
Honolulu, HI
Information: Michael Mirarchi, Synergistic Management, 3100 Route 138, Wall Township, NJ 07719. Tel: (908) 280-2024. Fax: (908) 681-9314.

16-21 **MTT-S International Microwave Symposium**
San Francisco, CA
Information: Derry Hornbuckle, Hewlett-Packard; Tel: (707) 577-3658; Fax: (707) 577-2036, or Jerry Fiedziusko, Space Systems/Loral Corp. Tel: (415) 852-6868. Fax: (415) 852-5068.

24-25 **Time and Frequency Seminars: Introduction-Level 1**
Boulder, CO

26-28 **Time and Frequency Seminars: Fundamentals-Level 2**
Boulder, CO
Information: Wendy Ortega Henderson, National Institute of Standards and Technology, 325 Broadway, Boulder, CO 80303-3328. Tel: (303) 497-3593; Fax: (303)497-6461. E-mail: ortegaw@boulder.nist.gov

July

21-26 **1996 IEEE AP-S International Symposium and URSI Radio Science Meeting**
Baltimore, MD
Information: Mr. Jon Moellers, Steering Committee Chair, 445 Hoes Lane, P.O. Box 1331, Piscataway, NJ 08855-1331. Tel: (410) 993-6774; Fax: (410) 993-7432.

RF courses

adar Cross Section Reduction
March 26-29, 1996, Atlanta, GA
rinciples of Electronic Counter-Countermeaures
April 9-11, 1996, Atlanta, GA
hased-Array Radar System Design
April 29-May 2, 1996, Atlanta, GA
nformation: Continuing Education, Georgia Institute of echnology, Atlanta, GA 30332-0385. Tel: (404) 894-2547.

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ntennas: Principles, Design and Measurements
May 14-17, 1996, St. Cloud (Orlando), FL
nformation: Kelly Brown, Northeast Consortium for Engi-eering Education, 1101 Massachusetts Ave., St. Cloud, FL 4769. Tel: (407) 892-6146. Fax: (407) 892-0406.

SP Without Tears
April 17-18, 1996, Fort Lauderdale, FL
May 1-3, 1996, Boston, MA
nformation: Z Domain Technologies, Inc., 325 Pine Isle ourt, Alpharetta, GA 30202. Tel: (770) 587-4812; Fax: 770) 518-8368.

F and Wireless Engineering
April 23-25, 1996, Las Vegas, NV
ractical High Frequency Filter Design
April 23, 1996, Las Vegas, NV
scillator Design Principles
April 24, 1996, 1996, Las Vegas, NV
igital Modulation and Spread Spectrum for Wireless ommunications
April 23, 1996,
F Power Transistors and Amplifiers
April 24, 1996, Las Vegas, NV
nformation: RF Design Seminar Series, Argus Trade hows, 6151 Powers Ferry Rd., N.W. Atlanta, GA 30339. el: (800) 828-0420.

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laximize Performance of Transmission Lines
March 5-7, 1996, Orland Park, IL
April 9-11, Orland Park, IL
May 7-9, 1996, Orland Park, IL
June 4-6, 1996, Orland Park, IL
nformation: Andrew Corporation, Dept. 355, P.O. Box: 000, San Fernando, CA 91341-9978. Tel: (800) 255-1479 xt. 117.

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licrowave Radio Systems
March 16-29, 1996, Washington, DC
CS: A Telecommunications Revolution
April 8-12, 1996, Washington, DC
lectromagnetic Interference and Compatibility EMI/EMC): A Practical Approach to Testing and dvances in Video Technology for Communications and roadcasting
April 8-11, 1996, Washington, DC
lodern Receiver Design
April 15-19, 1996, London, U.K.
Vireless Infrastructure Network Engineering for Cellular, CS, LEO and WPBX
April 15-19, 1996, Washington, DC
licrowave Tubes, High-Power Transmitters, and licrowave Systems: Basic Principles
April 29-May 3,1996, Washington, DC
Grounding, Bonding, Shielding and Transient Protection
April 29-May 3, 1996, San Diego, CA
Satellite Communications Engineering Principles
May 7-10, 1996, Washington, DC
Digital Image Processing
May 13-14, 1996, Washington, DC
Mobile Satellite Communications Systems
May 13-15, 1996, Washington, DC
Modern Digital Modulation Techniques
May 13-17, 1996, Washington, DC
Modern Digital Video Processing
May 15-17, 1996, Washington, DC
Global Positioning System: Principles and Practice
May 20-23, 1996, Washington, DC
Electromagnetic Interference and Control in Modern Communications Systems
May 20-24, 1996, Washington, DC
Mobile Communications Engineering
May 22-24, 1996, Washington, DC
Hazardous RF Electromagnetic Radiation: Evaluation, Control, Effects, and Standards
June 12-14, 1996, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Academic Center, Room T-308, 801 22nd Street, N.W., Washington, DC 20052. Tel: (202) 994–6106 or (800) 424–9773. Fax: (202) 872–0645.

Grounding & Shielding Electronic Systems, and Circuit Board Layout
June, 1996 (dates TBA), Chicago, IL
August 14-16, 1996, San Jose, CA
Information: Continuing Education, University of Missouri-Rolla, 103 ME Annex, Rolla, MO 65409-1560. Tel: (314) 341-4132; Fax: (314) 341-4992.

51st Engineering + Management Program
March 24-29, 1996, Los Angeles, CA
Digital Video Technology
April 10-12, 1996, Los Angeles, CA
Wireless Voice and Data Communications
April 16-19, 1996, Los Angeles, CA
Advanced Communications Systems Using Digital Signal Processing
April 22-24, 1996, Los Angeles, CA
Wired and Wireless Telecommunications Networking
May 20-24, 1996, Los Angeles, CA
Information: UCLA Extension, 10995 Le Conte Ave., Suite 542, Los Angeles, CA 90024. Tel: (310) 825-1047; Fax: (310) 206-2815. E-mail: mhenness@unex.ucla.edu

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RF IC Design of Wireless Communication Systems
April 23-26, 1996, San Francisco, CA
Information: Mead, Microelectronics, Inc., 7100 NW Grandview Dr., Corvallis, OR 97330. Tel: (541) 758-0828; Fax: (541) 752-1405.

Training Program for Cellular, PCS Staff
Independent Learning Program
Information: Virginia Polytechnic Institute and State University, Mobile and Portable Radio Research Group, 840 University City Blvd., Pointe West Commons, Suite 1, Blacksburg, VA 24061-0350. Tel: (540) 231-2970; Fax: (540) 231-2968.

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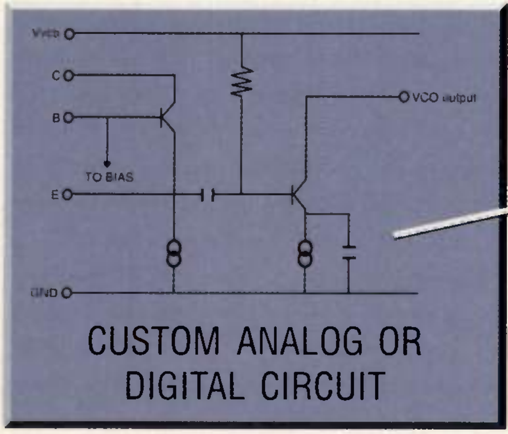

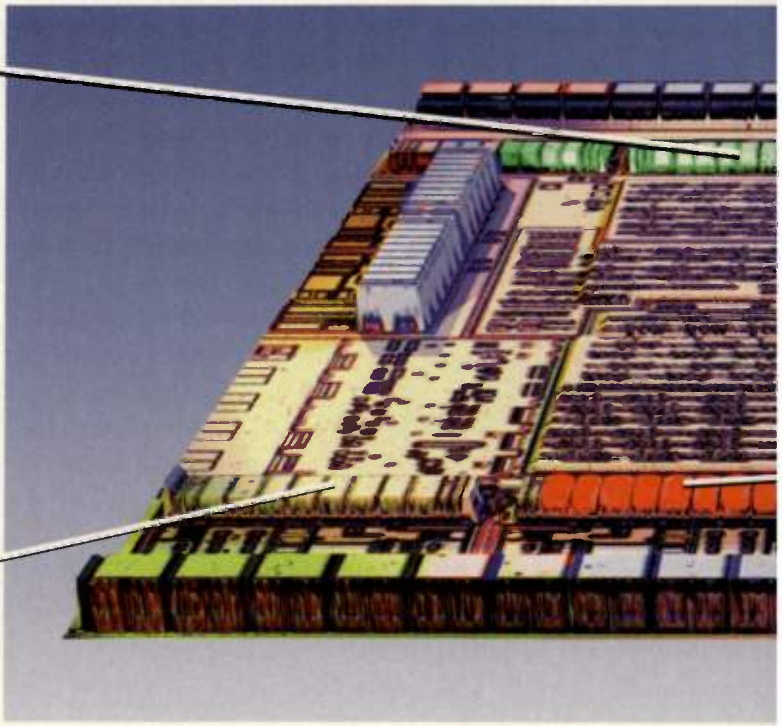

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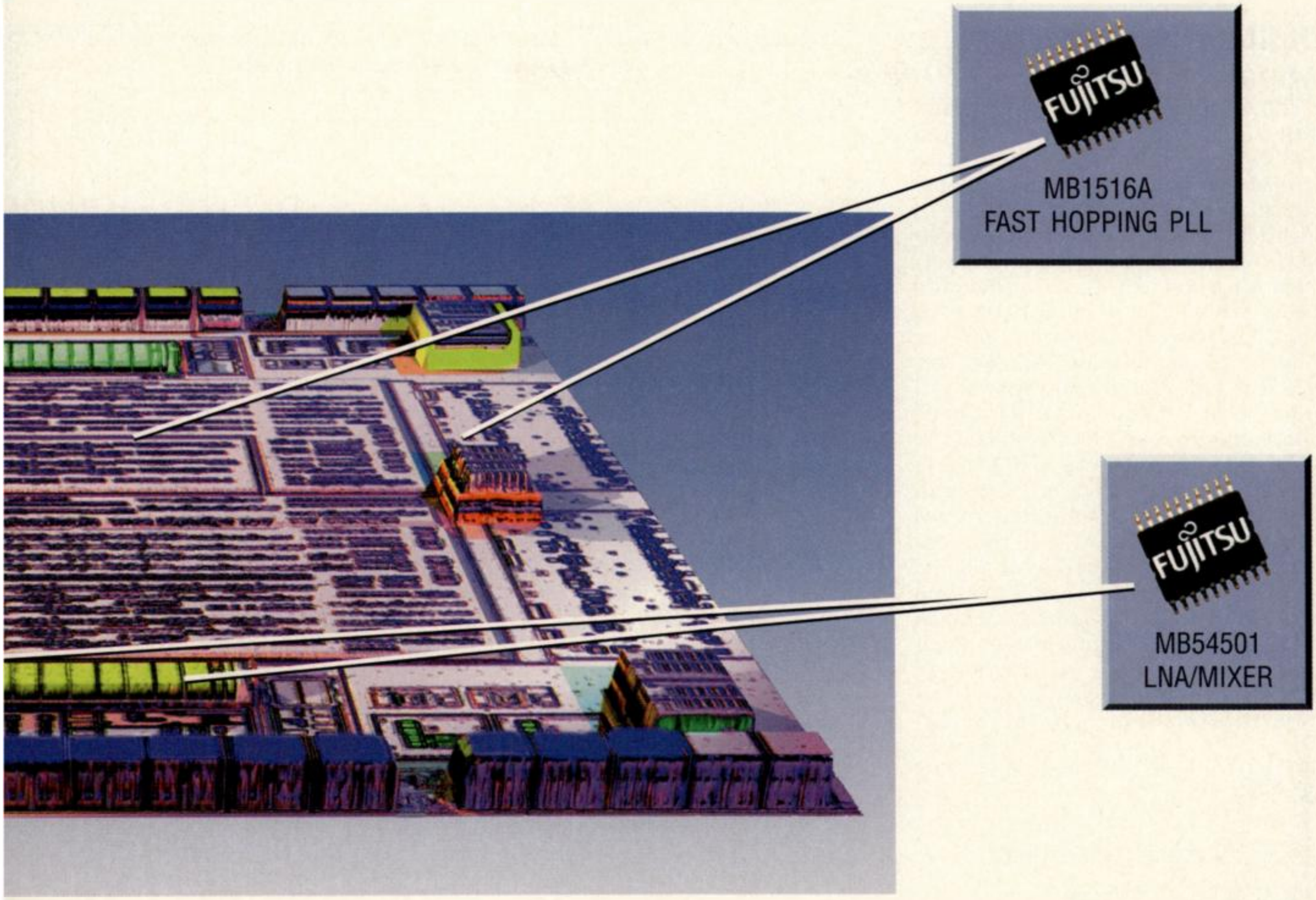

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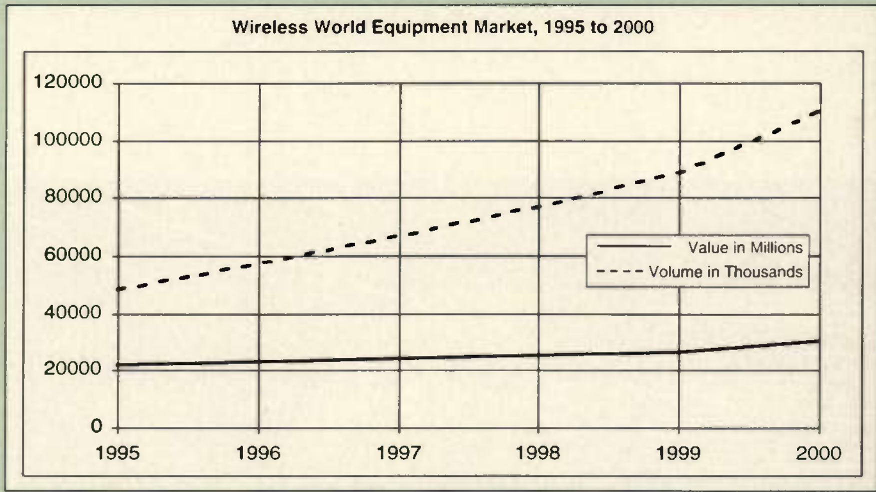
INFO/CARD 16

Digital Radio Moves Ahead in Canada

The Canadian Radio-television and Telecommunications Commission (CRTC) has announced that it is accepting applications for transitional digital radio licenses, to remain in effect until a long-term policy is developed. Along with a prior draft of spectrum allocation for digital radio, and the approval of a standard (Eureka-147/DAB), Canada is moving into the implementation phase of this new service. The new digital radio service will operate in L-band, with the Draft Allotment Plan specifying 1452-1492 MHz. The Plan divides the 40 MHz into 23 channels, each of which can carry up to five CD-quality stereo broadcast programs. Up to five stations having approximately the same coverage area may share a single transmitter. Experimental stations are operating in Toronto, Montreal, and soon Vancouver. Information on these developments can be obtained from Digital Radio Research Inc. (DRRI), 7925 Cote St-Luc, Montreal, Quebec, Canada H4W 1R5; fax: (514) 485-5885.

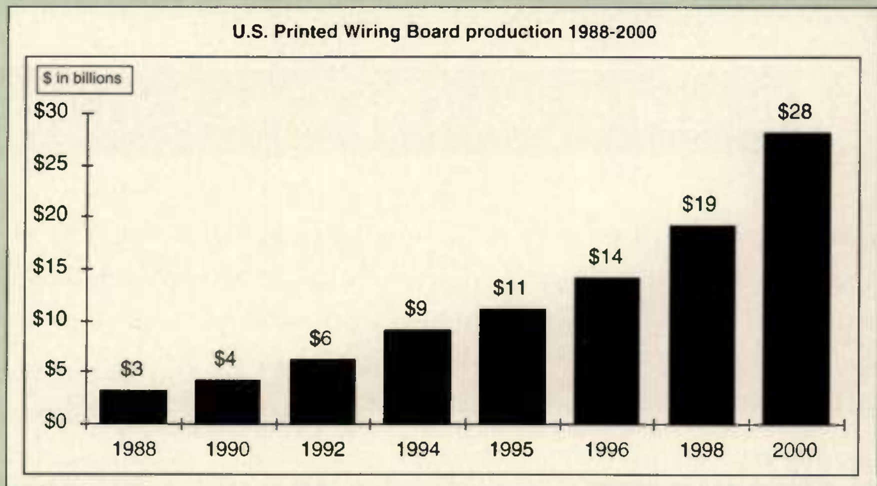
HP EEsof Sponsors Design Contest

An HP Omnibook 600C notebook PC and HP 48GXX scientific calculators are the prizes to be awarded in a contest for HP EEsof users. Entries are papers, and eventually 20-minute talks, on the entrant's design experience with HP EEsof tools. Entry categories are: 1) High frequency system design; 2) High frequency circuit design; and 3) Component or antenna design with EM simulation. Abstracts are due April 1, 1996 and should be no more than one page in length. Finalists will be notified by May 2. Completed papers are due June 10, for presentation during the MTT-S Microwave Symposium, June 18-20. Entries will be judged on innovation, performance, measured vs. modeled results, demonstrated use of HP EEsof tools, and return on investment (time savings). Abstracts should include the author's name, title, company, address, telephone, fax and e-mail address. Submit abstracts to: Tom Phillips, Customer Support Manager, HP EEsof, M/S 2USX, 1400 Fountain Grove Parkway, Santa Rosa, CA 95403; tel; (707) 577-4345.



Report Outlines Wireless Growth

Allied Business Intelligence, Inc. is completing a report, "The Wireless World Strategic Outlook," which projects the growth of six primary wireless technologies — Cellular Telephony, Direct Broadcast Systems (DBS), Global Positioning Systems (GPS), Personal Communications Systems (PCS), Very Small Aperture Terminals (VSATs) and Wireless Local Area Networks (WLANs). Cellular and PCS subscribers are expected to grow at a compound rate of 25% through the year 2000, VSATs will grow from 150,000 terminals in 1994 to over one-half million by 2000. DBS and WLAN are in their infancy, and are expected to grow into substantial markets. GPS alone is expected to be a $5 billion market by 2000, including stand-alone and in-system applications. For a copy of the entire report, contact Bill Britton or Andy Fuertes at (516) 624-3113; or check the firm's Web page at: http://www.alliedworld.com/



Healthy Growth Predicted for Printed Circuits and Assembly Industries

The Institute for Interconnecting and Packaging Electronic Circuits (IPC), projects that U.S. electronics manufacturing service providers (contract assemblers) will grow at a 20 percent average annual rate from 1994 through 1999. U.S. printed wiring board production is expected to grow at a 9.2 percent during that time. A study of 49 contract assembly companies indicates that the industry is now $11.4 billion. The accompanying chart (above) shows how the manufacturing services industry has grown, and how it is expected to continue to grow.

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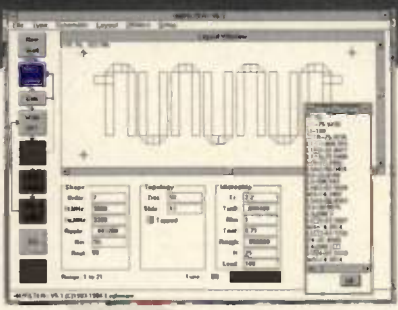

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Dielectric resonator
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Port SAW hybrid
Port SAW MOSFET
Pierce and Colpitts crystal
Driscoll crystal
Butler overtone
Overtone with multiplier

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L-C tee
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T-line single/double stub
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INFO/CARD 17

Wireless Technology Workshop Seeks Papers

A Call for Papers has been issued for the Advanced Technology Workshop on Wireless Communications, August 21-23, 1996 in Boulder, Colo. Sponsors are ISHM, IEEE MTT and IEEE CPMT. Presentations on significant work in wireless engineering are sought — systems, packaging, components, antennas, interference, measurements and modeling. Send abstracts by April 1996 to: Dr. Modest Oprysko, IBM T. Watson Research Center, P.O. Bo 218, Yorktown Heights, NY 10598; fa (914) 945-1974; e-mail: modest@wat son.ibm.com

Additional information can b obtained from the General Chair: D Roger Marks, NIST, 325 Broadwa MC 813.06, Boulder, CO 80303; te (303) 497-3037; fax: (303) 497-7828; e mail: marks@nist.gov

Business Briefs

Svetlana receives $3 Million investment from Defense Enterprise Fund — Svetlana Electron Devices has received an investment of $3 Million from the Defense Enterprise Fund, a private venture capital fund capitalized by the U.S. Government for financing of joint ventures that promote defense conversion in Russia. The new funding will strengthen the Svetlana joint venture and contribute to its growth, particularly in the rapidly growing foreign industrial and medical markets.

Schaffner opens liaison office in China — Schaffner announces the first field test instrumentation liaison office onside China, dedicated solely to providing advice and technical support on the issues of electromagnetic compatibility. With expanding imports, and an absence of its own national standards, Chinese manufacturers are beginning to design and test products to international IEC standards.

Pinpoint Communications begins vehicle location system — Pinpoint Communications, Inc. has begun operation of its first commercial network for vehicle location and mobile data communications in Dallas, Texas. The technology used is a proprietary land-based two-way vehicle tracking and communications system capable of tracking thousands of vehicles per second. Efficiencies gained in dispatch of customer vehicles is expected to result in significant savings, with payback of investment in less than a year for some customers.

NTS offers fast-track EMC/EMI product testing — Fast-track testing for CE Mark compliance is offered by National Technical System at west and east coast laboratories. Eight facilities are operated nationwide by NTS, along with a mobile EMC test van for on-site testing.

Haefly Trench and Tettex Instruments at single address — North American operations of Haefly Trench, Inc. and Tettex Instruments have been combined at a single location — Haefly Trench, Inc., 1308 Devils Reach Road, Woodbridge, VA 22192; tel: (703) 494-1900.

RF Industries adds new distributors — RF Industries, Ltd. RF connector products are now available from Mueller Electric Company, an Ohio-based distributor, and Neumann Electronics, based in California, with office in Arizona and Colorado.

Microdyne Corp. announces CE Mark compliance — Microdyne Corp. has six key Aerospace Telemetry Division product categories that have received the CE Mark for European electromagnetic compatibility certification. The products include receivers, combiners, accessories and adapters for tracking, data reception and display.

Merix and Hewlett-Packard cooperate in DYCOstrate® technology — Merix Corporation and Hewlett-Packard Company's Printed Circuit Organization have entered a sub-licensing agreement that allows Merix to use the DYCOstrate interconnect substrate technology developed by Dyconex AG of Switzerland. The technology uses plasma etching to generate fine-diameter vias for high-density circuit boards.

Contracts

Scientific-Atlanta to delive Inmarsat-M systems — Farhan Com mercial Company of Saudi Arabia ha signed a contract with Scientific Atlanta for six vehicle-mounted an five semi-transportable Multi-M sate lite communications systems. The sys tems will be used to provide digita two-way voice, and fax services t remote areas of Saudi Arabia using a Inmarsat satellite. The first of the sys tems was shipped in December 1995.

Loma Scientific signs MMDS agreement — Loma Scientific Inter national of Torrance, California, ha entered into an agreement with CMN Consorcio Multipunto Multican S.R.L. to provide broadcast transmit ters and other equipment for a majo MMDS/wireless cable facility servin Asuncion, the capital of Paraguay. Th system will deliver 31 channels o broadcast quality programming usin microwave frequencies.

AOA to provide GPS receivers – Allen Osborne Associates, Inc. (AOA will provide the new GPS receiver using in the NAVSTAR GPS Contro Segment's Monitor Stations. Th receivers will replace receiver installed in the early 1980s, and wil greatly increase the reliability of th satellite measurements. AOA will pro vide a modified version of its SNR-1 SM TurboRogue®, a 12-satellite receiv er capable of directly tracking th encrypted P code (Y-code) and of remov ing Selective Availability errors. Th receiver is capable of producing pseudo range measurements with a precisio of 10 mm and carrier phase measure ments with a precision of 3 mm.

Wyle Laboratories awarded con tract to operate Standards Labo ratory — Wyle Laboratories has bee awarded an $18.9 million contract t operate the Air Force Primary Stan dards Laboratory (AFPSL) in Newark Ohio. Wyle's contract runs from December 1995 through Septembe 1996, with one-year renewal options through September 2000. Under th contract, Wyle will calibrate standard for the Air Force Precision Measure ment Equipment Laboratories (PMELs) worldwide, for numerous other government organizations and industrial firms, plus many foreig military laboratories. RF

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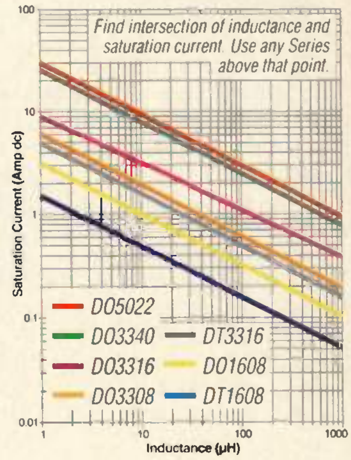
Selection Guide



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INFO/CARD 22

Packaging Drives RF Power Device Development

By Gary A. Breed
Editor

The amplification of signals to medium and high power is one of the most universal tasks in RF engineering. To make en engineer's job easier, RF power device manufacturers continue to develop new processes and continue to work to provide the most advantageous packaging for current communications and industrial applications.

Power semiconductors have several criteria placed upon them by their users, which may include any combination of the following:

- Low-inductance leads
- Efficient heat removal
- Convenient mounting configuration
- Low-cost standard packaging
- Capability of automated assembly

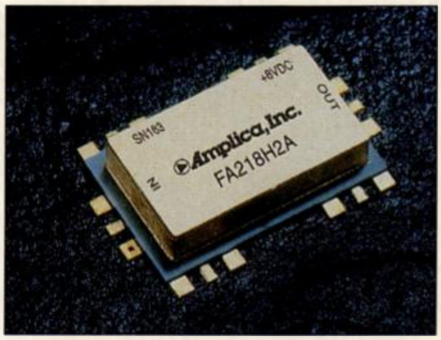
In the past, RF power devices were almost always assembled in ceramic packages with stripline leads and a substantial heat-dissipating mounting surface. With demand growing for lower cost solutions, manufacturers are now using different types of packages for these devices.

Instead of mounting die in expensive packages designed specifically for RF power applications, they are now mounted on standard lead frame packages, such as those illustrated to the right. Hermetically-sealed ceramic has been replaced by plastic encapsulation in the most cost-sensitive applications.

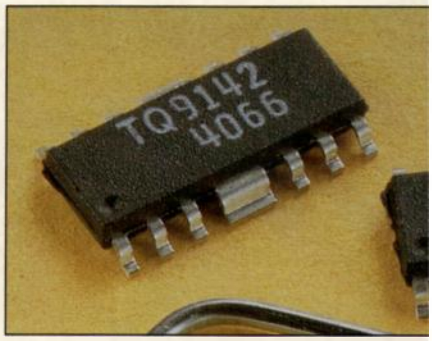
The primary difficulty in using inexpensive packaging is maintaining good RF performance in an environment that was not designed with this use in mind. Most devices use multiple-pin grounding, sometimes with a solid bus replacing adjacent pins, to lower the impedance of the ground connection. Devices are also characterized to include the inductance and stray capacitance introduced by the non-ideal packaging, or internal matching may be adapted in compensation. In many cases, the tradeoffs have been considered acceptable, and good efficiency has been obtained without resorting to external matching.

At high power (more than a few watts) the conventional ceramic packages remain the best choice for maintaining RF performance while providing adequate heat dissipation. For lower-frequency operation, such as 13.56 or 27.12 MHz ISM, and medium-wave to HF communications, some manufacturers are providing high power devices in conventional TO-220 or TO-247 packages. Package limitations appear to restrict these applications to under 50 MHz.

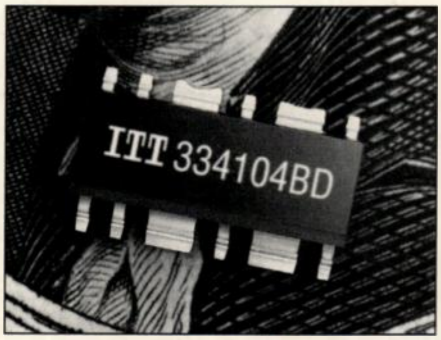
In summary, size and cost, plus standardization for manufacturability are becoming new criteria for RF power devices. As applications with these requirements grow, we will see more RF power products move to these packaging styles. *RF*



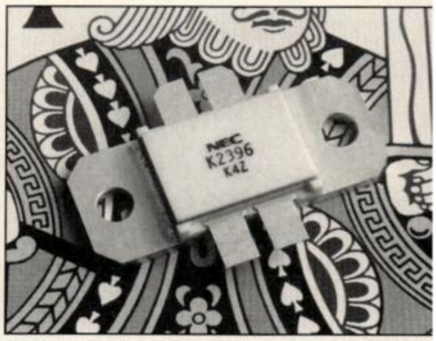
For 2-18 GHz operation, Amplica has adopted the EZ-Pak® surface mount package, which provides an effective transition from the amplifier package to the circuit substrate.



TriQuint's new GaAs cellular amplifier IC is provided in a SO-16 plastic package for lowest cost, modified for improved heat sinking by replacing the center pins with a solid bus.



ITT GTC also uses a modified SO-16 package for its 1.2-watt cellular amplifier, combining adjacent pins for heat sinking and lower impedance ground connection.



A more conventional stripline configuration is maintained for higher power devices like the NEC 100-watt MOSFET part designed for Class AB operation in the UHF television band.

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10.7 MHz

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2	TE5000	3	3.75	20	18.0	-	-	2	1.0	50	1800//+4
4	TE5010	3	3.75	30	14.0	-	-	3	2.0	60	1500//+3
6	TE5020	6	3.75	60	12.5	-	-	4	2.0	70	1500//+3
8	TE5030	6	3.75	60	10.0	90	12.5	5	2.0	80	1500//+3
2	TE5040	3	6.50	20	30.0	-	-	1	1.0	50	2700//0
4	TE5050	3	6.50	30	15.0	-	-	2	2.0	75	3100//0
6	TE5060	6	6.50	60	19.5	-	-	3	2.0	90	3100//0
8	TE5070	6	6.50	60	13.0	80	17.5	4	2.0	100	3100//0
2	TE5080	3	7.50	20	35.0	-	-	1	1.0	50	3000//0
4	TE5090	3	7.50	30	17.5	-	-	2	2.0	75	3300//0
6	TE5100	6	7.50	60	22.5	-	-	3	2.0	90	3300//0
8	TE5110	6	7.50	60	15.0	80	20.0	3	2.0	100	3300//0
2	TE5120	3	15.0	20	70.0	-	-	1	1.0	35	5000//-1
4	TE5130	3	15.0	30	35.0	-	-	2	2.0	60	5000//-1
6	TE5140	6	15.0	60	45.0	-	-	2	2.0	90	5000//-1
8	TE5150	6	15.0	60	30.0	80	40.0	3	2.0	100	5000//-1

21.4 MHz

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

45.0 MHz

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

70.0 MHz

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

90.0 MHz

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

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A Cable Telephony Transceiver Test System

:y Mike McNatt
'ellabs Operations, Inc.

Data and telephone communications :an coexist with TV broadcast stations n a Community Access Television :ATV) system over the same cable. RF ·ansceivers or "cable modems" make hese communication links possible. lowever, intermodulation products 'rom multiple stations transmitted hrough network nonlinearities), dis- ortion and interference are serious oncerns in any CATV system. These ignals are often just below a visible or udible threshold, and even very small istortion contributions by an RF ·ansceiver can cause problems. The olution is to test the transceivers thor- ughly prior to installation on an ctive system.

The automated test system presented here is used to verify the perfor- nance of various RF transceivers :able modems) used in a telephony or ata-over-cable TV system. The two nodules tested are a Forward Trans- eiver at the CATV head end and a Return Transceiver at the customer ite. The Forward Transceiver trans- nits in the HI or downstream band 250-750 MHz) and receives in the LO r upstream band (5-42 MHz). The Return Transceiver receives in the HI and and transmits in the LO band. 'ull-duplex RF-modulated forward nd return telephony and data signals re transmitted simultaneously with ıormal CATV signals over the same 5 ohm cable.

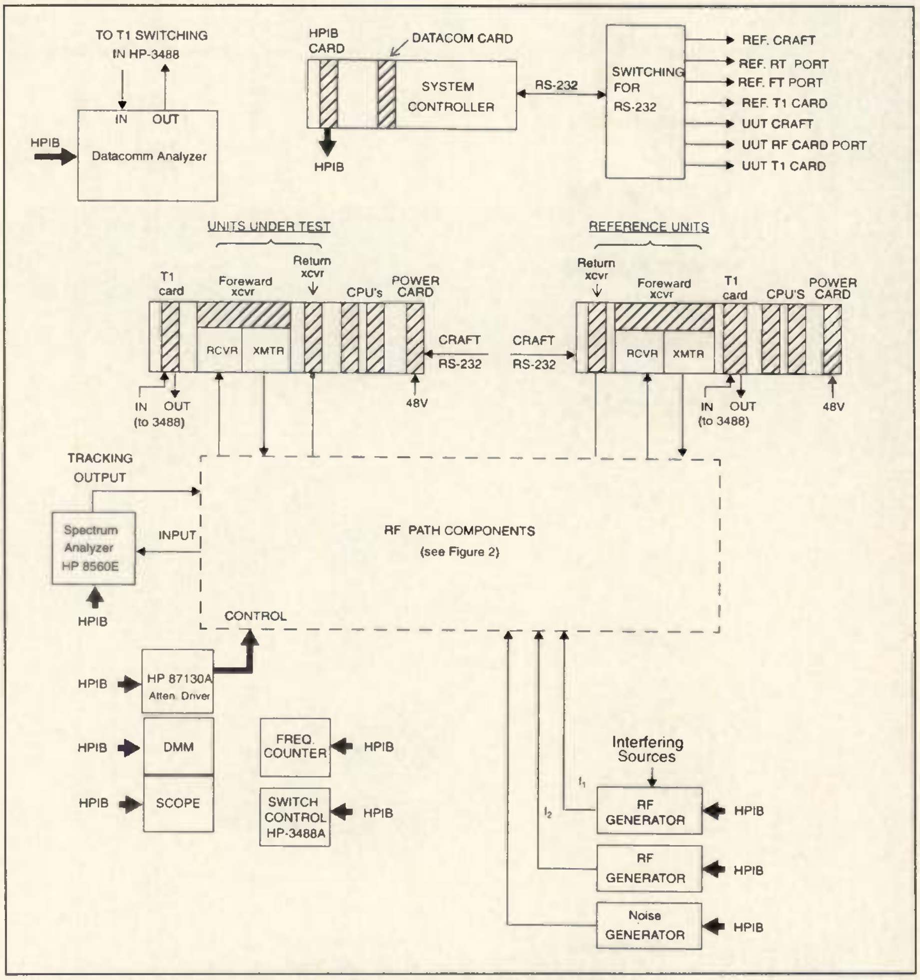


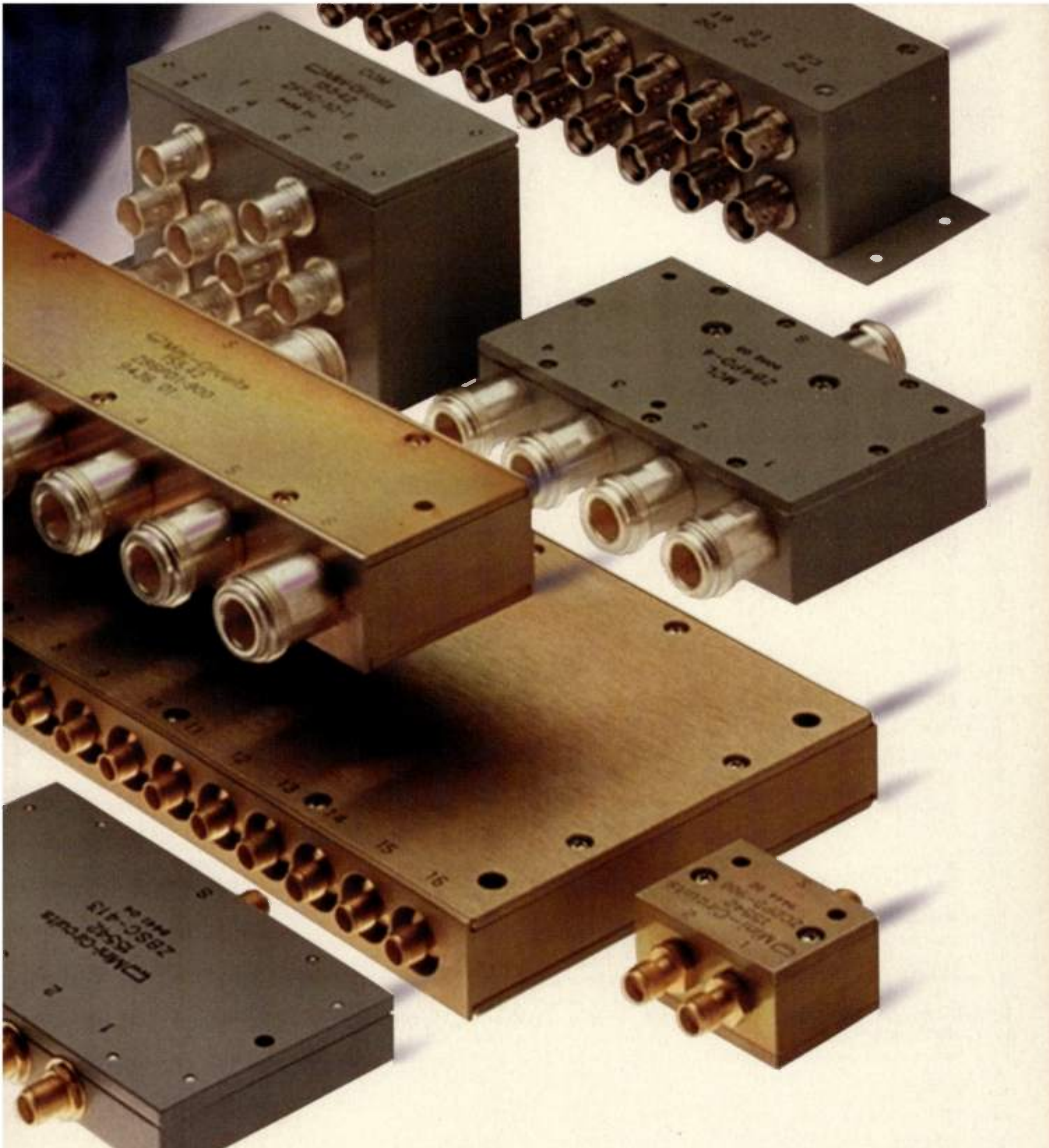
Figure 1. Block diagram of the cable modem test system.

System Overview

Figure 1 shows a block diagram of he system. The System Controller ontrols all IEEE-488 instruments nd communicates with all modules y use of RS-232 interfaces. The data- omm analyzer provides a T1 refer- nce data transmitter and analyzes it errors by use of a data receiver. 'he T1 data rate is 1.544 Mbps, vhich consists of 24 channels of DS0 64 kbps) telephone signals plus 8 :bps of signaling. Note that the 64 kbps data rate results from sampling an analog phone signal at 8 kbps with an 8-bit code.

The T1 data stream is input to the T1 interface card, which outputs a serial data stream to the unit under test (UUT) where it is transmitted to a reference receiver. A reference transmitter returns the same data stream to the UUT receiver, where it is decoded and routed to the datacomm (bit-error-rate) analyzer for comparison to the transmitted data stream. A spectrum analyzer, the key RF measurement device in the system, performs the following measurements:

- Transmitter output power, center frequency and bandwidth
- Transmitter carrier phase noise, harmonic generation and adjacent channel interference
- Receiver input power level and the level of disturbing signals, such as the noise generator and the RF generators



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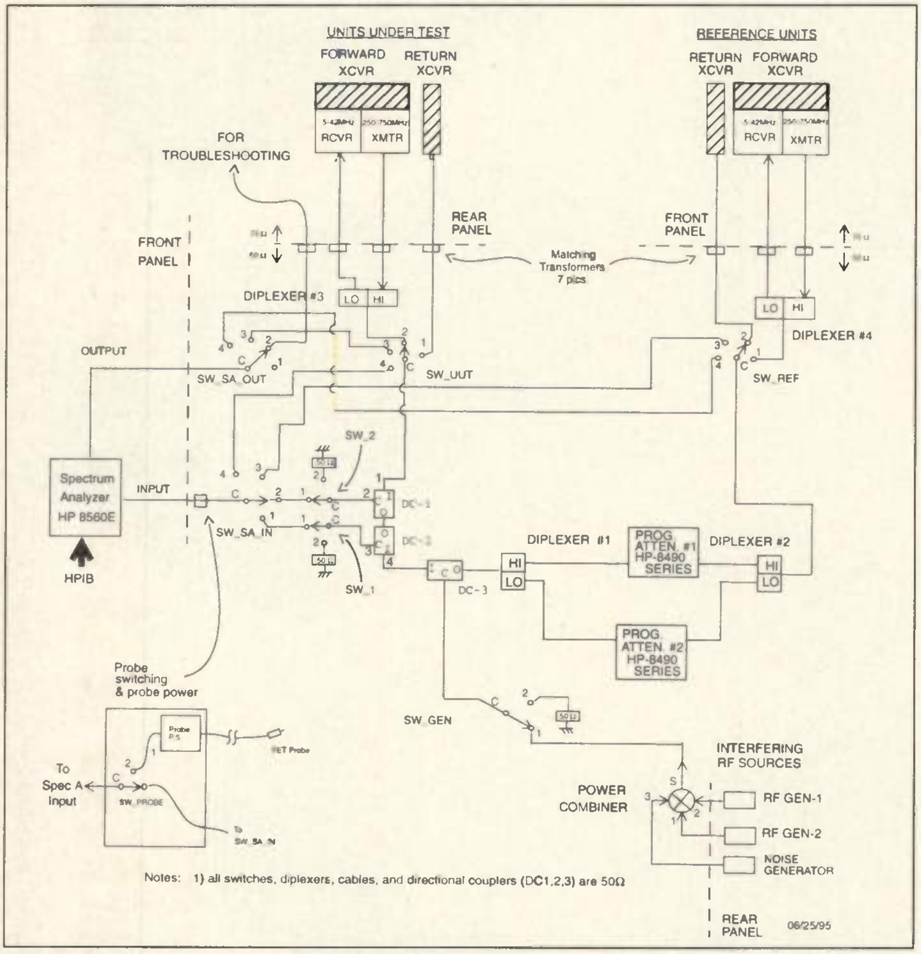

Figure 2. Detail of the "RF Path Components" portion of the system diagrammed in Figure 1.

Many non-RF tests are performed (LEDs, power drain, alarm functions, clock tracking, etc.) that are beyond the scope of this article.

Test Philosophy Using Reference or "Gold" Units

Two main approaches were considered in the design of the test system. The first involved specialized equipment that would generate and then receive/decode the complex QPSK (Quadrature Phase Shift Keying) data used by the system. This approach involved the use of expensive, specialized test equipment. Therefore a more economical and completely sufficient second technique — that of using reference, or "Gold" modules — was selected. Figure 1 illustrates the use of reference modules. To test a Return Transceiver UUT (unit under test), the reference Forward Transceiver transmits the forward RF signal to the UUT and receives the UUT's return RF signal. A T1 (digital data) test set generates the test signal. The system T1 card converts the serial data stream to a serial time division multiplexed data stream on the backplane, which inputs to the reference Forward Transceiver (FT). The FT then constructs a QPSK digital signal which modulates the RF transmission. The RF is received by the UUT, which demodulates the QPSK and converts it to the TDM signals for the backplane. These signals are received by the T1 card, looped back and returned via the backplane to the UUT. The UUT then transmits by use of the return RF channel back to the reference FT. This information is then returned to the T1 test set for comparison to the data transmitted. Testing an FT is just the reverse of the above: the Reference Return Transceiver transmits the return RF signal to the UUT and receives the UUT's forward RF signal. RS-232 interfaces are used under program control to change the transmit and receive frequencies and output levels of the transceivers.

Figure 4 illustrates in more detail how the T1 cards in the UUT and reference systems are connected to accomplish bit-error-rate testing. Note that a complete RF link is used — both transceivers are operating full duplex continuously.

A Tutorial on Some RF Test System Components

Figure 2 is an expanded section from Figure 1. It shows the various components used to route, attenuate and sample transmitter power from several sources. These sources include reference modules, UUT modules and generators that provide interfering test signals. The interference comes from two RF sources for intermodulation and adjacent-channel testing, and a noise source. IEEE-488 bus commands are used to control most equipment. The components in Figure 2 are described below.

Diplexer — Within this component are two filters; a high-pass filter (HPF) and a low-pass filter (LPF), both connected to the common port. Note that identical Diplexers #1 though #4 each have one common port, a LO port (connected to the LPF) and a HI port (connected to the HPF). Signals in the HI band can pass bidirectionally between the HI port and the common port, but signals in the LO band are blocked from taking this path. Conversely, signals in the LO band can pass bidirectionally between the LO port and the common port, but signals in the HI band are blocked from taking this path. Insertion loss for in-band signals is about 1 dB; attenuation for out-of-band signals is 45 dB minimum.

Directional Couplers — These devices are industry-standard directional couplers with port labeling as follows: I = input port, O = output port and C = coupled port. Bi-directional attenuation from port-to-port is as follows: about 2 dB insertion between ports I and O, about 10 dB coupling between ports I and C, about 35 dB isolation between ports O and C.

DC-1 and DC-2 — In order to sample transmitter outputs, two directional couplers (DC-1 and DC-2) are connected "back-to-back," or output-to-output, as shown in Figure 2. DC-2 samples transmitter power from the Reference Transceivers and DC-1 sam-

ples transmitter power from the UUT. Note that due to RF Switch SW_SA-IN, the spectrum analyzer displays only one transmitter output (either the UUT or the Reference module) at a time. DC-3: This unit couples interfering test signals into the RF path from port C to port I. The interfering signals from the two RF generators and the noise generator are thereby coupled to the UUT receivers (Forward Transceiver or Return Transceiver). However, the interfering signals are prevented from traveling to the Reference Transceiver from port C through port O. Note that when the generators are not used, RF switch SW_GEN connects a termination to port C to insure proper operation of the device.

Programmable Step Attenuators #1 and #2 (PA-1 and PA-2) — Each PA is actually two PAs connected in series: one PA supplies 0-110 dB in 10 dB steps and the other PA supplies 0-11 dB in 1 dB steps. Thus the maximum attenuation of the pair is 121 dB. Note that the operation of Diplexers #1 and #2 causes PA-1 to attenuate only HI band RF signals and PA-2 to attenuate only LO band RF signals. The attenuators are under IEEE-488 control by use of the HP-87130A Attenuator/Switch Driver.

Matching Transformers — All transceiver modules have RF ports that are 75 ohms. Matching transformers are used to match the 75 ohm modules to the 50 ohm devices used in most of the RF path. These "N" connector transformers have less than 1 dB insertion loss at 1 GHz.

RF Switches — Two types of RF coaxial SMA-female-connector 4 GHz switches are used: SPDT and SP4T. The meaning of each switch label, as referred to in Figure 2, is as follows (SW = switch):

SW_REF — Reference module control
SW_UUT — UUT module control
SW_SA_IN — Spectrum analyzer input control
SW_SA_OUT — Spectrum analyzer output control
SW_GEN — Generators (connects the interfering sources to UUT receiver)
SW_1, SW_2 — These switches terminate the unused port of directional couplers DC-1 and DC-2
SW-PROBE — Spectrum analyzer probe

All RF switches are under IEEE-488 control by use of the HP-87130A Attenuator/Switch Driver.

Power Combiner — This commercial power splitter/power combiner combines three signals into one output. The attenuation between any input and the output is typically 6 dB, and the isolation between inputs is typically 20 dB or better.

Probe Switching — A FET-input oscilloscope probe is used for troubleshooting UUTs. The scope probe is a fraction of the cost of a higher-quality laboratory RF probe, but provides satisfactory performance for a production-line environment. RF switching connects the probe to the spectrum analyzer when in the troubleshooting mode.

Distribution Board (HP-84941) — This board distributes the control signals from the HP-87130A Attenuator/Switch Driver, an IEEE-488 controlled instrument, to 31 4-pin connectors. The signals to each 4-pin connector control one SPDT switch, one section of a SP4T switch or one section of a programmable step attenuator. It takes four of the 4-pin connectors to control one SP4T switch or one programmable step attenuator. The distribution board has no active components and connects to the HP-87130A with a 68-wire cable. Only one distribution board can be controlled by the HP-87130A.

Driver Board (HP-84940A) — This board allows expansion of the HP-87130A (see above) to control 31 additional switches and/or programmable attenuator sections. Up to 7 driver boards can be used with the HP-87130A for a total capacity of 248 switches. The driver board has active components and connects to the HP-87130A by use of a 36-wire cable. The board includes a distribution area, similar to the HP-84941 Distribution Board, of 31 4-pin connectors.

dBmV vs. dBm

The CATV unit of "dBmV" relates to dBm as follows:

dBmV (75 ohm system) = dBm (50 ohm system) + 48.75 dB

As an example, the FCC requires that each CATV channel reaching a television must be at least 0 dBmV, or about –49 dBm.

Spectrum Analyzer Measurements — Path Loss Compensation Example

In Figure 2, "RF Path Components," various combinations of components in the RF path cause spectrum analyzer (SA) measurement errors or losses that vary depending on the particular path. These losses are corrected in the test software by compensation factors (in dB) that are added or subtracted to the SA readings. These path losses cause a difference in the SA power displayed and the actual power at the point of interest — whether at a receiver input or at a transmitter output. Figure 3 shows a typical compensation factor calculation for measuring UUT receiver input level. The figure shows why the receiver input level observed at the SA input is 6.5 dB less than the power actually delivered to the receiver input (25.5 – 19 dBmV). See the box above for a discussion of the CATV "dBmV" unit.

In Figure 3, the two "10 dB" losses occur in directional couplers DC-1 or DC-2 from the input to the coupled port. The box with the "1 dB" label represents losses in the RF path due to switches and cables. The "2.5 dB" box represents the loss in Diplexer #3. Note the 1 dB insertion loss of each directional coupler.

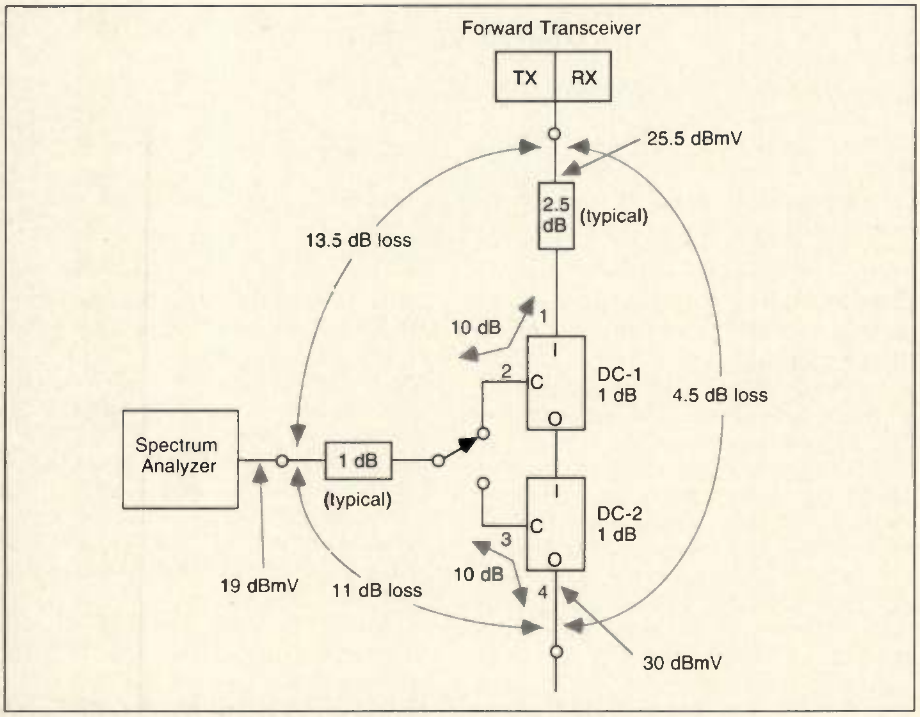

Figure 3. Typical RF path loss measurement compensation.

Assume a signal of 30 dBmV at the input of DC-2 due to the reference transmitter. The result is 19 dBmV at the SA input and 25.5 dBmV at the receiver input. This can be seen by subtracting the various path losses from the 30 dBmV input. Note also that there is a 13.5 dB loss shown between the SA input and the transmitter output.

Other RF path measurements are also similarly corrected. Actual measurements on the paths show close agreement with the above discussion, with some obvious frequency dependencies taken into account.

Spectrum Analyzer Noise Measurement Compensation

In certain types of spectrum analyzer (SA) noise power measurements, the SA reads 2.5 lower than the actual power present — thus 2.5 dB must be added to the figure reported by the SA in these cases. The factors causing thi are discussed in HP Application Not 150-4, "Spectrum Analyzer Noise Mea surements."

The SA "Channel Bandwidtl (Power) Measurement" feature is use by the system to measure transmitte output power, receiver input powe levels, etc. These measurements mus be corrected by the 2.5 dB factor onl when testing a modulated channel The unmodulated (carrier-only) signa does not use the correction. The cor rection is made automatically in th test software as required. However when using the Channel Bandwidtl Power Measurement feature manual ly while troubleshooting modulate signals, the 2.5 dB must be added b the operator to obtain correct mea surements.

Conclusions

This article has provided three lev els of information about test system for CATV transceivers. A testing approach has been detailed — that o using reference units. This keep equipment costs lower than a approach that requires special, com mercial equipment to both generat and then analyze test signals. Th design of the test system has bee shown. This includes the presentatio and discussion of an overall block dia gram and a detailed component-leve schematic. Finally, a tutorial of man RF test system components has bee presented. Components such as diplex ers, directional couplers, power split ters, impedance transformers, an various RF switches have bee described. Typical uses of each of thes components has been illustrated in practical, industrial manufacturin test system.

Acknowledgments

I am very grateful to Walt Read fo providing technical leadership for th Test System project, to Dave Irion fo creating the software that runs th test system, to Steve Heitke for th design of the telephone test interface to Bill Lang and his team for the con struction of the System, and to Ceci Deisch for his help with various R issues. *RF*

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1. Robert A. Witte, *Spectrum and Net work Measurements*, Prentice-Hall 1991. Excellent information on th nuances and finer techniques of usin

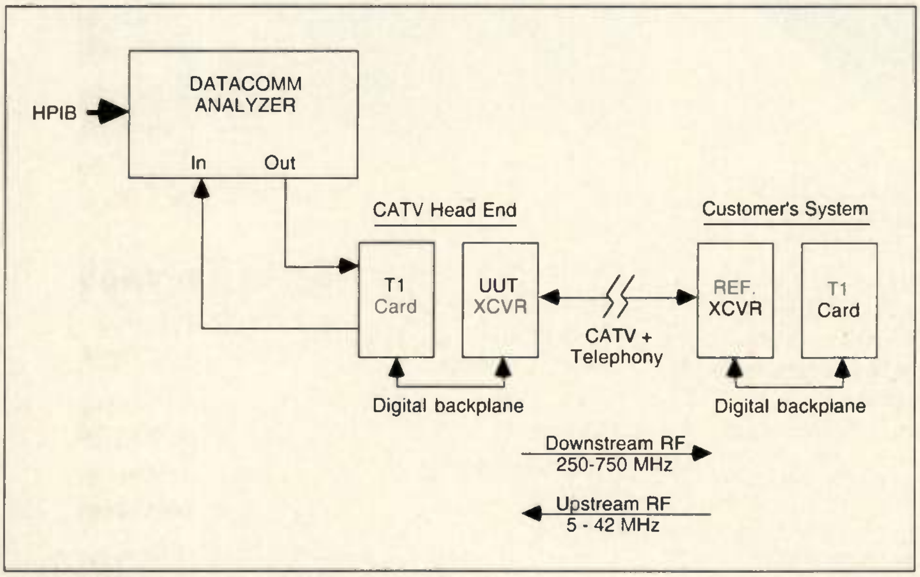

Figure 4. Bit error rate testing signal flow.

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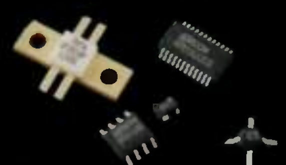

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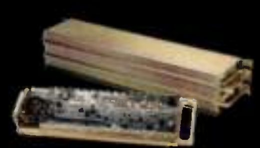

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About the Author

Mike McNatt is a Staff Engineer at Tellabs in Bolingbrook, Ill., and designs test systems for RF products. Mike received his B.S.E.E. at Oklahoma State University and M.S.E.E. from the University of Missouri (Columbia, Mo.). Prior to joining Tellabs, he was at Amoco Technology where he was on the design team for an RF-to-fiber-optics interface for CATV systems. Mike has had articles and "Ideas for Design" published in Electronic Design, EDN and Byte Magazine. He is a member of the IEEE, a registered professional engineer and is a licensed radio amateur (WB5RRP). He can be reached at Tellabs Operations, Inc., 1000 Remington Blvd., Bolingbrook, IL 60440; by phone at (708) 378-6198 or via e-mail: mcnatt@tellabs.com.

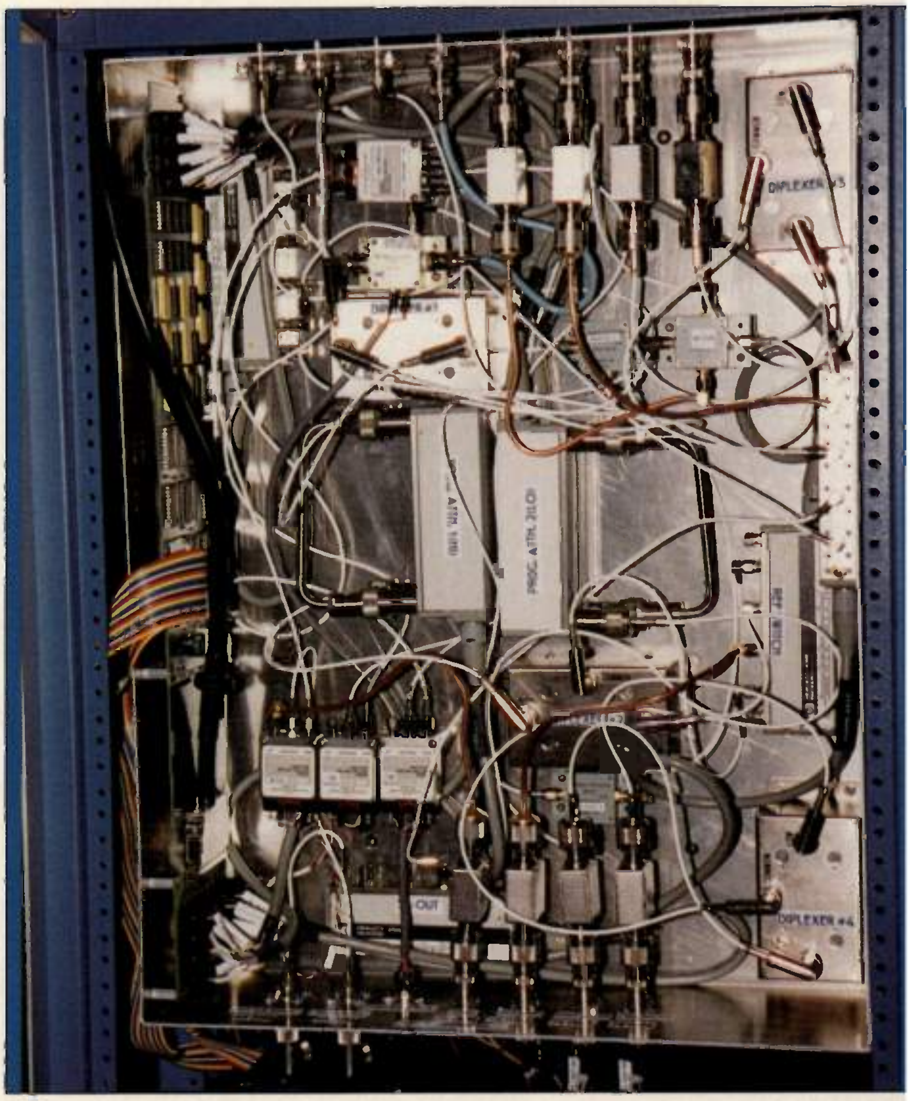
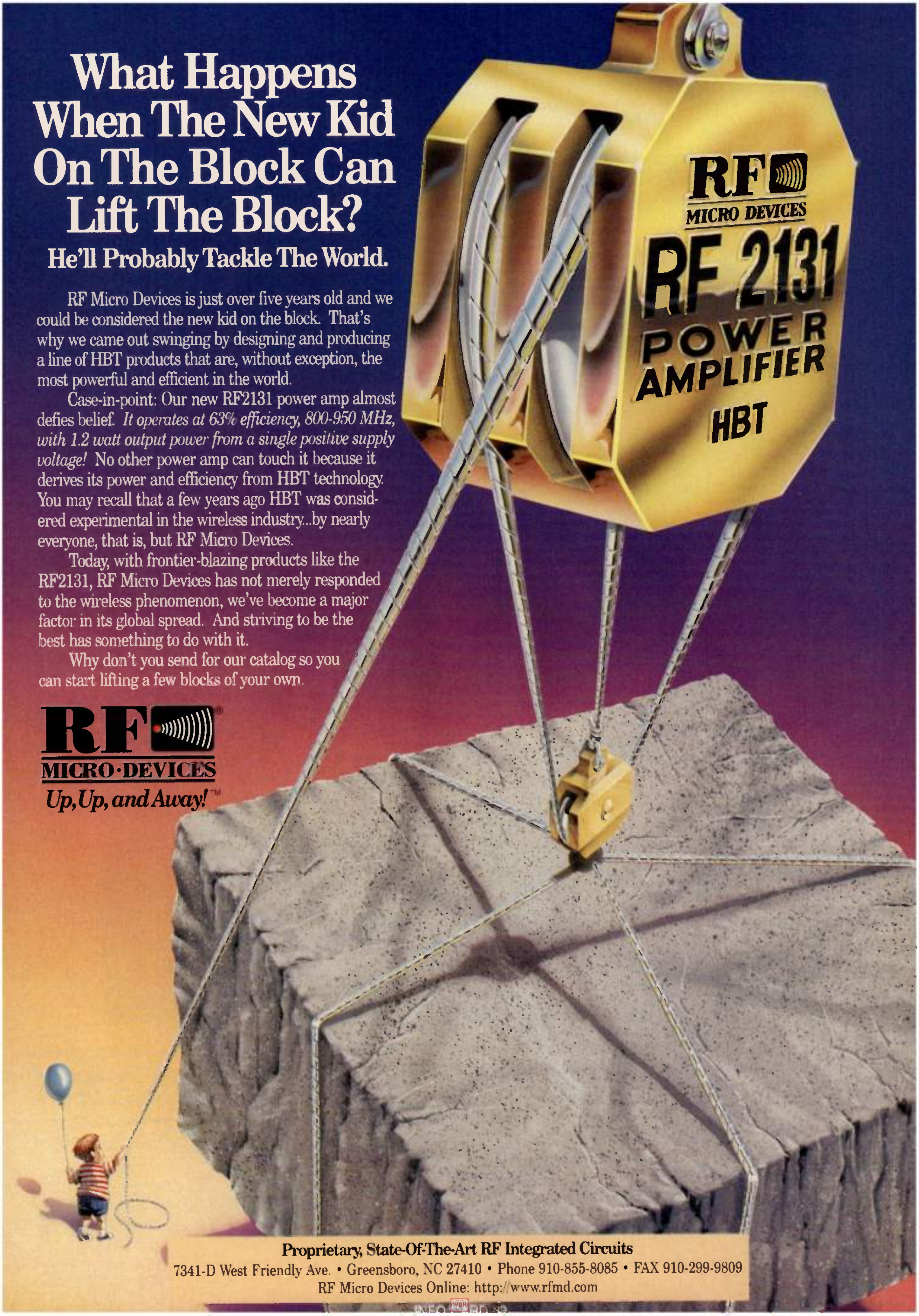


Photo of the cable telephony test system RF components assembly describe in the article. (Photo courtesy of Dan Dawdy)

Applications of Cable Modems — A Hot Technology

Cable modems will allow cable TV operators to offer several services, some of which are potential "killer applications" for CATV. These include telephony-over-cable, interactive TV, and Internet connections, with the latter providing much excitement in domestic markets. For Internet applications, cable modems can supply data rates of one to 40 Mbps. As a comparison, inexpensive personal computer analog modems can provide up to a 28.8 kbps thoughput.

International applications will be concentrated in telephony-over-cable. In many countries, the installed base (miles) of CATV lines greatly exceeds the installed base of telephone lines. Thus to provide telephone service where CATV lines exist only involves adding cable modems without adding more miles of cable or copper twisted-pair. In addition, powering these new services is not a problem. CATV service in many areas carry 60 or 90 VAC right over the same cable for powering line amplifiers and other devices.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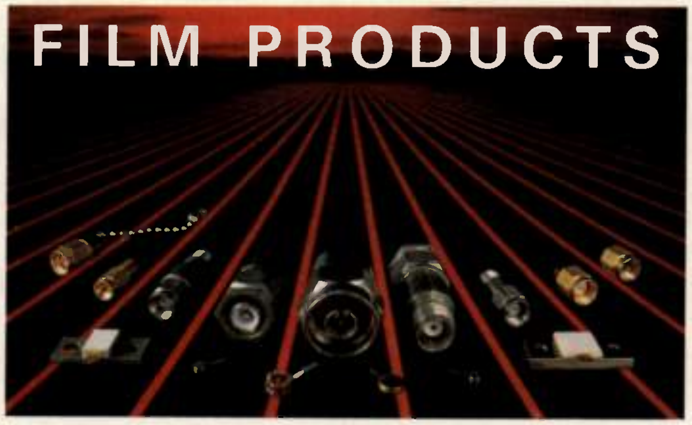

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High Linearity HBT Amplifier Targets Multicarrier Systems

William J. Pratt
RF Micro Devices, Inc.

A monolithic amplifier using GaAs HBT technology has been developed which provides extremely flat frequency response and high dynamic range. This inexpensive amplifier is packaged in a standard SOIC-8 plastic package and has a flat frequency response to beyond 1 GHz (–3dB at 2 GHz), and an IP_3 of +40 dBm. It is capable amplifying multicarrier signals with very high fidelity. The primary applications of the device are in cellular base stations, CATV signal distribution, and other uses requiring excellent linearity.

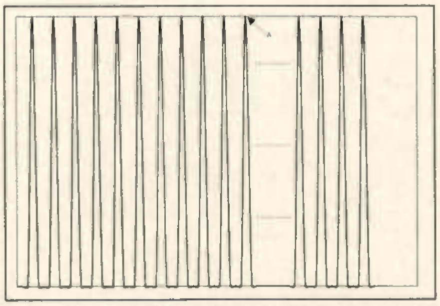

Figure 1. "Picket fence" spectrum of *n* CW carriers.

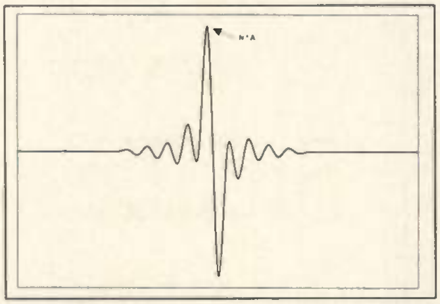

Figure 2. Time waveform of coherent CW carriers.

The need for high linearity amplifiers arises from stress placed on communications channels by the addition of more data and the requirement to handle digitally modulated signals with high fidelity. As the amount of data increases and the necessity to reduce spurious interference increases, the channel linearity must be improved over current implementations.

A particularly difficult problem is obtaining the linearity and bandwidth needed for the amplification of multicarrier signals such as those found in the TV distribution industry. Cable TV, for example, is expanding rapidly, requiring more bandwidth as more channels are added to the system. These systems are also incorporating new technology, such as two-way channels for voice and data, and digital video modulation. As this additional capacity is added, the bandwidth and signal handling capability of the various links in the distribution chain must be improved.

In addition to the need for more capacity, the linearity of the channel must further improve because digital modulation schemes are less tolerant of IM distortion products than the analog systems currently in use. The infrastructure needed to accommodate these technology innovations must be in place throughout the distribution chain before it can become available to the consumer. Hence, there is considerable pressure to improve channel linearity and bandwidth in the many links of the various distribution channels.

RF Micro Devices (RFMD) is developing a line of high linearity, wide bandwidth products to address these requirements. The first in the series is the RF2312 amplifier which has been released as a standard product. Other higher power units are expected to be released in 1996. The RF2312 is essentially a low cost, traditional Darlington "gain block" circuit with the exception that performance far outstrips previously available units. The gain is essentially flat over the bandwidth from 1 MHz to well over 1 GHz and it has an IP_3 of over 40 dBm. The secret to the performance of the unit is the use of GaAs Heterojunction technology in the fabrication of the device. The unit will replace more expensive and less reliable discrete amplifiers and permit much better distortion levels for a given power level.

Characteristics of Multicarrier Channels

A multicarrier channel consists of many independent RF carriers each with its own frequency and modulation. It is important that each of these signals remain "uncontaminated" by all of the other signals in the channel. For simplicity, consider that if *n* CW carriers are present, have the same amplitude, and are coherent with each other, the resulting frequency spectrum is a "picket fence" as shown in Figure 1. The time waveform resembles an impulse doublet function as shown in Figure 2. The amplitude of the voltage peak is $20\log(n)$ dB above the amplitude peak of a single tone. If, however, each carrier is independent and has a random phase with respect the other carriers, the time waveform is similar to that shown in Figure 3 while the frequency spectrum remains the same as in Figure 1. In some newer systems, this spectrum can extend from a few MHz to about 1 GHz. The result is a noise-like signal which has the same average power as in the coherent case, but with much smaller peak excursions. This is an important fact, since the amplifier has to be designed to handle peak excursions in a linear fashion. Designing the part for voltage excursions of $20\log(n)$ dB above the single tone amplitude would be very difficult (and would be design "overkill").

Because the signal is noise-like, the peak excursions of the voltage waveform are a statistical function. Therefore, the maximum level is indeterminate except to say that it is smaller than $A*20\log(n)$, where A is the amplitude of each carrier and *n* is the number of signals. In fact, successive measurements may yield different answers. Because of the nature and quantity of the signals present, howev-

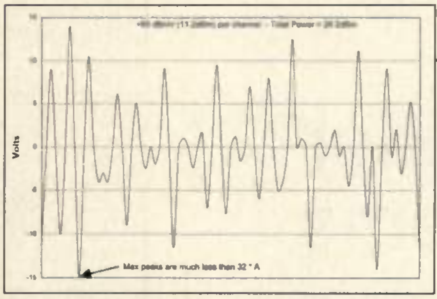

Figure 3. Portion of a 32-channel waveform.

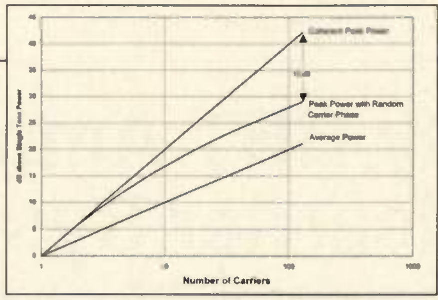

Figure 4. Peak power in multi-carrier channels.

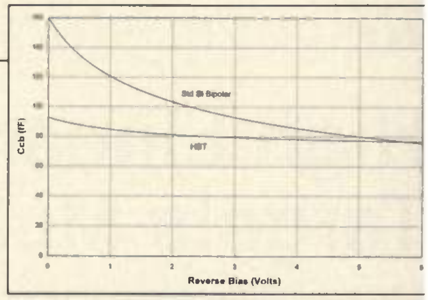

Figure 5. C_{cb} in HBT and Si bipola transistors.

er, it is possible to estimate the max mum excursion of the waveforms Many measurements have shown tha as the number of signals increases, th actual peak excursion is alway reduced markedly from the theoretica maximum. The curve of Figure shows the maximum expected pea power level for CW signals with ran dom phase along with theoretical max imum and average power level. Not that for 64 carriers, the differenc between coherent maximum an expected maximum is about 10 dB Therefore, the design of an amplifie which is to handle 64 signals mus have a linear signal handling abilit 26 dB above that needed for a singl signal, but the average output powe will be only 18 dB above the power i a single tone.

Causes of Distortion

Four different sources of distortio in an amplifier are listed below. Thes issues must be addressed either in th design of the amplifier or in the choic of technology used for implementation

Voltage compliance — As seen i Figure 4, an amplifier which has t handle 110 signals must be capable o an output voltage swing of about 2 dB higher than the peak voltage o each carrier signal and 31 dB highe than the RMS voltage of each signal So, if the output is set to +35 dBm (the RMS voltage of a single signal i 35 dB above 1 mV = 56.2 mV_{RMS}), th output swing must be at least 6 dBmV which is a 4 V_{p-p} voltage swing If the amplifier saturates during thi voltage swing, severe distortio results. In practical circuits, there i an emitter degeneration resisto which raises emitter voltage causin early saturation. It is also necessary t guard band the saturation of the out put device, since approaching V_{sat} wil cause distortion as well. So the DC V_{CE} (V_{DS} for FETs) used for the design must be at least 5.5 V for this applica tion and V_{CE} maximum must be a least 9.5 V. In general, high transisto

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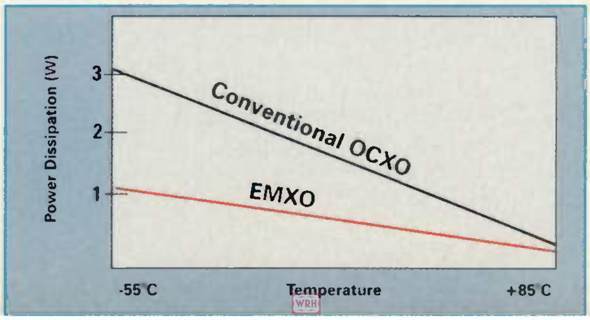

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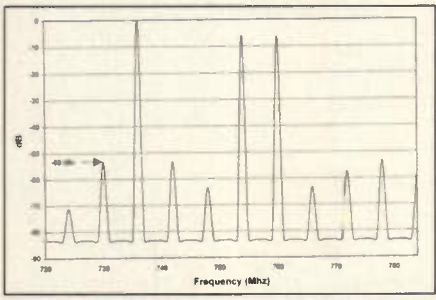


Figure 6. Distortion in a HBT circuit with a three-tone input.

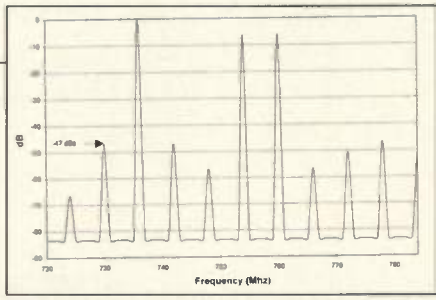


Figure 7. Distortion in a Si bipolar circuit (C_{bc} modulation).

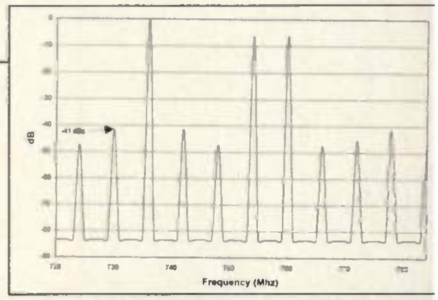


Figure 8. Distortion in a GaAs MESFET circuit.

breakdown voltages are needed for these applications in order to prevent voltage compliance distortion.

Drive Current — Although the exponential I-V characteristic of bipolar devices is thought to make the device nonlinear, experimental results and analysis prove otherwise. The use of emitter degeneration, feedback and ample DC collector current permit the device to obtain arbitrarily good linearity of G_m (transconductance). It is necessary to bias the device at a high enough current to achieve suitably linear operation. In general, devices used in this application have more current drive than would theoretically be necessary to drive the load at the required power level. The excess current is used to provide superior linearity over amplifiers designed for single carrier applications.

Distortion due to non linearity of base-collector capacitance — Because devices used for multicarrier systems are, of necessity, fairly large devices (because of power density), the base-collector capacitance may be larger than other amplifiers of the same power level. Ideally, the transistor's capacitance is constant, and therefore may cause roll off in frequency response, but not contribute distortion. Real world devices, however, exhibit a nonlinear capacitance vs. voltage curve which contributes significantly to distortion in the output signal, particularly when the voltage swing is quite large.

Non linear R_{out} of the transistor — Non linear loading of the output by the amplifying device itself also causes considerable distortion in the output waveform. Ideally, R_{out} is high and constant. If it is not, the loading of the output changes with signal voltage level, in effect modulating the output signal and causing distortion.

Advantages of HBT

The RF2312 is fabricated with a GaAs HBT technology, a proven technology which uses a GaAs/AlGaAs heterostructure for producing bipolar devices which have very high f_T, very high Early Voltage, high BV_{CEO} and a capacitance vs. voltage curve that is nearly flat. This HBT process is the most reliable commercially available HBT process in the world and has been qualified for class S space applications. This level of ruggedness is absolutely needed for space applications, but it is also demanded by commercial applications such as TV distribution systems and cellular base stations.

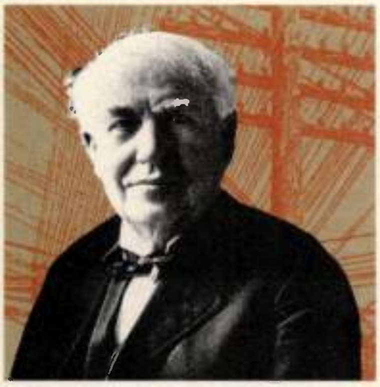
RFMD has supplied high volumes of power amplifiers utilizing this process with excellent results. The HBT process and products built with this process have been tested to determine that failure rates are at least 10^7 hours at 125 degrees C.

The main reason that this technology was selected for this product is that it provides superior linearity and bandwidth compared to other technology choices. The desired performance requires amplifying devices which have a very high f_T. The minimum f_T is about 15 GHz, with 20 GHz or more is desirable in order to obtain a really flat response through 1 GHz. Silicon devices could not be used for this application because, as the f_T of silicon bipolar monolithic transistors increases, a heavy price is paid in terms of breakdown voltage. The best Si BJT devices (for ICs) which have an f_T in excess of 15 GHz, have a breakdown voltage of 12 V (BV_{CBO}) and higher f_T devices result in even less BV_{CBO}. So, there is a limit to the voltage swing on the output which is too low for any multicarrier applications. GaAs devices (both HBT and MESFET) can have breakdown voltages which are much higher (18 V for MESFET and over 22 V for the HBT) even though they have F_T well over 20 GHz.

For the HBT, whose f_T is over 25 GHz, the DC collector voltage can be as high as 10 V and the resultant voltage swing is then 20 $V_{p\text{-}p}$. This makes the HBT the best candidate to provide the best voltage compliance for a wideband amplifier. The second very real advantage of the HBT is the fact that base-collector capacitance is small and much more constant than either the Si BJT or the MESFET. Figure 5 shows the capacitance variation vs. voltage for a Si BJT and an HBT. Note that the HBT capacitance is nearly constant with collector voltage. This has a significant effect on linearity as will be shown later.

RF Micro Devices has approached the marketplace for RF products using a philosophy of Optimum Technology Matching (OTM) which uses the best technology at our disposal (GaAs MESFET, silicon bipolar, and GaAs HBT) to address the needs of the market. In other words, we don't restrict the selection of the technology utilized, a priori, but instead, seek that which will yield the most effective solution for our customers. In this case, the overriding concern is to maximize linearity with minimum power consumption, and of course, do so at a reasonable price. Over the years, our experience is that the most linear amplifiers we can build are invariably GaAs HBT devices. In order to illustrate why we selected HBT for ultra linear amplifier products, three different parts have been simulated for comparison — an HBT part, a Si BJT part, and a GaAs MESFET part. The modeling used to simulate these devices is very accurate and has been tested in practice for many types of circuit. One of the parts is the RF2312, itself, hence, we have laboratory confirmation of the accuracy of this simulation.

The first circuit is the RF2312 using HBT devices. Its performance is shown in Figure 6. This spectrum is the output of the RF2312 with a three tone input. The test tones are those used for DIN4500B testing which is a standard test used for European TV systems. The large tone is at a frequency of 736 MHz. The smaller tones are each 6 dB below the large tone, and the frequency offsets are 18 and 24 MHz, respectively. Normally, the DIN rating is the highest level for the large signal which produces distortion products which are –60 dBc. As seen



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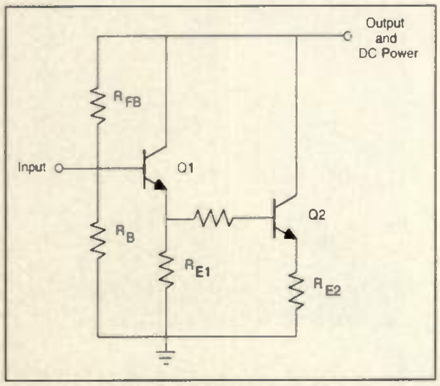

Figure 9. Diagram of a Darlington "gain block" amplifier.

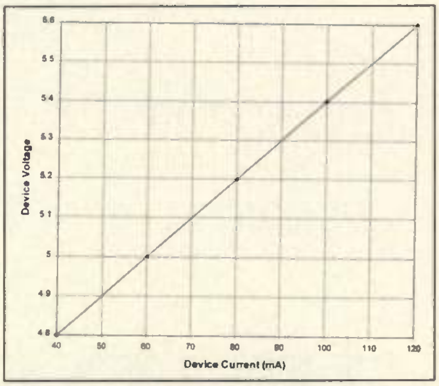

Figure 10. Device voltage versus device current for the RF2312.

in Figure 6, the RF2312 distortion is –53 dBc for the +60.5 dBmV level used for the test. In order to obtain –60 dBc, the level of the large signal must be decreased to about 57 dBmV. Since we are only interested in comparative data, the signal is left at +60.5 dBmV.

The spectrum of Figure 7 is the output of the same circuit except that silicon bipolar transistors are used. The transistors are actually "fake" transistors, in that they have identical properties as the HBT circuit except for the base-collector capacitance which is set to have capacitance vs. voltage variations like the one shown in Figure 5. The bias conditions, both current and voltage are identical for the two simulations. This silicon circuit is a very optimistic case since no silicon BJT IC process exists which has the combination of Early Voltage, f_T, and B_{VCEO} which is associated with the HBT device. But, if one could find a silicon process with these attributes the result would be as shown. Note that the distortion products increase by approximately 6 dB, but only because the base-collector capacitance varies with voltage. The degradation in performance shows clearly the advantages of the HBT over any process which has a significant variation in capacitance vs. voltage.

Figure 8 shows the same circuit using MESFET devices for the active elements of the design. Again, the bias conditions are identical. In addition, the gate width and pinch off voltage for these FETs were optimized for best performance. This circuit is not "fake" in the previous sense, because it is possible to physically realize the circuit. The linearity of this circuit suffers because of varying capacitance as in the silicon device, and also because of the strongly non linear output conductance of MESFETs. In addition, because of the lower G_m for the MESFET, no source degeneration is possible which hurts distortion considerably. In fact, it proved impossible to obtain the same gain as the HBT and Si BJT implementations, so the input signals had to be increased to get the same output power. The results are clearly inferior to both the Si bipolar design and the HBT. So, while it is possible to obtain MESFET devices which have sufficient breakdown and

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Gain Flatness +/-0.3dB (.05 to 1 GHz)
Bandwidth (3 dB) 2.5Ghz
Noise Figure: 3.8 dB
PMAX +21 dBm, @ 500 MHz
OIP3 +38 dBm @ 500 MHz
OIP2 57 dBm
Input VSWR 1.4:1
Output VSWR 1.6:1
ICC 100 mA
Vd 5.4 V

Table 1. Typical performance of the RF2312.

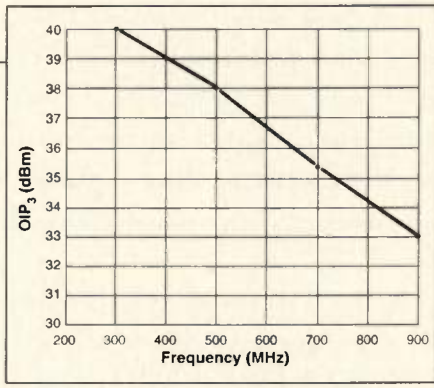

Figure 11. Plot of OIP_3 versus frequency

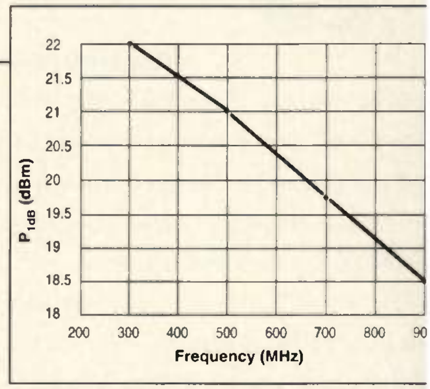

Figure 12. Performance of P_{1dB} versus frequency.

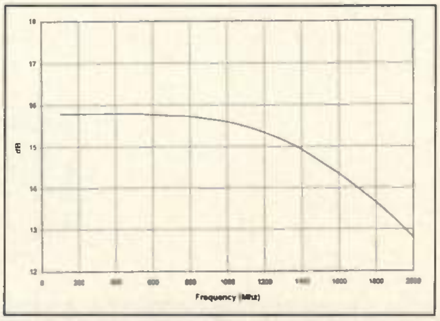

Figure 13. RF2312 gain versus frequency performance.

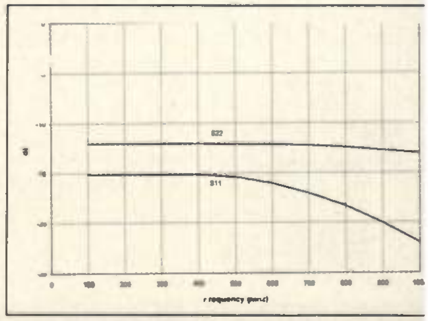

Figure 14. Input and output matcl (S_{11} and S_{22}) versus frequency.

f_T, they are not a good choice for high linearity amplifiers.

Chip Design

The RF2312 is a simple Darlington "cascadable gain block" design as shown in Figure 9. This is a broadband feedback amplifier which has its inputs and outputs matched to 75 ohms by virtue of the feedback resistor. Gain is determined by the feedback resistor (R_{FB}) and the emitter resistor (R_{E2}). These resistors, along with R_B, also set the DC bias conditions. The output is connected to the power supply through an inductor (RF choke) and a series resistor which sets the device current. The value of the resistor is determined by calculating the necessary voltage drop at the desired current level. The voltage drop is the difference between the power supply voltage and the device voltage given in Figure 10. For example, if the power supply voltage is to be 10 V and the device current is desired to be 100 mA, we see that the device voltage is 5.4 V and the drop across the external resistor must be 10–5.4 = 4.6 V. The current is 100 mA, so R = 4.6/.1 = 46 ohms. The desired bias conditions may be set up for power supply voltages from about 7 to 12 V. Voltages outside of that range are not recommended.

The inductor value is set such that the reactance at the lowest frequency of operation is at least 200 ohms. The input and output coupling capacitors are set such that their reactance at the lowest frequency of operation is less than 10 ohms.

Chip Performance

A summary of the measured performance of the RF2312 is shown in Table 1. Note that the table reflects measurements done in a 50 ohm system. This is because we were equipped for only 50 ohm measurements of some parameters. For consistency, all of the performance measures are listed as 50 ohm measurements. To perform all of the measurements at 75 ohms, a 50 to 75 ohm transformer must be connected to both the input and output ports. Unfortunately, these transformers have some loss and don't do a good job over the full bandwidth of the part. Because the imaginary parts of Z_{IN} and Z_{OUT} are very low, the VSWR is reasonably good at both 50 ohms and 75 ohms, so 50 ohm measurements were performed for Table 1 data.

A concern with any wideband amplifier of this sort is its stability, particularly when driving filters and severely mismatched loads. The RF2312 has been tested under a variety of conditions using various bandpass and lowpass filters without evidence of any oscillations. The parts has also been driven and loaded simultaneously with 10:1 tuners and showed no unstable properties.

Figures 11 and 12 show the Output Third Order Intercept Point (OIP_3) and power output at the 1 dB gain compression point (P_{1dB}) of the unit vs. frequency. Note that power handling ability does decrease as frequency increases. This is inherent in the Darlington configuration. It is the result of Beta decreasing in the second transistor (Q2) which forces the emitter follower stage (Q1) to provide more current at higher frequencies. A lower f_T, as may be found in a Si device, aggravates that situation; i.e. the higher the f_T of the devices, the bettei Distortion resulting from this effec can be decreased by increasing the col lector current of the first stage (lowe the value of R_{E1}). This, of course increases overall power consumptior The current level in the RF2312 is se to achieve the performance goals fo the part, and yield good multicarrie performance up to 1 GHz.

The frequency response is shown i Figure 13. Note the flatness in the GHz band — excellent frequenc response unmatched by other ampli fiers. This response was measured o a test board and includes the effect o the package, PCB and connectors Finally, Figure 14 shows the retur losses for both input and output.

Conclusion

The RF2312 offers the highest lin earity and best frequency response o any commercially available integrate circuit in its power class. The OIP_3 i +40 dBm and the frequency respons is extremely flat across a full GHz o bandwidth. The part operates from single power supply from 7 to 12 V Pricing is less than $4.00.

For more information, contact RI Micro Devices, 7341-D West Firnedl Avenue, Greensboro, NC 27410; tel (910) 855-8085; fax: (910) 299-9809 — or circle Info/Card #250. RI

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TUF DUCK CENTURION

INFO/CARD 41

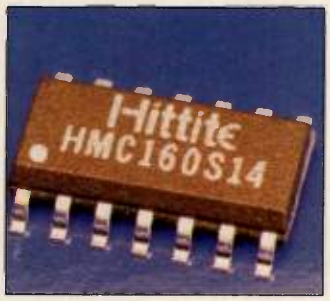
IC is Single-Chip DSSS Radio System for RFID

Micron Communications, Inc. announces the development of the RFIDEngine™ integrated circuit prototype. The IC is a single-chip spread spectrum radio frequency identification (RFID) circuit. The Micron RFIDEngine integrates a direct sequence spread spectrum (DSSS) microwave radio, a microcontroller, and low power static RAM (SRAM) on one silicon chip. This high level of integration should provide cost and performance advantages over current technology. The IC uses Micron's proprietary CMOS fabrication process. The development of the RFICEngine will lead to test marketing by Micron of very small, low-cost RFID tags under the MicroStamp™ brand name. The device will also be available on an OEM basis to other interested RFIC manufacturers. This announcement reports an advance in technology; RFIDEngine products are under development and will be available at a later date.

Micron Communications, Inc., a subsidiary of Micron Technology, Inc.
INFO/CARD #209

Diversity Switch for T/R Applications to 2 GHz

Hittite Microwave Corporation introduces the HMC160S14 low-cost diversity switch for use in transmit-receive applications, provided in a 14-lead SOIC package. The device can handle up to 3 watts power at frequencies up to 2 GHz, and is especially suited for 900 MHz operation. The switch is used to exchange two antennas between the transmitter and two receivers, providing antenna diversity. The switch has exceptional intermodulation performance: 54 dBm third order intercept at +8 volts bias. On-chip circuitry allows single positive supply operation (3 to 8 volts) at very low DC current. Control inputs are compatible with CMOS and most TTL logic families. Price is \$9.15 (100-999 qty.)



Hittite Microwave Corp.
INFO/CARD #210

Ultra Low Noise Oscillator Offers Lower Cost

The Ultra Low Noise series oscillators from Wenzel Associates, Inc. provides high performance at \$1400 in quantities of 10 (typical), versus more than \$1500 for the standard. The Ultra Low Noise series features excellent phase noise performance of better than –140 dBc at 10 Hz offset, and greater than –175 dBc at 1 kHz offset, for the 5 MHz SC-cut model. Temperature stability is better than $\pm 5 \times 10^{-9}$ from 0 to +50°C. Package size is 1.75 × 2.94 × 1 inch. Fast six-week delivery (part number 500-04903). Additionalinformation is available at the company's Web site: http://www.wenzel.com

Wenzel Associates
INFO/CARD #211

7/16 DIN Cable Assemblies Target Base Stations

A new series of LMR® cable assemblies with 7/16 DIN connectors is now available from Times Microwave Systems. Designed for both indoor and outdoor installation, the assemblies are fabricated with LMR-600 or LMR-900 cable. The connectors have gasketed seals for compete weatherproofing, and strain relief is provided by an adhesive-backed shrink boot. Cables are swept for VSWR and loss from 800-2000 MHz. A typical 3-foot LMR-600 assembly has 1.25:1 maximum VSWR, 0.3 dB maximum insertion loss, and a price of \$103.75.

Times Microwave Systems
INFO/CARD #212

3-V Power Amplifiers for Mobile Phone Systems

Oki Semiconductor's new low-cost, high-performance 3-volt GaAs power amplifiers target the high volume Advanced Mobile Phone Systems (AMPS) market. Designated the KGF1607 and KGF1637, these new efficient devices operate at frequencies up to 850 MHz. The devices are suited for cellular radios, wireless modems, LANs, high-efficiency portable computers with datacom LANs, cable TV systems, portable satellite terminals and wireless cable TV distribution systems. Both devices provide >31.5 dBm; the KGF1607 with an effiency of >70 percent, and the plastic-packaged KGF1637 at >60 percent. A 1000+ pieces, the KGF1607 is priced at \$8.76, and the KGF1637 is \$7.60.

Oki Semiconductor
INFO/CARD #213

Product Focus: Communications Test Equipment

One-Box Tester for Pagers

Hewlett-Packard announces the Pager Signaling Option 1EP for the HP 8648A signal generator. This option allows testing of pagers to the new FLEX and FLEX-TD paging standards from Motorola, and the POCSAG protocol. The 1EP includes an internal digital pager encoder, and another option, 1E5, offers a high-stability timebase that meets stringent FLEX requirements. The HP 8648A with Option 1EP is priced at \$5865; Option 1EP alone is \$1435.

Hewlett-Packard Company
INFO/CARD #214

RF/Microwave Test Set

Marconi introduces the 6200B series of RF and microwave test sets that integrate a scalar network analyzer, synthesized sweep generator, power meter, frequency counter, plus voltage and current sources. Fault location software allows you to measure return loss on coaxial or waveguide transmission lines with 0.1% accuracy. Five versions of the instruments exist, with frequency ranges of 10 MHz to a maximum of 2, 8, 20, 26.5 or 46 GHz. Power sensors, bridges and scalar detectors allow a wide range of transmission and reflection measurements, including time domain.
Marconi Instruments
INFO/CARD #215

Modulation Spectrum Analyzer

Models R3263 and R3465 from Advantest are spectrum analyzers with multi-standard modulation analysis capability. The R3465 covers 9 kHz to 8 GHz, and addresses the North American Digital Cellular (NADC), Personal Digital Cellular (PDC) and Personal Handy phone System (PHS) standards. Model R3263 has a 9 kHz to 3 GHz frequency range and covers the standards for the Global System for Mobile communications (GSM). The R3465 is \$32,995; the R3263 is \$27,000 with the GSM Tx Plus option.
Advantest Corporation,
distributed by Tektronix, Inc.
INFO/CARD #216

Mobile Station Interface Set for TDMA and CDMA

The Noise ComMS-800/MS-1900 Series instruments include all the necessary RF components, such as attenuators, switches, duplexers, and two synthesized CW signal generators or $\pi/4$ DQPSK signal generators to perform integrated testing of CDMA and TDMA. The units are designed to be used with the Wireless Telecom Group's MP-2500 multipath simulator, and UFX-BER-892 Carrier-to-Noise generator, for a complete mobile test setup. The MS-800/MS-1900 Series prices start at \$27,500 with delivery in 2-10 weeks.

Noise Com, Inc.
INFO/CARD #217

DISCRETE COMPONENTS

High Permeability Toroids

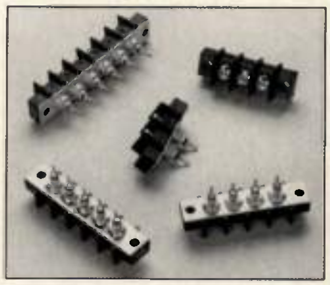
Ceramic Magnetics has achieved a 50% increase in the permeability of ferrite toroids. Their Peak-Perm™ toroids are produce using a proprietary material which holds 15,000 perm at 10 kHz at room temperature. The toroids are available in diameters from 0.100 to 0.50 inches O.Dd, and can be produced in thicknesses from .030 to .250 inches.
Ceramic Magnetics
INFO/CARD #218

Ultrasonics Starter Kit for Experimenters

Valpey-Fisher's Ultrasound and Optical Products offers an easily accessible kit of piezoelectric crystals which can be used for classroom instruction, laboratory research, or proof of concept for new product ideas. The kit includes crystals ranging from quartz and lithium niobate to various grades of PZT and lead metaniobate. Crystals are furnished with technical data, and have solderable electrodes. Price of the Ultrasonics Starter Kit is \$195.
Valpey-Fisher Corporation
INFO/CARD #219

Filtered Terminal Blocks

EMI-filtered terminal blocks from TUSONIX offer capacitance ranging from 2500 to 5000 pF, insertion loss of 50 dB at 100 MHz, and a terminal torque resistance of 9 in-lbs. The filtered terminal blocks are designed for easy attachment, and acan be custom built for specific applications.



TUSONIX, Inc.
INFO/CARD #220

Crystal for Communications

Connor-Winfield has developed a new crystal, Part no. CW49-GL, designed for use with the Motorola MC145572 transceiver IC. the crystal is available in frequencies of 20.0 through 28.0 MHz, with center frequency stabilities from ±25 ppm to ±100 ppm. Standard frequencies of 20.480 and 25.088 MHz are available from stock. Prototype price is $12.50 each in quanties of 10.
Connor-Winfield Corp.
INFO/CARD #221

Ceramic Resonator Available to 20 MHz

AVX has introduced an all-ceramic resonator with an extended frequency range up to 20.00 MHz, providing a surface-mount alternative to clock oscillators in amny applications. The PBRC series has a low profile (2 mm), and can withstand nearly all soldering processes. They are available in frequencies from 8.01 to 20.00 MHz (PBRC-A) or 2.00 to 8.00 MHZ (PBRC-B). In quantities of 10,000 the prices of the PBRC-B series is $0.35.
AVX Corporation
INFO/CARD #222

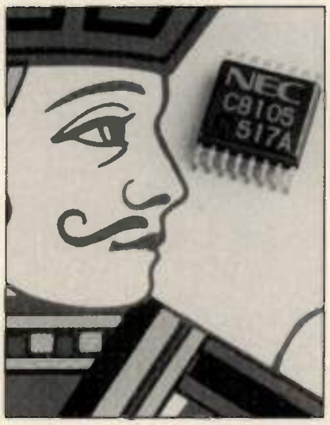
SEMICONDUCTORS

Low-Distortion Voltage-Feedback Op Amps

Maxim Integrated Products' MAX4108 operational amplifier features 400 MHz 3 dB bandwidth and ±5 V operation with a 1200 V/µs slew rate and a spurious-free dynamic range (SFDR) of 93 dB at 5 MHz. The MAX4109 is compensated for a closed-loop gain of 2 V/V or greater and provides 225 MHz –3 dB bandwidth and a SFDR of 90 dB at 5 MHz. Applications include imaging, instrumentation and RF signal processing. Pricing starts at $3.88 (1000 qty.)
Maxim Integrated Products
INFO/CARD #223

MMIC Quadrature Modulator

California Eastern Laboratories has introduced another NEC silicon MMIC device for wireless applications. UPC8105GR is a quadrature



modulator with an on-chip phase shifter, 400 MHz modulator bandwidth, 10 MHz I-Q bandwidth and a *Power Save* function. Supply voltage is 2.7-5.5 volts, and the device is provided in a 16-pin SSOP package. Current consumption is 16 mA at 3 volts. The UPC8105GR is fabricated using the NESAT III process with transistors that approach 20 GHz FT. Pricing is $3.90 in production quantities.
California Eastern Laboratories
INFO/CARD #224

High-Performance Phase/Frequency Detectors

R.F. Research Australia announces the BK7001 (single) and BK7002 (dual) phase/frequency detectors. Their proprietary design produces lock-up times down to half that of conventional detectors. Included is a true lock detect capable of driving a lock alarm LED directly. Inputs and outputs are TTL or HCMOS compatible, and the maximum operating frequency is 35 MHz. Packaging is a 44-pin J-lead.
R.F. Research Australia
INFO/CARD #225

SMT Power MOSFET

Motorola announces the MRF5007, an RF power MOSFET for broadband applications at frequencies up to 520 MHz. The device supplies up to 7 watts ouput power at 7.5 volts with a minimum gain of 10 dB at 512 MHz. Other features include low feedback capacitance (C_{rss} = 13 pF) and the ability to withstand 20:1 load VSWR at any phase angle. The MRF is packaged in a true surface-mount package which can be handled by automatic assembly equipment. Pricing is $16.20 in low volumes.
Motorola SPS
INFO/CARD #226

GaAs IC for GSM

Anadigics Inc. has introduced a new GaAs power amplifier for use with the GSM telephone standard. The AWT0904 has three stages, providing 32 dBm output power with 0 dBm input. Power control during the transmit pulse, low noise in the receive band, and a heat-sinked SSOP28 package are additional features of the device. A similar device for DCS1800 and PCS1900 will sooon be available.
Anadigics, Inc.
INFO/CARD #227

SIGNAL SOURCES

Low Cost TCXOs

Networks International Corp. introduces low cost temperature controlled crystal oscillators (TCXOs) with a standard range of 9-25 MHz and ±1 ppm stability over 0 to 50°C. Output is 1.0 Vp-p, clipped sine DC cut. The TCXOs are provided in a .740 × .460 × .340 package.
Networks International Corp.
INFO/CARD #228

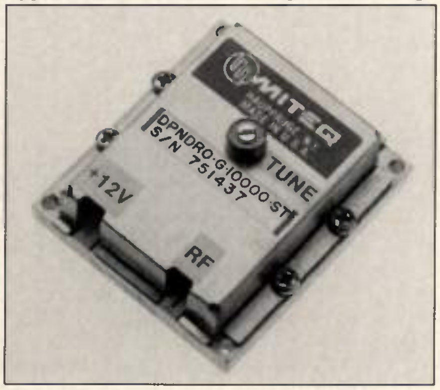
SONET Clock Oscillator

MDR3001-2 from Magnum Microwave is a dielectric resonator oscillator for a center frequency of 9953.28 MHz, the line rate frequency for SONET OC-192 lightwave communications. It is electrically and mechanically tunable and can be phase

locked for use in clock recovery and clock regeneration circuits. Free-running stability is 3 ppm/°C, power output is +11, ±2 dBm, and typical phase noise is –110 dBc at 100 kHz offset.
Magnum Microwave
INFO/CARD #229

Drop-in DRO

Miteq's DPNDRO series of drop-in dielectric resonator oscillators is available with SMA connectors or direct microstrip launch for wireless communications and other applications. Available frequencies range



from 4.5 to 15 GHz, with power output of up to +10 dBm. Mechanical tuning is provided, and a built-in voltage regulator and reverse voltage protection are standard features. Electronic tuning and low-noise versions are available.
Miteq
INFO/CARD #230

Synthesizer Products

Syntest announces a complete line of low-cost stand-alone synthesizers and modules. The SM/SI series are available in a range of 0.1000 Hz to 159.999 MHz emloying single phase-locked loop technology. Temperature stability is guaranteed to ±10 ppm (±1 ppm optional) over 0-50°C. Output levels are avilable in TTL or ECL. Frequency tuning is accomplished by TTL or ECL, thumbwheel switches, or remote BCD parallel lines. Prices start at $247.00 in OEM quantities.
Syntest Corporation
INFO/CARD #231

AMPLIFIERS

PCN/PCS Amplifiers

Wessex Electronics has introduced a series of PCN and PCS high power amplifiers, available in modular form, instrument case, or 19 inch rack mount. Model ANP-0135S is conservatively rated at 20 watts in either the 1805-1880 MHz or 1930-1990 MHz bands. Gain is 27 dB minimum, harmonics are –50 dBc and spurious outputs are –70 dBc. A full complement of protection and operational features are provided in these units.
Wessex Electronics
INFO/CARD #232

SIGNAL PROCESSING COMPONENTS

High Power Diode Switch

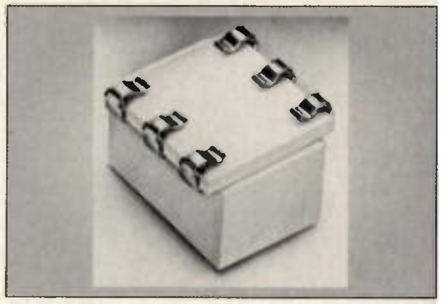
Hill Engineering offers the tiny series H22 SPDT switch that handles 2 kW peak power and 80 wats average power over 500-4000 MHz. Switching speed is 1 μs, insertion loss is 0.5 dB and isolation is 60 dB. Case size is 1.6 × 1.6 × 0.5 inches.
Hill Engineering
INFO/CARD #233

High Isolation Combiner

Model REC-9A2NA from Renaissance Electronics is a power combiner offering three in-phase inputs in the 800-900 MHz band. With 1.5 dB insertion loss, the combiner features 80 dB isolation between input ports, and ±5 degree phase balance. The standard product line includes couplers and power dividers for 800 to 1990 MHz, supplied in 2, 3, 4, 6, 8 and 16-way, configurations.
Renaissance Electronics
INFO/CARD #234

VHF/UHF 0° Power Splitter

Mini-Circuits introduces a 1-watt broadband 3-750 MHz 2-way power splitter with 28 dB isolation, 1.2:1 VSWR and insertion loss of 0.4 dB (typical specifications). The



JPS-2-1W is priced at $8.95 ea. (1-9 qty.) and is available from stock.
Mini-Circuits
INFO/CARD #235

Notes on Electrically Small Antennas

By Gary A. Breed
Editor

To meet size requirements, new wireless applications, as well as many current radio systems, use antennas which are smaller than typical self-resonant configurations. To better understand the performance tradeoffs of small antennas, this tutorial presents information on their impedance, matching methods, and loss mechanisms.

Small antennas can be defined as those with dimensions smaller than equivalent self-resonant types. Self-resonant configurations of antennas that might be used in common communications applications include:

- 1/4-wavelength monopoles, fed against a ground plane or counterpoise.
- 1/2-wavelength dipole
- 1-wavelength loop
- 1/2×1/2-wavelength square microstrip patch
- 0.6-wavelength diameter circular microstrip patch
- 1/2-wavelength slot in a conducting surface (or waveguide)

The graphs in Figures 1 show the relationship between frequency and wavelength for a 1/4-wave and 1/2-wave. Although these are trivial calculations, the graphs offer a quick review of the self-resonant dimensions of antennas at various frequencies.

For example, in the cellular band (800-900 MHz), a 1/4-wavelength monopole will be around 9 cm in length. This length may be practical for many fixed and portable phones, but the trend in manufacturing in toward smaller products. A manually-extendable antenna is included in some products, but with consumer convenience in mind, it may be desirable to conceal the antenna within the cellular phone structure, requiring a less-than-full-size antenna.

In the traditional paging band (around 160 MHz), the problem is quite different. A quarter-wave in this frequency range is 47 cm, much larger than anyone would want to hand on their belt or place in a shirt pocket! A comparison of size is shown in Figure 2, on the next page.

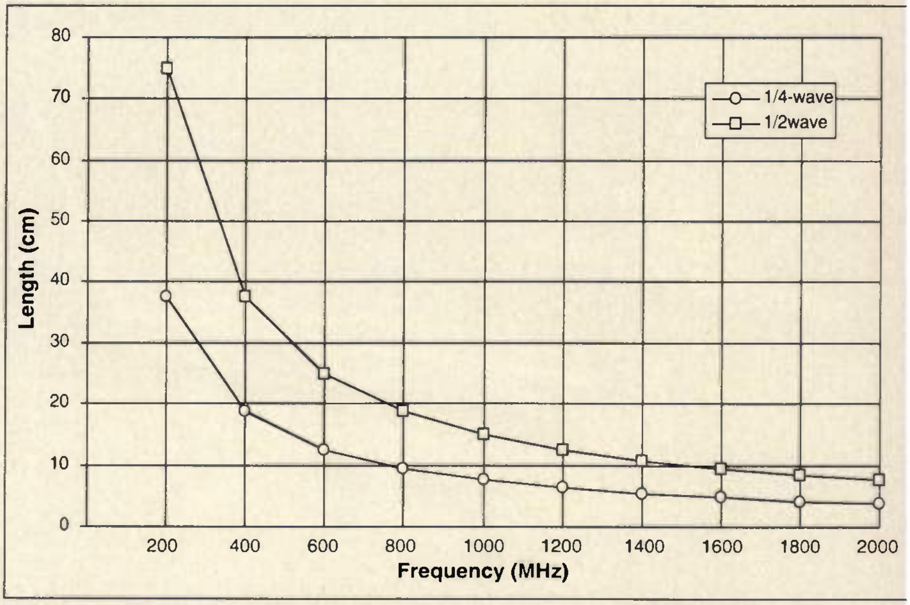


Figure 1. 1/4 and 1/2 wavelengths versus frequency; 200-2000 MHz.

Behavior of Small Antennas

Dipole and monopole antennas — At short lengths, the feedpoint impedance of dipoles (and monopoles with a proper counterpoise) shows increasing capacitance and decreasing resistance. This impedance must be transformed to the operating impedance of the transmitter and/or receiver circuitry in order for the antenna to be useful. Intuitively, this is logical, since a capacitor is two conductors separated by a dielectric space. As the conductors get shorter, the capacitance decreases and the capacitive reactance increases. The resistive behavior is not intuitive, since the antenna is already an open circuit, but it may be useful to think of the shorter length as representing a smaller value resistor.

Loop antennas — As the circumference of a loop antenna gets smaller compared to its self-resonance at one wavelength, the feedpoint impedance is inductive, with decreasing resistance. The inductive reactance becomes relatively high once the size falls out of the region near resonance and decreases as the size decreases. This is intuitively explained by looking at the loop as a one-turn inductor that is getting smaller (less inductance and lower reactance). Also intuitively the resistance of a smaller length conductor will decrease with size.

Matching to reactive loads is typically done in two steps: removing the reactance by adding the necessary opposite type, then transforming the resistance to the desired value. In the case of a dipole or monopole, the first step means adding inductance to cancel the capacitive reactance. For the loop, it means adding capacitance that has equal reactance as the inductive component of the impedance.

Losses in Antenna Systems

Small antennas do not radiate much less efficiency than their full-size counterparts — *if we have lossless con-*

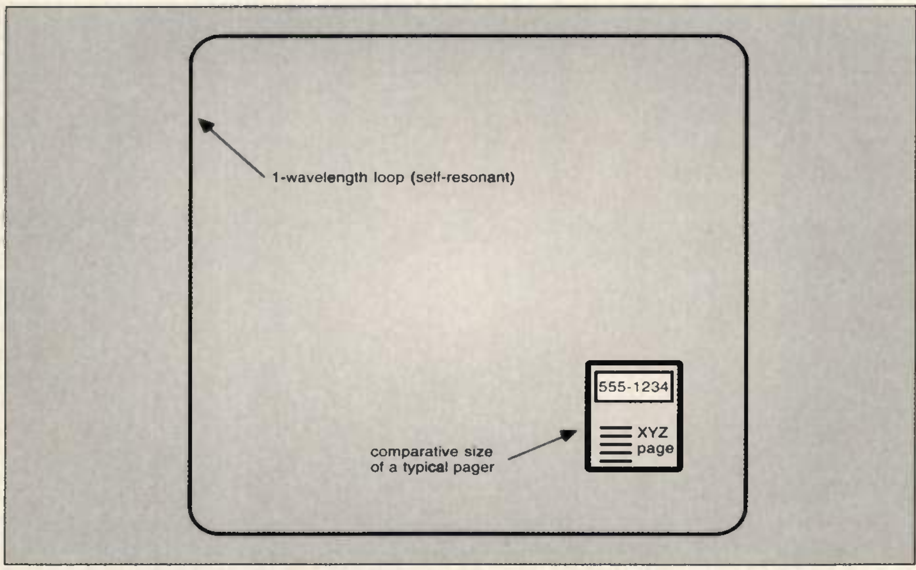


Figure 2. Some pagers use small loop antennas printed on the outside edges of the circuit board. These antennas will be no more than 1/8 the size of a 1-wavelength loop self-resonant at that frequency.

ditions. Theoretical losses occur from cancellation of fields, but are not large until the antenna is perhaps 1/10 of "normal" size.

Most losses arise from actual resistive losses in the matching circuitry. The main culprit is the inductor. Inductors have significant ohmic losses which become proportionally larger as the resistive impedance decreases. For example, an inductor with one ohm resistance will dissipate half the power delivered to an otherwise pe fectly-matched antenna with one oh resistive impedance.

In the case of the loop, the inducto is the antenna conductor, with greater degree of importance on ohmi losses than a full-size antenna. T help minimize losses, small loops ar often constructed from copper tubin with silver-soldered joints.

Components in the matching net work are another source of losses Very low resistance results in propo tionately higher currents flowing i the inductors, capacitors or transform ers used for matching. Random selec tion of a matching network is not rec ommended, despite the fact that a least 12 different schemes might pro vide the desired transformation. Whil a matching network cannot reduc losses in the antenna conductor or th required series inductors, selection c the right type can avoid excessiv additional losses.

Combating the Problem

The "brute force" approach is the us of more transmit power and mor receiver gain to overcome losses. Whil a valid approach, this solution ha limitations — regulation of transmi power, power consumption, and th noise figure of the receiver.

Enhancement of efficiency can b gained by coupling to the body of th user, or to metallic objects. A page antenna will be more efficient by a fac tor or 6 to 10 dB when placed on th user's body, even with our moderat conductivity. The user's hand ca enhance the counterpoise of a portabl cellular phone, compared to having th unit sit on a table, for example (safet concerns must be considered, howev er!) You may have noticed improve reception from a portable AM-FM radio when it is placed on an automo bile hood or trunk lid.

In all cases, the expected efficienc of the small antenna must be include in the calculation of receive and trans mit performance. Appropriate margin must be designed-in to allow reliabl communications. *RF*

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2. K. Fujimoto, J.R. James, *Mobile Antenna System Handbook*, Artech House, 1994.

What You Should Know Before Returning To School for that Advanced Degree

By. Jeffrey H. Reed, Theodore S. Rappaport and Brian D. Woerner
Virginia Tech

Nearly every engineer at some point in their career has pondered the idea of returning to school for additional education. Students whose four year degree is not in electrical engineering may even return to school and obtain a MS or PhD in electrical engineering, especially for those students with significant technical experience.

There are valid and erroneous motivations for returning to school. Valid motivations include:

1. Gaining more knowledge to satisfy intellectual curiosity
2. Gaining new skills to change to a new technical field. This is an especially important aspect for those engineers who are converting from a military oriented profession to a commercially oriented profession
3. Improving or expanding career opportunities to include teaching, consulting, or more research oriented work
4. Enjoying a culturally rich and diverse atmosphere

Erroneous motivations, or myths about the advantages of returning to school include:

1. Gaining a higher degree is a guaranteed way to earn more money
2. Returning to school is a solution to "job burn-out"
3. Obtaining a new or advanced degree is a "vacation" which requires little work

In this article, we explore the motivations for returning to school, examine educational alternatives, discuss some of the advantages and disadvantages of returning to school and provide advice on how to select the best school to fit your needs. Much of the perspective for this article has been furnished by students who left full-time jobs to return to school for advanced electrical engineering degrees at Virginia Tech's Mobile and Portable Radio Research Group (MPRG).

Mechanism for Further Education

There are four ways for an engineer to acquire further education:

1. Short courses
2. Part-time education while working full-time in industry
3. Full-time education while working part-time in industry
4. Full-time education while working at the university

Short courses are obviously the most convenient of these options in terms of minimizing impact on one's personal life. It is usually a cost effective way to learn since most employers will cover the expense of courses relevant to one's job. The drawback is that this type of learning can be very superficial. It is very difficult to retain information presented at the fast pace of a short course. The learning process requires time, concentration and a chance to wrestle with concepts so they become integrated into the problem solving thought process. Unfortunately, short courses offer little, if any, opportunity for the student to work problems or participate in design projects.

Part-time education offers a compromise between short courses and full-time education. Part-time education is less disruptive of personal life than returning to school full time and less financially draining. Drawbacks are that the quality of education is not as high as a full-time education since time is split between work and school. Coordinating time between education and a full-time job can be frustrating, particularly if the student is married and has a family. Part-time education is becoming increasingly important for keeping up in a technical field, but it requires patience and commitment to get a degree in a part-time program. Typically, a part-time MS student takes three to five times longer to obtain a degree as a full-time MS student.

Part-time PhD student status is not usually feasible, unless it is combined with an extended period of being a full-time PhD student. This combination can be a excellent strategy. Beginning school as a part-time student allows one to refresh math skills and complete a few classes while searching for an advisor, dissertation topic and preparing for any entrance exams. A financial advantage to part-time education is that many employers will reimburse the cost of work-related courses.

Full-time student status has many merits. Besides being a quicker mechanism for obtaining the degree than part-time status, the ability to focus exclusively on education allows for a better quality of education. Furthermore, one can enjoy the rich diversity of culture that college campus have to offer. New friends and associates made in school often times become important contacts for years to come.

Frustrations of Returning Engineers

For the engineer who returns to school as a full-time student, there are some potential surprises awaiting. The decision to return to school can often be met with skepticism from co-workers and management. Co-workers can often be discouraging, pointing out the loss of income and hard work in going back to school, as one returning MPRG student put it "my co-workers thought I was crazy." Management may not be supportive of this move as well. For instance, returning back to the same company after earning a degree may not lead to improved salary, especially if one returns to the same job title. A company can easily take for granted the benefits of the enhanced education of their employee.

Rusty math, technical skills, test taking, and study skills are the bane of the returning engineer. Most find that these skills return quickly, and are quite pleased to know that this information is not lost. The returning engineer does have an advantage, the ability to better discern what the important material is and what the relevant applications are to industry. This is a tremendous asset in studying for exams and working design projects, which have become a pervasive part of most technical graduate courses. One older returning graduate student in the MPRG describes his ability to compete with non-returning students this way, "Old age and guile will beat youth any day."

There are psychological barriers for returning to school. One MPRG student expressed his concern on returning to school as being viewed as the "old man on campus." This myth was quickly dispelled,

"Now the only time I feel like an old man is when I walk into a bar and find my self surrounded by people that are barely twenty one. I find that there are a lot of students who are like me, returning to school after working for some period of time. Quite a few of my friends are actually older than I am."

Another psychological barrier for returning students is in moving from an engineering position where one has control over resources and people to being powerless. A good electrical engineering department will respect the experience of returning engineers and the benefits they bring. As one MPRG student stated,

"I've noticed that the professors treat someone who has experience with a bit more respect. I think professors do genuinely value the opinion of someone who has a bit of real-world and real-life experience. Maybe I've just grown more assertive from being 'out there' and demand more attention in class than I used to (compared to my undergrad days)."

Probably the greatest barrier for returning students is the financial burden, especially for those with a family. It's difficult to live on a graduate student budget after acquiring a comfortable lifestyle. In the long term returning back to school usually offers a financial advantage. However, it takes several years to make up for the loss of income during the time spent in school. It would be unwise to return to school strictly with the expectation of financial gain.

Faculty advisors may give preferential funding to returning engineers because of the value of their experience. Student pay is generally quite low. Typically, research assistantship stipends (PhD level) run from $1300-$1600 per month and may or may not include tuition. Research assistantships are more likely to be given to returning engineers. Often universities will grant tuition wavers or fellowships to highly qualified students. For example, at Virginia Tech, the Bradley Fellowship program supports eight U.S. citizen PhD students in the electrical engineering department and provides a stipend of $1550 per month, full tuition, all book costs and a small housing allowance.

Picking the Right University

Picking the right university for full-time study can be a very difficult task Look especially hard at the strength o the particular department, emphase within the department and use thes as the most important criterion instead of the name recognition of th university. A strong department wil have adequate resources including fac ulty, equipment, lab and office spac and a wide selection of relevant class es. A strong research university wil require a lower number of teachin hours from its faculty, strong ties t industry and the faculty will have a lengthy publication record in leadin technical journals. Reviewing studen theses and dissertations is also a excellent way to judge the quality o school and its research program.

Meeting with potential researc advisors, instructors, and fellow stu dents is an effective way to evaluat the quality of the faculty. Ask toug questions of the faculty and adminis trators, e.g., "What is the percentag of students being funded thoug research contracts and teaching assist antships?", "Are tuition wavers grant ed?", "What is the minimum and maxi mum class size?", "What is the per

About the Mobile and Portable Radio Research Group (MPRG) at Virginia Tech

The MPRG was founded in 1990 by Dr. Ted Rappaport to conduct research on emerging wireless communication technologies. The group has research, teaching and service missions which are national in scope and is a leading producer of professionals for the wireless communication industry. MPRG consists of three principal faculty members, seven support staff and approximately thirty five graduate students, many of whom are engineers returning to school for advanced degrees. Major MPRG research thrusts include:

- Simulation and analysis of wireless communications systems using real-world channel models.
- Development of signal processing algorithms for interference rejection, antenna steering, error correction, and fast synchronization.
- Advanced modem development.
- Site-specific prediction of propagation characteristics.
- Measurement and analysis of propagation characteristics for wireless channels.

Core funding is provided by the MPRG Industrial Affiliates Foundation, a coalition of major corporations in the wireless field who serve as academic "stock holders" of the research group. MPRG Industrial Affiliates include: AT&T, Bell Communications Research (Bellcore), BellSouth, BNR, the Federal Bureau of Investigation (FBI), Grayson Electronics, GTE, MCI, Motorola, National Semiconductor, Southwestern Bell, and Texas Instruments.

The MPRG prepares quarterly research reports for its industrial affiliate members and promotes aggressive technology transfer through frequent visitation. Most industrial affiliate members have hired one or more MPRG graduates. In return, affiliate members provide financial support and serve as unofficial "stockholders" of the group, providing technical and managerial advice. The MPRG also conducts funded research for individual government or industry sponsors.

ɪntage of students who pass the PhD ıtrance exam?", "Are faculty evaluat- ı on their teaching skills?", and Vhat are the infrastructure resources ..g., computers, office space, lab ıuipment) available to graduate stu- ents?". These same questions should lso be posed to students attending ıe university.

Examine the faculty's attitude ıwards returning students. A quality ıgineering department will recognize ıe importance of returning engineer's ıills. Some universities, such as Vir- ınia Tech, allow especially qualified ıudents to serve as co-principle inves- gators on research contracts. A quali- ı advisor will encourage close and ɔntinuing contact of the students ith research sponsors. Some faculty rovide business cards to MS and PhD ıudents who are expected to eventual- ı manage their own research projects.

Its also advantageous to examine ıe relationship of the faculty with ıdustry. If a faculty member has a ɔod relationship with industry it is kely that their classes and research ısters skills that make graduate stu- ents attractive to employers. A uni- ersity's ties to industry are an excel- ınt indicator of how easy it will be to ɛt a job once a degree is obtained.

hould You Return to School?

Returning to school is a personal ecision based on the expectations of ıe educational experience. Higher ducation is an investment which does ot immediately lead to greater finan- al gains, but rather an investment in ıtellectual maturing and technical ophistication. We have seen few ɛturning students who have regretted ırthering their education, and believe ıat more engineers error in making ıe decision not to return to school.

cknowledgment

We wish to thank the returning ngineering students in the MPRG ho have provided valuable insight ıto this article. *RF*

About the Authors

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Reducing Active Filter Cost Using Symbolic Three-Pole Synthesis

By: Glenn A. Parker
Eagleware Corporation

Often in active filter design, the overall cost per unit is predominately determined by how many operational amplifiers the circuit contains. Since they are by far the most expensive element in common RC active filters, the ability to realize higher orders with fewer op-amps is an attractive option. However, symbolic solution of the transfer function for element values in terms of the desired poles can be a difficult task.

Active filter design has been limited to cascading first and second order sections to realize conjugate and real pole placements. This is because the symbolic solutions for third order filters are not documented. Williams [1] presents tables of capacitor values for the third order unity-gain lowpass section, but does not encourage attempts at solving the system of equations resulting from transfer function coefficients.

This article presents the equations necessary for designing third order, single op-amp filters. The transfer function is first shown for both the unity-gain, and the 6 dB gain (K = 2) case. The solutions are then presented along with a practical design example utilizing the three-pole filter, and contrasts single op-amp sensitivity with a typical two op-amp alternative. Highpass and lowpass solutions are presented.

Lowpass Transfer Function

The general voltage-gain transfer function for a third-order lowpass filter is:

$$T(s) = \frac{K}{As^3 + Bs^2 + Cs + 1} \qquad (1)$$

Where K determines the filter gain at DC ($\omega = 0$), and A, B, and C determine the transfer function pole placement and the filter characteristics.

Unity Gain Lowpass Filter

The schematic for a normalized unity-gain third order active lowpass filter is shown in Figure 1. All resistors are normalized to 1 ohm. The transfer function for this filter is given by Equation (1), where:

$K = 1$
$A = C_1C_2C_3$
$B = 2C_1C_2 + 2\,C_2C_3$
$C = C_1 + 3C_2$

By solving these equations for C_1, C_2, and C_3 in terms of the denominator coefficients, a symbolic expressio for the capacitors in terms of th desired poles is easily found. Taking $= \alpha \pm j\beta$ and $s = \sigma$ to be the roots of th denominator, the solutions are:

$$L = \alpha^2 + \beta^2$$

$$M = -\frac{4\alpha^2}{81L^2} - \frac{10\alpha + 9\sigma}{162\sigma L} - \frac{1}{81\sigma^2}$$

$$N = \frac{8\alpha^3}{729L^3} - \frac{\alpha(10\alpha + 9\alpha)}{486\sigma L^2} - \frac{5(2\alpha - 9\alpha)}{972\alpha^2 L} + \frac{1}{729\sigma^3}$$

$$C_2 = \sqrt[3]{N + \sqrt{M^3 + N^2}} + \sqrt[3]{N - \sqrt{M^3 + N^2}} - \frac{2(2\alpha\sigma + L}{9\sigma L}$$

$$C_1 = -3C_2 - \frac{2\alpha}{L} - \frac{1}{\sigma}$$

$$C_3 = 3C_2 + \frac{\alpha}{C_2\sigma L} + \frac{1}{2C_2L} + \frac{2\alpha}{L} + \frac{1}{\sigma}$$

Lowpass with Gain

The schematic shown in Figure has a lowpass frequency response wit 6 dB of gain. All capacitors are nor malized to 1 farad. The transfer func tion for this filter is given by Equatio 1, where:

$K = 2$
$A = R_1R_2R_3$
$B = 2R_1R_3 + R_2R_3$
$C = R_1 + R_3$

Solving for the normalized resistors we have:

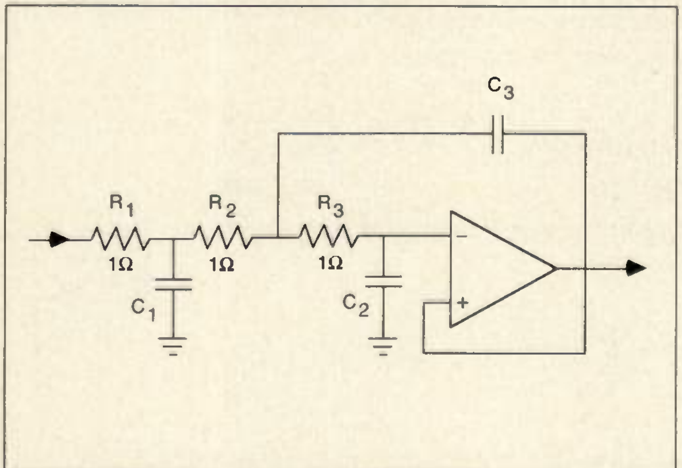

Figure 1. Schematic for a K=1 lowpass filter

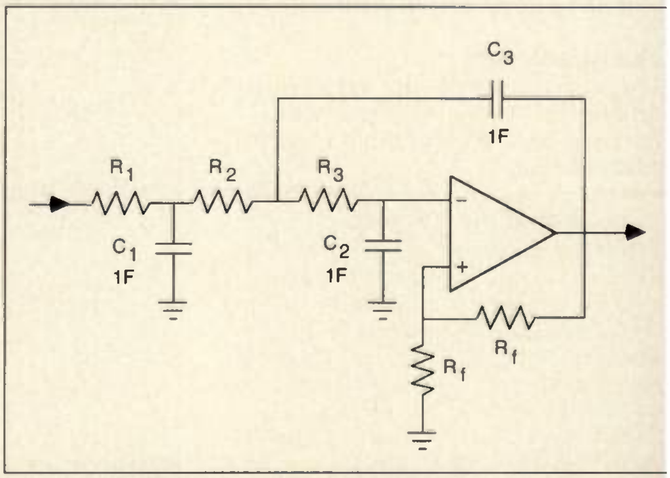

Figure 2. Schematic for a K=2 lowpass filter

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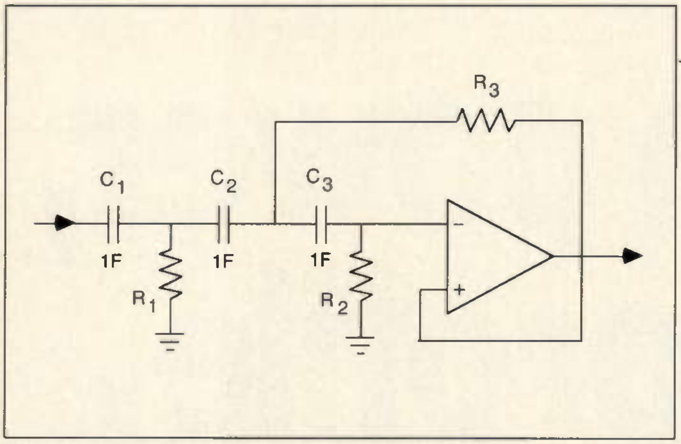


Figure 3. Schematic for a K=1 highpass filter.

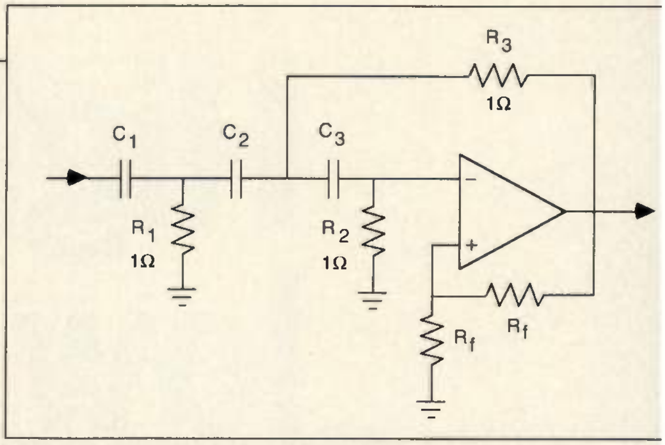


Figure 4. Schematic for a K=2 highpass filter.

$$L = \alpha^2 + \beta^2$$

$$M = -\frac{4\alpha^2}{9L^2} - \frac{2\alpha - 3\alpha}{18\sigma L} - \frac{1}{9\sigma^2}$$

$$N = -\frac{8\alpha^3}{27L^3} - \frac{\alpha(2\alpha - 3\sigma)}{18\sigma L^2} - \frac{\alpha + 3\sigma}{18\sigma^2 L} - \frac{1}{27\sigma^3}$$

$$R_1 = \sqrt[3]{N + \sqrt{M^3 + N^2}} + \sqrt[3]{N - \sqrt{M^3 + N^2}} - \frac{2\alpha}{3L} - \frac{1}{3\sigma}$$

$$R_2 = -\frac{2\alpha + \sigma}{\beta^2(\sigma R_1 + 1) + \alpha(\alpha\sigma R_1 + \alpha + 2\sigma)} - 2R_1$$

$$R_3 + -R_1 - \frac{2\alpha}{L} - \frac{1}{\sigma}$$

The values for the two feedback resistors (labeled R_f in the schematic) are arbitrary.

Unity-Gain Highpass

Figure 3 shows the schematic for a unity-gain highpass filter. The element values for this filter can be found by inverting the values in the corresponding lowpass filter (Figure 1).

Highpass with Gain

$$R_{1_{HP}} = \frac{1}{C_{1_{LP}}}$$

$$R_{2_{HP}} = \frac{1}{C_{2_{LP}}}$$

$$R_{3_{HP}} + \frac{1}{C_{3_{LP}}}$$

Figure 4 shows the schematic for a normalized highpass filter with 6 dB gain. The element values for this filter are found by inverting the lowpass values calculated for Figure 2 as follows:

$$C_{1_{HP}} = \frac{1}{R_{1_{LP}}}$$

$$C_{2_{HP}} = \frac{1}{R_{2_{LP}}}$$

$$C_{3_{HP}} = \frac{1}{R_{3_{LP}}}$$

Frequency and Impedance Scaling

The filters just described are normalized for –3 dB cutoff at $\omega = 1$ radian/second. To adjust this cutoff frequency and substitute more realizable component values, scaling is required. The component values for an arbitrary cutoff frequency are found as follows:

For Figures 1 and 4: First, choose a resistor value. This value is Z.

Next, calculate the frequency scaling factor: $FSF = 2\pi f_c$, where f_c is the desired –3 dB cutoff frequency in Hz. Scale the resistor values:

$$R_{1,2,3} = Z$$

Finally, all capacitors are scaled by the same amount:

$$C_i^D = \frac{C_i^N}{Z \times FSF}$$

Where C_i^N is the normalized capacitance value, and C_i^D is the denormalized (scaled) value.

For Figures 2 and 3: First, choose a capacitor value. This value is C.

$$C_{1,2,3} = C$$

$$FSF = 2\pi f_c$$

Where f_c is the desired cutoff frequency in Hz.

Now, scale the resistors:

$$R_i^D = R_i^N\left(\frac{1}{C \times FSF}\right)$$

Where R_i^N is the normalized resistance of R_i, andR_id is the denormalized (scaled) value.

Design Example

To design a filter from the equations presented, first choose a topology. In this example a 15 KHz, 6 dB gain lowpass filter is designed and simulatec A sensitivity analysis is then pe formed and compared with a conver tional two-section filter, which realize the same transfer function.

The poles for a third order (three pole) Butterworth lowpass filter ar given in [1] as follows:

Pole #1: $s = -0.5 + j0.866$
Pole #2: $s = -0.5 - j0.866$
Pole #3: $s = -1$

Therefore,

$$\alpha = -0.5$$
$$\beta = 0.866$$
$$\sigma = -1$$

From the equations presented in th lowpass gain section above,

$L = 0.99996$
$M = -0.11111$
$N = 0.21297$
$R_1 = 1.565\ \Omega$
$R_2 = 1.469\ \Omega$
$R_3 = 0.4348\ \Omega$

Choosing capacitor values of 470 pF, the resistors are scaled (denormal ized) as follows:

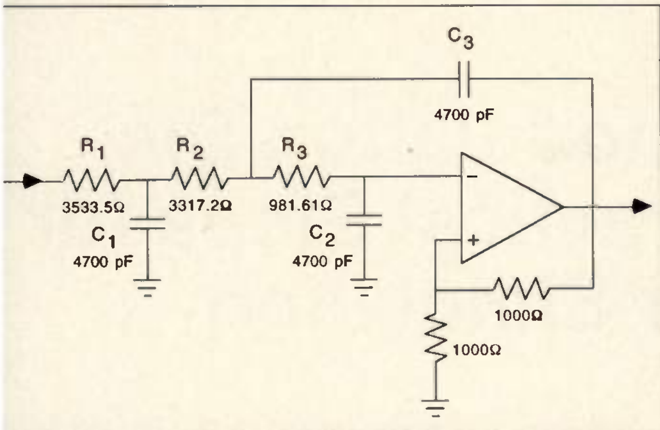
$$FSF = 2\pi(15 \times 10^3)$$

$$R_1 = \frac{1.565}{(4700 \times 10^{-12})(30000\pi)} = 3533.5\Omega$$

$$R_2 = \frac{1.469}{(4700 \times 10^{-12})(30000\pi)} = 3317.2\Omega$$

$$R_3 = \frac{0.4348}{(4700 \times 10^{-12})(30000\pi)} = 981.61\Omega$$

Since the two feedback resistors ca be any value (as long as they ar equal), 1 kΩ is chosen for this example The schematic for the final circuit i shown in Figure 5. For comparison Figure 6 shows a schematic for a typi



'igure 5. Schematic for 15 KHz lowpass design example.

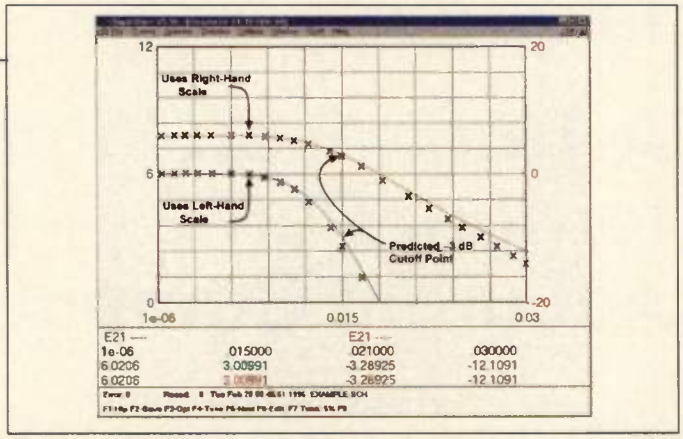
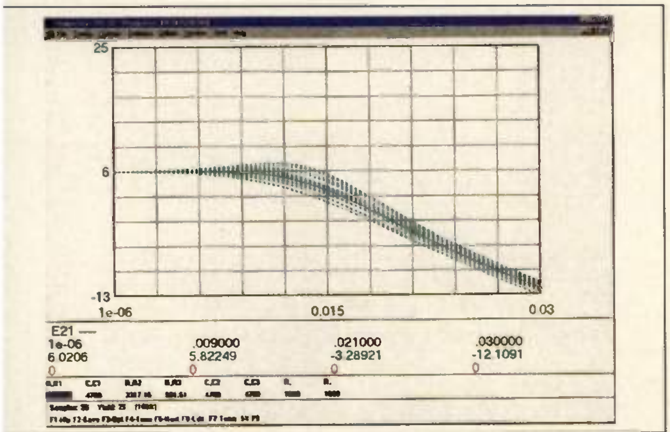


Figure 7. Predicted and measured response for the filter shown in Figure 5.



'igure 8. Sensitivity analysis for the filter of Figure 5.

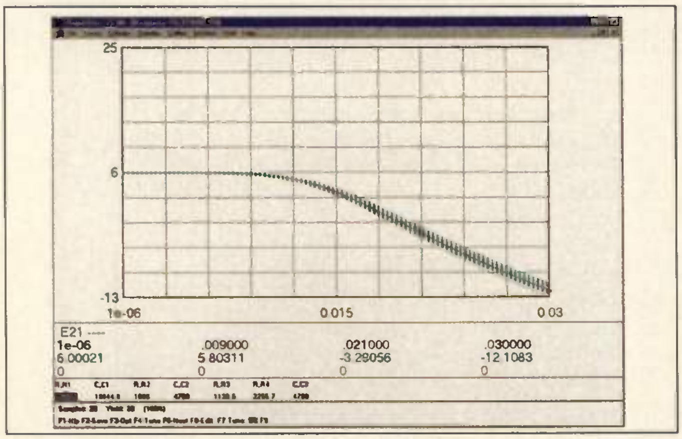


Figure 9. Sensitivity analysis for the filter of Figure 6.

al 2 section, third order filter that 'ealizes the same transfer function as 'igure 5.

The schematic in Figure 5 was con- ;tructed using 1% resistors, 5% capaci- ors, and a Harris CA3450 op-amp. [he circuit was simulated using Eagleware's =SuperStar= simulator 2], and the simulated and measured 'esponses are shown in Figure 7. The solid traces correspond to the predict- ed voltage gain versus frequency, whereas the Xs mark the actual mea- sured gain at each frequency. The horizontal axis is labeled as frequency in MHz.

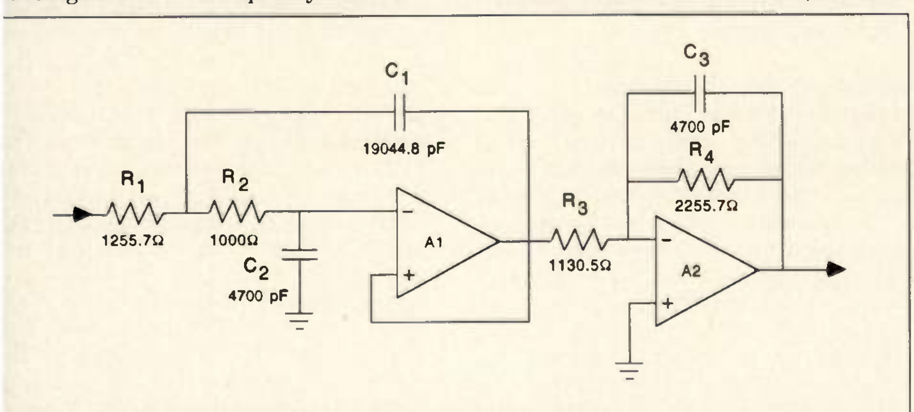
A sensitivity analysis is shown in Figures 8 and 9 for the single section, and two-section filters, respectively. The sensitivity analysis performed is called a Monte Carlo analysis, and consists of repeated simulations after varying each component within a specified tolerance.

Twenty-five random samples were taken, with each component value lying randomly within a given part tolerance (1% for each resistor, and 5% for each capacitor).

By tuning parts individually, C_2 and C_3 are found to be the most sensitive components in the single-section filter. This means that a series or parallel trimmer cap may be necessary for these two components, or tighter tolerance parts may be used. ***RF***

References

1. Arthur B Williams, Fred J. Taylor, *Electronic Filter Design Handbook*, McGraw-Hill, 1988
2. =SuperStar= is a product of Eagleware Corporation, Stone Mountain GA, (770) 939-0156



'igure 6. A typical 2-section, 3 pole lowpass filter.

About the Author

Glenn Parker received a BEE (Bachelor of EE) from Georgia Tech in 1994. He has been employed at Eagleware in Stone Mountain, GA as a software/hardware engineer for 3 years. Glenn led the development of Eagleware's active filter program, =A/FILTER=. He can be reached be telephone at (770) 939-0156 or by e-mail at eagleware@ eagleware.com.

A Tutorial on Intermodulation Distortion: Part 2 — Practical Steps for Accurate Computer Simulation

By Jeffrey Pawlan
Pawlan Communications

This is the second of a two-part series on intermodulation distortion (IMD). In this article, a real amplifier is modeled and the results of the non-linear simulation are compared to the actual performance of the amplifier. The necessary steps to getting an accurate model are presented, along with potential problems.

Last month, in the first article, it was mentioned that a wide dynamic range amplifier was built using lossless feedback technology. The realization presented here uses an RF transformer with three windings, which is non-trivial for microwave simulators. The initial design was modeled, and both linear and non-linear analysis were done on the computer using Microwave Harmonica [1]. The amplifier was designed for the HF-VHF region of 5 MHz to 80 MHz with goals of 8.5 dB minimum gain and a minimum third order intercept point of +43 dBm. Second order or harmonic distortion was not as critical. The goal was –40 dBc for the harmonics at a carrier power output level of +12 dBm. Several high performance low cost surface mount bipolar transistors were evaluated for this application. The non-linear simulation would also show the effect of various bias currents on the third order intermodulation distortion. Once a good model had been established for the circuit, several different transistor models could be tried in the circuit. The results of using two different transistors are presented. The final working schematic of the amplifier is shown in Figure 1.

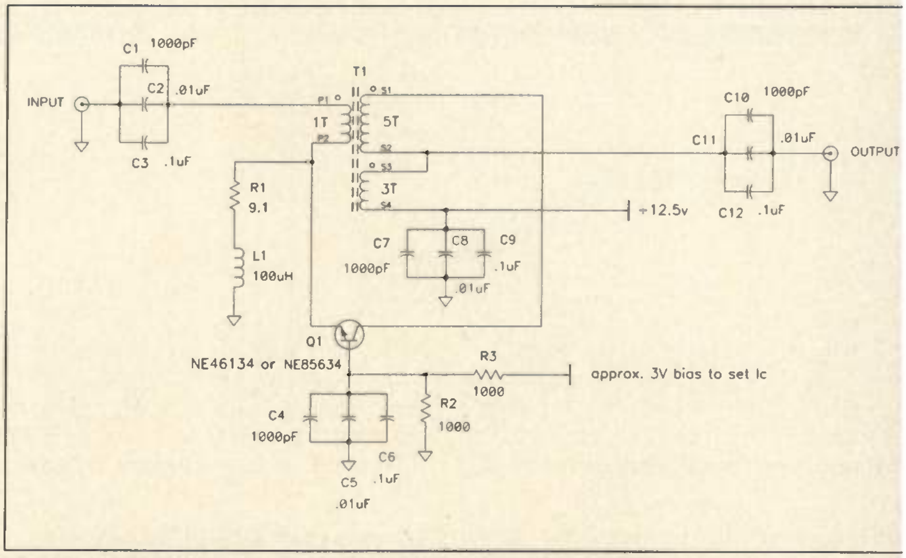


Figure 1. Final working schematic of the example amplifier.

Crucial Steps in the Creation of an Accurate Circuit Model

1. The most important element is the transistor. The model parameters provided by the vendor may not be complete or accurate; therefore, they must be verified.

2. The next most important element is the transformer. As it is not an ideal transformer, a special model had to be developed.

3. The wire leads on the transformer need to be included in the circuit model.

4. The capacitors and resistors did not require special modeling because they were size 0805 chip components and at this frequency they are nearly ideal components.

Modeling the Transistor

Some transistor manufacturers provide non-linear model parameters, as well as models for their package parasitics. Unfortunately, not all manufacturers provide them and those that do have them may not have them in the package that you want to use. Often, a few of the parameters are missing, as well. Given a sufficient project budget, one can contract custom parameter extraction of any device, bipolar or otherwise. If you have an ongoing need for parameter extraction, then it may be prudent for your company to inves in an automated parameter extractio system consisting of test fixtures, vector network analyzer, compute controlled voltage and current source bias tees, and parameter extractio software to control and automate th process. In this particular case, cus tom parameter extraction was not nec essary.

In addition to the non-linear mode parameters, it is highly desirable t have a set of S-parameters at the actu al operating bias selected for th devices. This will allow you to chec your circuit using the linear simulato It is highly unlikely that you will fin published S-parameters at just th right bias voltage and current for you application, so you will need to gener ate them. Again, if you have th appropriate test equipment, you coul measure them. However, once yo have a good non-linear model for th device you will be able to generat them on the computer. In fact, the wa that your non-linear model can b

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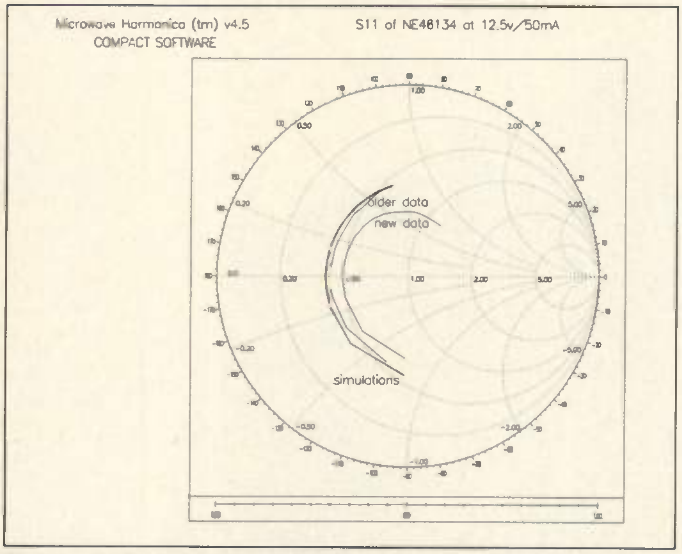

Figure 2. Comparison between the non-linear simulated S_{11} of a NE46134 and the published data from California Eastern Labs (polar plot).

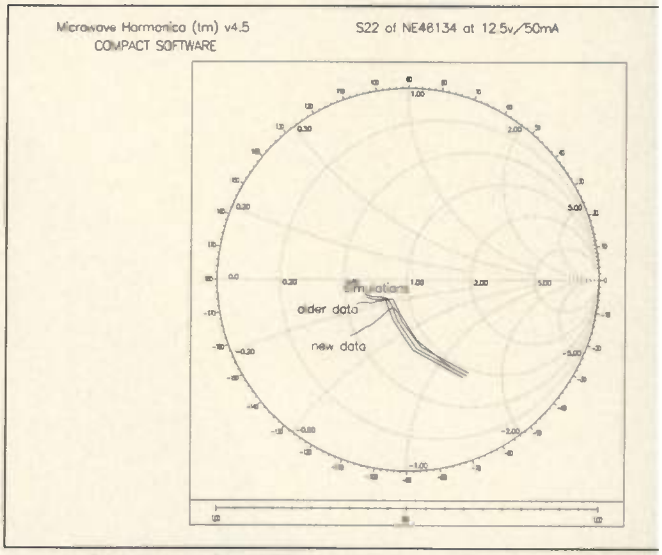

Figure 3. Comparison between the non-linear simulate S_{22} of a NE46134 and the published data from Californi Eastern Labs (polar plot).

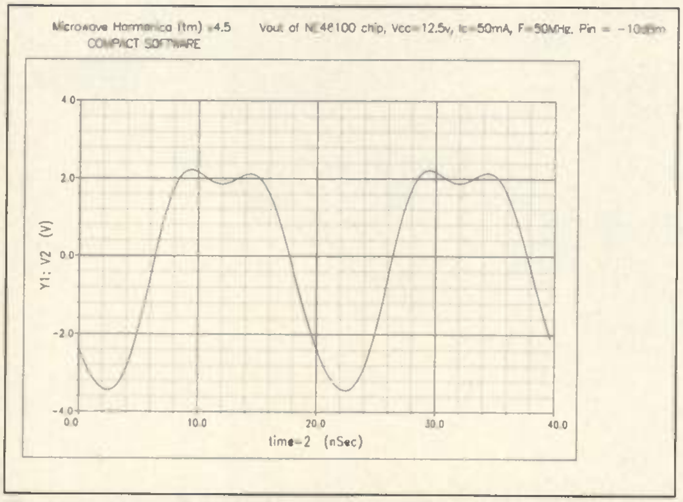

Figure 4. The distorted output waveform from a NE46100 chip at 50 MHz when a power input level of –10 dBm is used (rectangular plot).

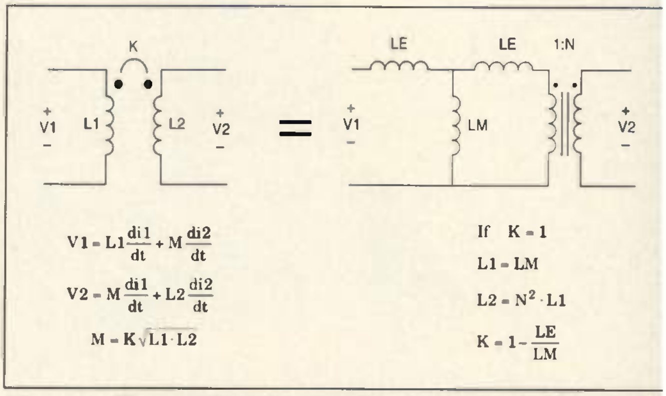

Figure 6. The equivqlent circuit for two mutually couple inductors (compliments of Intusoft). The coupled inducto transformer on the left is computationally efficient, but i cannot provide access to LE and LM or be used as a build ing block. The ideal transformer with discrete inductance and their relationship to the coupling coefficient is show on the right.

tested for accuracy is by generating S-parameters at the same bias voltage and current as published S-parameter data from the manufacturer and seeing how well they match. When you have the non-linear model verified, you can generate the S-parameters at exactly the bias that you want. Two SPICE circuit files exist that will do this for you [2]. Better still, Microwave Harmonica will generate large signal S-parameters. These are available after non-linear simulation as GC parameters.

There are two very important caveats to this. First, sometimes the published S-parameters are not taken in the same de-embedded fixture as the data points used to create the non-linear model parameters. This occurred with the NEC S-parameter disk version 8. The data on this disk for the NE46134 transistor was significantly different from data that was on the disk version 4. The S-parameter data on the earlier disk agreed very well with the non-linear model, but the newer data did not. A call to California Eastern Labs revealed that they had made a radical change in their fixturing. They saw that the majority of their customers were newcomers to RF design and did not know that it was necessary to include parasitic capacitances of the printed circuit board i their circuit designs, especially whe using FR-4 material, which has a hig capacitance. CEL was trying to mak life easier for these designers so the mounted the transistors on FR-4 mate rial and included the capacitances i the S-parameter measurements Always test your non-linear mode with the package parasitics befor attempting to use it in a circuit. If yo do not get very good correlation to th measured S-parameter data, call th manufacturer and work with them t determine where the problem is.

Figure 2 shows the compariso between the non-linear simulation o

Input power (dBm)	GC11 MAG	ANG	GC21 MAG	ANG	GC12 MAG	ANG	GC22 MAG	ANG
–40	0.521	–91.929	41.579	127.737	0.021	59.481	0.590	–60.534
–30	0.522	–91.761	41.520	127.797	0.021	59.486	0.591	–60.344
–20	0.529	–89.898	40.849	128.460	0.022	59.597	0.603	–58.173
–10	0.632	–62.359	28.624	136.193	0.027	65.161	0.690	–36.011
0	0.861	–25.942	9.119	148.787	0.030	75.720	0.810	–19.625

Figure 5. Large signal S-parameters vs. input power of a NE46134 at 50MHz

S_{11} and the measured S_{11} of a NE46134. Figure 3 shows the comparison for S_{22}. Another very important pitfall that can cause major inaccuracy is the RF input power level of the measurement at frequencies where the device has significant gain. This applies equally to physical measurements with a Vector Network Analyzer and to non-linear simulation. The device will be driven into non-linear large signal operation and will distort if the input signal is more than –30 dBm. It may be necessary to reduce the input power to –40 dBm to ensure linear region operation. Figure 4 shows the distorted output waveform from a NE46100 chip at 50 MHz when a power input level of –10 dBm was used. It is not generally published how significantly different the S-parameters can be at various power levels.

The non-linear model of the NE46134 was entered in Microwave Harmonica and the power level was stepped at a single frequency: 50 MHz. The resulting large signal S-parameters vs. input power are shown in Figure 5. Note how much the S_{11} phase and the S_{21} magnitude change. This should be a sobering reminder to be very careful to document the power level when taking device measurements.

It is important to check to model parameter names supplied by the manufacturer against the non-linear element model in the software. Microwave Harmonica uses the SPICE conventions for most, but not all, of the Gummel-Poon parameters. The following short list shows the differences between Berkeley SPICE and Microwave Harmonica for the bipolar transistor parameters:

SPICE	Microwave Harmonica
VAF	VA
VAR	VB
RE	RE1
RC	RC2
FC	FCC
Fc	FC

Modeling the Transformer

The amplifier topology shown in Figure 1 uses a transformer with three windings. It is not ideal; therefore, it is more accurately called three mutually coupled inductors. SPICE can have many mutually coupled inductors in a circuit, but harmonic balance simulators usually only allow a maximum of two inductors to be coupled. The transformer model in Microwave Harmonica allows three windings, but they are perfectly coupled. Two coupled inductors with less than perfect coupling may be represented as shown in Figure 6. The ideal transformer provides the turns ratio and the external inductors provide the model for the

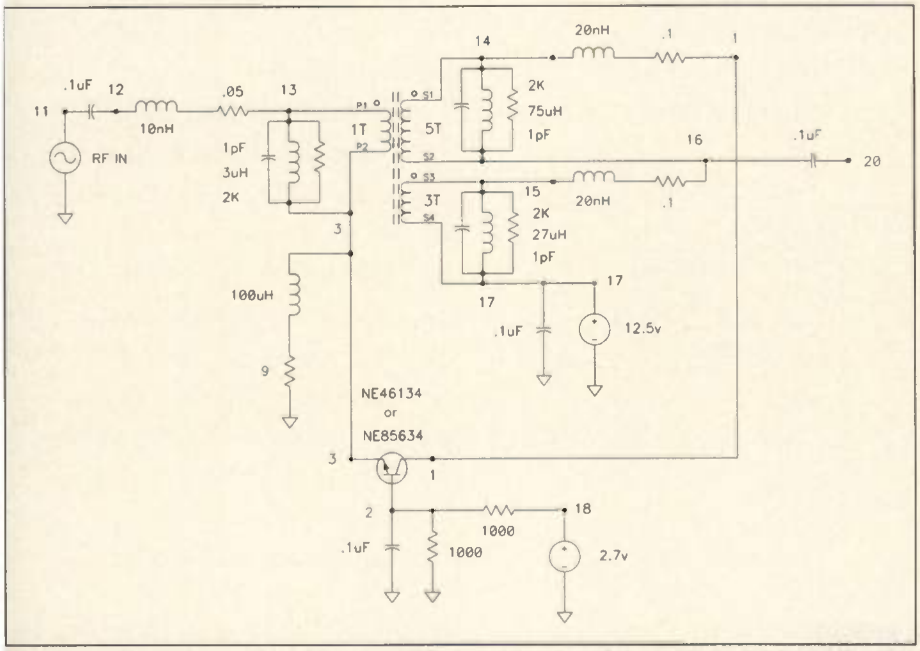

Figure 7. The schematic of the amplifier as simulated, including the transformer model.

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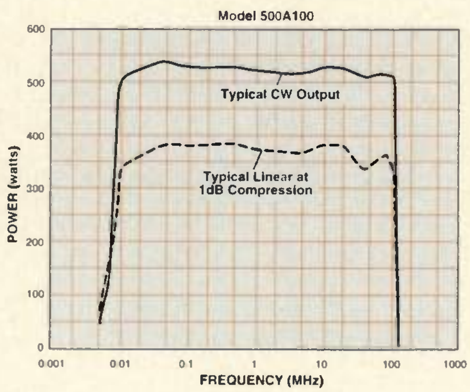

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```
* hfampn120.ckt NON- LINEAR ANALYSIS OF A TRANSFORMER FB HF AMP
* MICROWAVE HARMONICA FILE BY Jeffrey Pawlan
* this file has single tone swept frequency input for S21 & S11 generation
* This includes the NEC transistor and pkg model parameters for the NE46134
CTRL
CFNEWT=6E-1
END
LB: 1.2NH
LE: 1.2NH
CCB: .03PF
CCE: .18PF
CCBpkg: .18PF
CCEpkg: .18PF
CBEpkg: .01PF
Tbz: 60
Tbl: 50mil
Tbk: 1
NBLK
IND 4 5 L=LB
IND 6 7 L=LE
CAP 4 7 C=CBEpkg
CAP 5 1 C=CCB
CAP 4 1 C=CCBpkg
CAP 1 6 C=CCE
CAP 1 7 C=CCEpkg
TRL 2 4 Z=Tbz P=Tbl K=Tbk A=Tba
TRL 3 7 Z=Tbz P=Tbl K=Tbk A=Tba
BIP 5 1 6 MOD=NE46134 NAME=Q1
CAP 11 12 C=.1UF
SRL 3 0 R=9 L=100UH
SRL 12 13 R=.05 L=10NH
PRX 13 3 R=2000 L=3UH C=1PF
CAP 2 0 C=.1UF
RES 2 0 R=1000
RES 2 18 R=1000
CAP 17 0 C=.1UF
PRX 14 16 R=2000 L=75UH C=1PF
SRL 14 1 R=.1 L=20NH
TRF 13 14 15 3 16 17 N1=5 N2=3
BIAS 17 0 V=12.5
BIAS 18 0 V=2.7
AMP: 2POR 11 20
END
FREQ
INTM 3
TONE 1 STEP 1 MHZ 70 MHZ 1 MHZ
TONE 2 STEP 1.1 MHZ 70.1 MHZ 1 MHZ
END
NEXC
P1<H1+H0> -30DBM
P2<H0+H1> -30DBM
END
NOUT
TG21<H1+H0,H1+H0> DB
RL1<H1+H0> DB
RL2<H0+H1> DB
END
DATA
.MODEL NE46134 BIP(NPN BF=185 VA=30 VB=12.376 RC2=3 RB=6 RE1=.63
+IKF=.35 ISE=.57E-12 TF=12.9E-12 TR=1.7E-8 ITF=.4 VTF=19.875 CJC=2.5E-12
+CJE=4.9E-12 XTI=3 NE=1.8 ISC=1E-14 BR=5 VJC=.83 VJE=.6 IS=8.7E-16
+MJC=.33 MJE=.45 XTF=1.6 IKR=.01789 NC=1.95 FCC=.5 RBM=4 IRB=.004
+XCJC=.2 NF=.959 NR=1)
END
```

Figure 8. The complete Microwave Harmonica circuit file listing for the example amplifier (plain ASCII format).

inductance of each inductor along with the mutual coupling. This can be extended to three windings. A slight departure from this was used in the computer simulation of this amplifier. The schematic shown in Figure 7 shows the three winding transformer and the external parasitic elements. The leads are also modeled as a series resistance and inductance as they are made of fine wire and have significant length at these frequencies.

The Non-linear Simulation of the Complete Amplifier

The schematic as shown in Figure 7 was entered into the circuit file. The Microwave Harmonica non-linear circuit file must have the netlist for the amplifier entered in a NBLK section. This is acted on by a FREQ section that specifies the frequencies and the number of input signals to the amplifier. A NEXC section specifies the excitation for the circuit. Other sections such as output definition, optimization goals, data, simulation control, and variable initialization are optional. In the circuit file shown in Figure 8, the transistor model parameters are placed in the DATA section. The NEXC section defines the power level of the input signals. The syntax of the signals in both the FREQ and the NEXC sections specify whether the signals are applied at the input to the amplifier or the output. They also specify whether a harmonic or the fundamental is to be used.

It is important to simulate the circuit under the same bias conditions as the physical circuit. The true operating point of thetransistor is checked by generating a time domain plot of the Vce and the Ic of the transistor while the signals are applied to the input. You will be able to visually see the operating points from these plots. In this amplifier, the desired bias was approximately 12.5 V at 48 mA collector current. Figure 9 shows these plots. If the current is not correct, adjust the base bias supply voltage then re-run the simulation.

After the non-linear simulation is run, the program stores all generated data in a file. It is then possible to specify what output format and which data is to be displayed without ever running the simulation again, unless you change the circuit or a component value. The predicted gain of the amplifier is shown in Figure 10 and the complex input and output impedances are shown in Figure 11.

Third Order Intermodulation Distortion

In order to simulate third order intermodulation distortion, two closely spaced signals are applied to the input of the amplifier. These have equal power levels which are chosen to produce the desired output power level. It is important to choose an output power level that is representative of the level that the circuit will see in its operation. As shown in Part 1, the intercept point is unreachable because the amplifier is into saturation long before this point is reached. This is a fictitious point that is created by extrapolating both the desired and the distortion product signals then calculating the resulting intersection. Figure 12 shows the conversion from absolute power levels to third order intercept point and vice-versa. These formulae have the power levels and the third order intercept point in dBm.

It is easy to simulate intermodulation distortion performance using Microwave Harmonica. Unlike SPICE, the computational engine can be told to focus on the specific frequencies of the two input signals and then the exact frequencies of the intermodulation products. This makes the result much more accurate and significantly reduces the simulation time. The

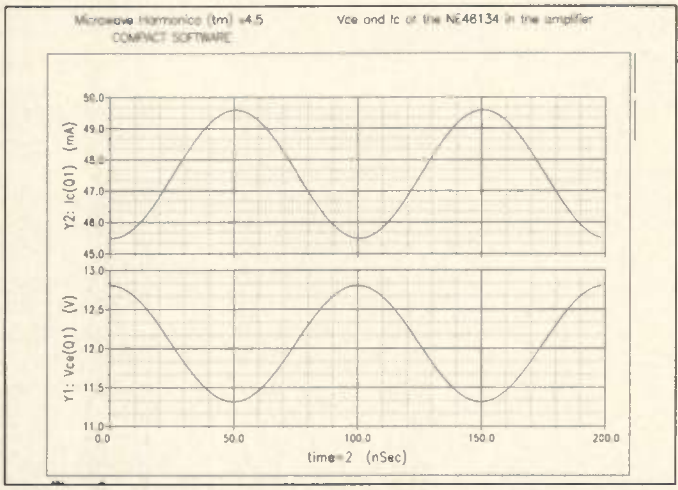

Figure 9. The V_{CE} and I_C of the NE46134 transistor in the amplifier, showing the operating points (rectangular plot).

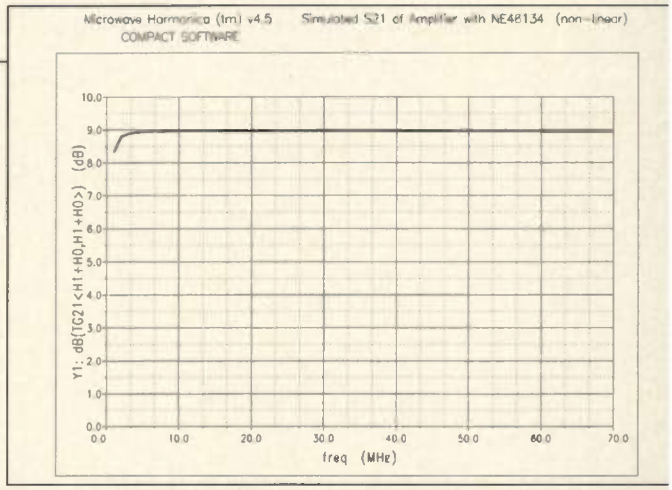

Figure 10. Predicted gain (S_{21}) of the amplifier (rect. plot).

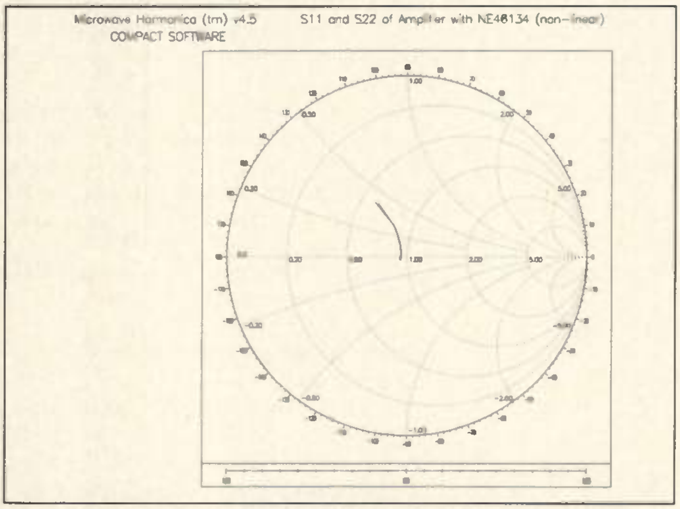

Figure 11. The complex input and output impedances (S_{11}, S_{22}) of the amplifier (polar plot).

3rd Order Intermodulation Distortion and the Intercept Point (IP_3)

1. Always measure the distortion products at the output level that the amplifier will normally be used.
2. No amplifier is ideal, therefore, the Intercept Point is a fictitious becnchmark derived by extrapolation. You cannot directly measure it.
3. The absolute power levels of the distortion products and the output signals may be converted to IP3 as follows:

Let: P_o = power level of the two output signals in dBm
P_3 = power level of the 3rd order products at the output
IP_3 = 3rd order Intercept Point at the ouput in dBm

$$IP_3 = \frac{3P_o - P_3}{2} \quad \text{and,} \quad P_3 = 3P_o - 2IP_3$$

Figure 12. Engineering formulae showing relationship o absolute power levels to 3rd order intercept point.

FREQ section shown in Figure 13 includes the two input frequencies and an additional line that specifies that a third order intermodulation analysis is to be done. This amplifier was designed to produce an output power of +14 dBm. Since the gain was previously predicted to be 8.9 dB, the levels of the input signals are set to +5.1 dBm each. The simulation was run, and a spectrum plot of the output was generated. A table was also generated so that the exact levels of the intermodulation distortion products would be printed. The plot and the table are shown in Figures 14 and 15 respectively. The two third order products are –54.6483 dBm and –54.3488 dBm. Since the two have slightly different levels, they should be averaged together. Using the formula given in Figure 12, the third order output intercept point was calculated to be +48.3 dBm.

The amplifier was also simulated with a different transistor at a different collector current. The model parameters for a NE85634 were substituted and the base bias was adjusted for a current of approximately 35 mA. The intermodulation performance was slightly less and is shown in Figures 16 and 17. The third order output intercept point was calculated to be +46.55 dBm for this transistor and lower bias current.

The prototype amplifier was built using all chip components except for the transformer. The circuit was built on a small piece of unetched copper clad circuit board. The transformer was handmade by starting with trifilar wire.The wires were twisted tightly to increase coupling. The core was a two hole ferrite balun. Initially, the transistor was the NE46134. After measurements were made, the transistor was changed to a NE85634.

The amplifier with a NE46134 installed was tested for S_{21} (gain), S_1 (input match), and S_{22} (output match with a vector network analyzer. Th plots of these are shown in Figures 18 19, and 20. A test setup for measurin the third order intercept point was cre ated as shown in Figure 21. The out puts of both synthesized signal genera tors are individually amplified the attenuated before combining. This is t insure a high degree of back isolation otherwise one signal generator's out put would be present at the output o

```
FREQ
INTM 3
TONE 1 28MHZ
TONE 2 28.4MHZ
END
```

Figure 13. Excerpt from a circuit file showing the FREQ section for third-order intermodulation analysis. (plair ASCII file).

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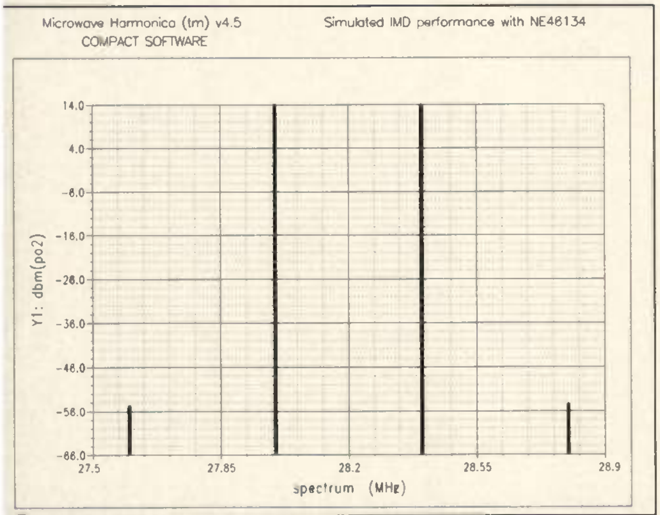


igure 14. Plot of simulated third-order intermodulation istortion for NE46134. (Rect plot).

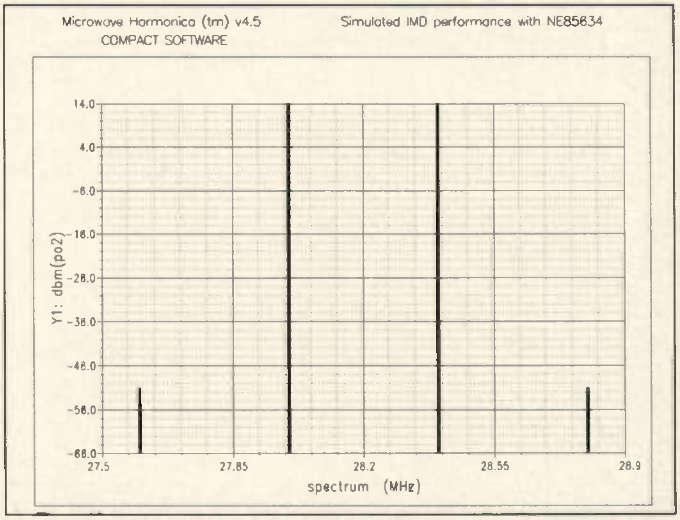


Figure 16. Plot of simulated third-order intermodulation distortion for NE85634, (Rect Plot).

Non-Linear Simulation of Intermodulation Distortion for Amplifier with a NE46134 transistor

Power level (dBm)	Frequency (MHz)
–76.9215	0.40
–54.6483	27.60
14.0500	28.00
14.0499	28.40
–54.3488	28.80
–38.3773	56.00
–32.6222	56.40
–38.2549	56.80
–54.5307	84.00
–45.9738	84.40
–45.9335	84.80
–54.4096	85.20

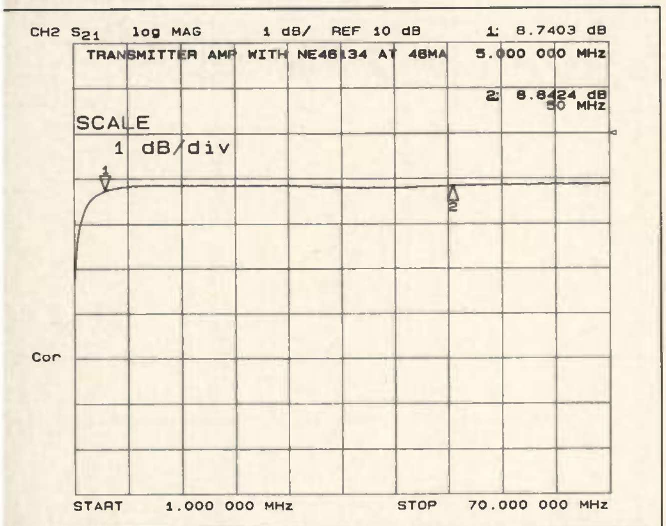
igure 15. Table of simulated inter- ıodulation distortion products for IE46134. (ASCII file).

Non-Linear Simulation of Intermodulation Distortion for Amplifier with a NE85634 transistor

Power level (dBm)	Frequency (MHz)
–65.6700	0.40
–51.0452	27.60
14.0262	28.00
14.0261	28.40
–51.0051	28.80
–45.0697	56.00
–39.4342	56.40
–44.9543	56.80
–56.9705	84.00
–48.3820	84.40
–48.3593	84.80
–56.8979	85.20

Figure 17. Table of simulated intermodulation distortion products for NE85634. (ASCII file).

the other signal generator, which would likely create intermodulation products and invalidate the measurement. The combiner must also be resistive; do not use a power splitter as the ferrite core in this may create non-linearities. The actual spectrum analyzer plot of the amplifier with a NE46134 installed is shown in Figure 22. The two desired output signals were set to +14 dBm. The measured third order intermodulation products were –55.2 dBm and –56.3 dBm. Their average was –55.75 dBm. The calculated third order intercept point was +48.88 dBm. The transistor was changed to the NE85634 and the bias set for a collector current of 35 mA and the amplifier was re-measured. The spectrum ana-



igure 18. Actual measured performance of amplifier with IE46134 for S_{21} (gain).

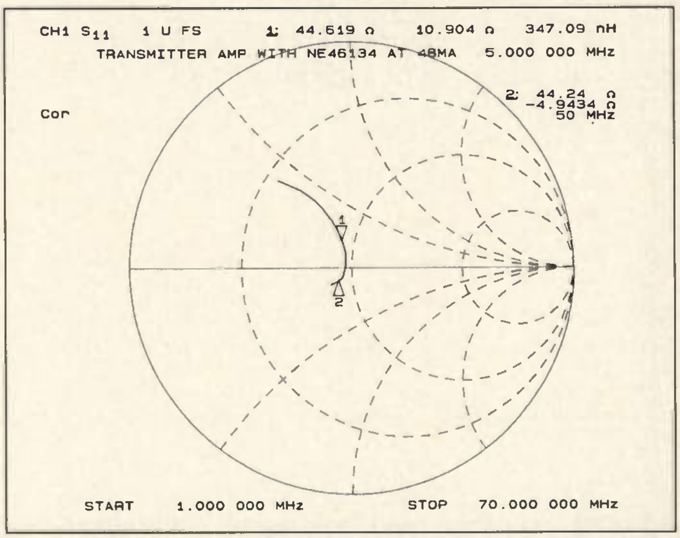


Figure 19. Actual measured performance of amplifier with NE46134 for S_{11} (input match).

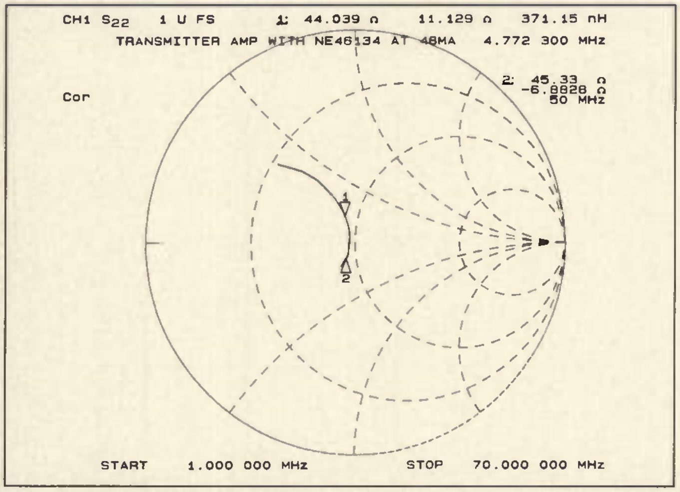


Figure 20. Actual measured performance of amplifier with NE46134 for S_{22} (output match).

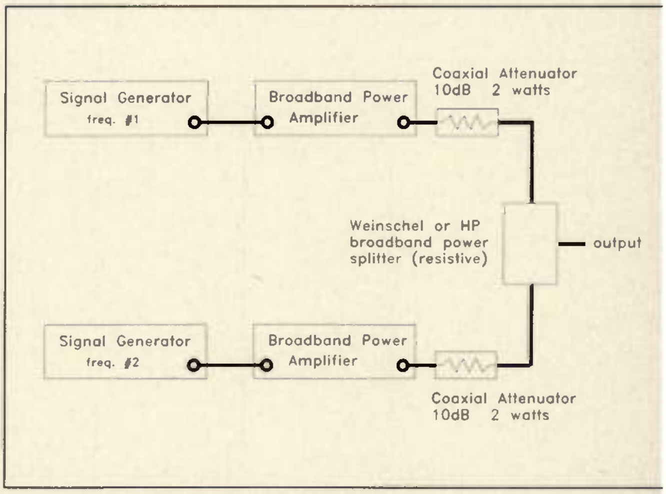


Figure 21. Block diagram of a test setup for measurin third-order intercept point.

lyzer plot of the amplifier with a NE85634 installed is shown in Figure 23. The measured third order intermodulation products were –52.0 dBm and –54.0 dBm. The average of these is –53 dBm. The calculated third order intercept point was +47.5 dBm.

Simulated vs. Measured Performance

The gain and frequency response of the amplifier agreed within .05 dB. The input and output impedances agreed in phase, and the magnitude differed only slightly, This was probably due to core losses of the transformer. The third order intercept points agreed within 1 dB, which is better than the stated accuracy of the spectrum analyzer making the measurement. These results along with ease of use show that harmonic balance simulation of non-linear circuits has come of age. The use of proper computer simulation tools can help shorten the development time of a product and can be used to create a design that is more manufacturable. It is still recommended that a prototype always be built before committing a design to production. *RF*

Notes

[1] Microwave Harmonica is available from Compact Software Inc., Paterson, NJ; (201) 881-1200.

[2] A SPICE circuit file called SPARAMS.CIR is included on the RF Library disk from Intusoft Corp. A different SPICE circuit file is included on California Eastern Laboratories' NEC diskette.

About the Author

Jeffrey Pawlan is the owner of Pawlan Communications, 14908 Sandy Lane, San Jose, CA 95124-4340. He can be reached at: tel: (408) 371-0256; fax: (408) 371-4302. For the last 5 years he has predominantly been an independent consultant and has designed many wireless LANs, RFID tag systems, and handheld products. His favorite research areas are low noise wide dynamic range amplifiers, and low phase noise oscillators. Past experience includes design of CATV hybrid amplifiers, mm-wave communications systems, L-band satellite ground stations, plus teaching analog circuit design and SPICE.

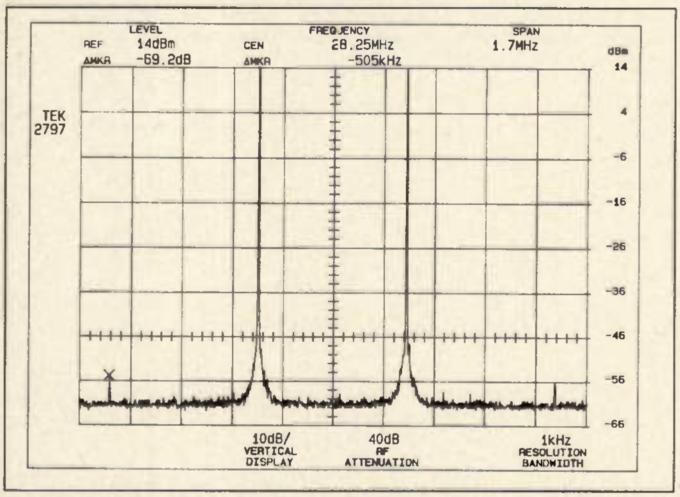


Figure 22. Actual spectrum analyzer plot of the output signals and intermodulation distortion of the amplifier with a NE46134 installed.

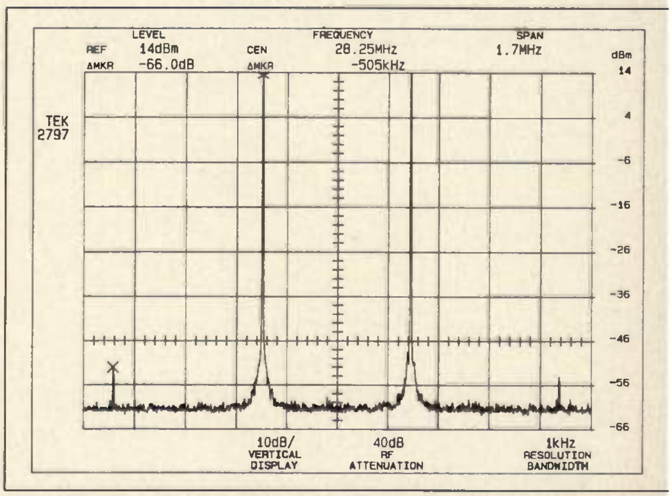


Figure 23. Actual spectrum analyzer plot of the output sig nals and intermodulation distortion of the amplifier with a NE85634 installed.

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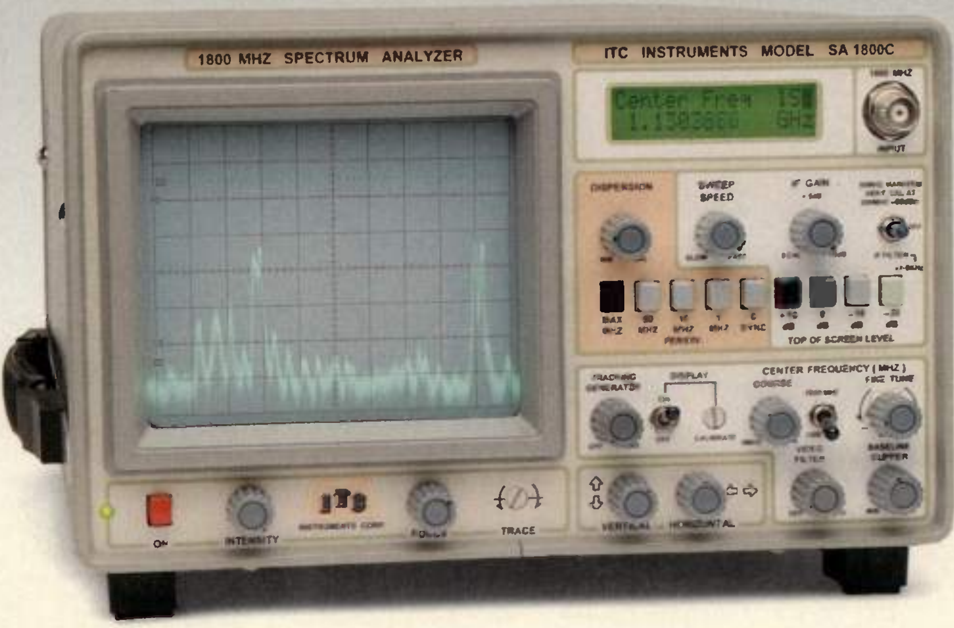

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HP EEsof
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Engineering Plastics Guide

DSM Engineering Plastic Products has released a new Advanced Engineering Plastics Material Guide. The brochure profiles DSM's line of high performance materials including 29 different extruded, compression and injection molded products. Electrostatic Discharge (ESD) dissipative materials are included in the product line.
DSM Engineering Plastic Products
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Coaxial Products Catalog

Pasternack Enterprises has published catalog #1996, containing connectors and adapters, cables, cable assemblies, amplifiers, switches, attenuators, couplers and other coaxial interconnection and signal processing products.
Pasternack Enterprises
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GaAs Semiconductor Guide

Oki Semiconductor has prepared a new guide to the KGF16xx family of GaAs RF semiconductors. The catalog includes product data for the KGF1606 Power FET for CDMA/Spread Spectrum, the KGF1607 Power FET for FM/FDMA applications and the KGF1608 Power FET for GSM/TDMA applications. These three devices are ceramic-packaged and characterized for 500-3000 MHz operation. Also described are the plastic-packaged KGF1637 and KGF1638, Power FETs for FM/FDMA and GSM/TDMA, respectively, operating from 500-3000 MHz. All devices are specified for operation with a 3.4-volt supply.
Oki Semiconductor
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Frequency Domain Reflectometry Application Note

Anritsu Wiltron announces an application note on Frequency Domain Reflectometry (FDR) for microwave antennas. Application note AN54100A-4 is a step-by-step guide through maintenance planning, FDR measurement theory, measurement procedures, data storage, evaluation of costs and failure prevention. Sixty percent of a typical site's problems are caused by problematic cables, connectors and antennas. FDR techniques can identify small degradations in RF performance before they become major problems.
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Microwave Amplifiers

Applied Systems Engineering's new product summary list the company's capabilities in producing amplifiers and transmitters for radar, communications, EMC and laboratory use. Amplifier types include traveling wave tubes (TWT), klystrons and magnetrons, for CW or pulsed operation.
Applied Systems Engineering, Inc.
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SMT Couplers and Power Dividers Catalog

Anaren Microwave has a new catalog, *Wireless Surface-Mount Components*. Included in the catalog are descriptions and specifications of 27 models of 3 dB hybrid couplers, directional couplers and power dividers. Measured performance graphs are included, and 22 pages of the catalog are devoted to coupler theory and applications in mobile communications base station amplifiers.
Anaren Microwave
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Coaxial Connectors

RF Industries new catalog presents an expanded line of coaxial connectors, with 300 new items. Connector types include BNC, TNC, N, UHF, Mini-UHF, MB, SMB, SMA, MCX, 7/16 DIN, FMR, LMR series, plus 1/2" and 7/8" corrugated cable. 1500 products are listed, also including cable assemblies, connector kits, adapters and hand tools.
RF Industries
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Measurement Products

Tektronix announces the availability of its 1996 Measurement Products Catalog. Among the 900 products presented in the catalog are nearly 100 new items, and products offered by Tektronix' alliances with Advantest and Rohde & Schwarz. the catalog lists all Tektronix worldwide sales offices, distrbutors and representatives.
Tektronix, Inc.
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Cable Assemblies Catalog

Florida RF Labs' new short form catalog outlines their complete line of flexible, semi-flexible and semi-rigid coaxial cable assemblies. The catalog provides detailed data regarding their high-performance, low-loss and low-cost assemblies. Performance specifications, connector types and ordering information are also included.
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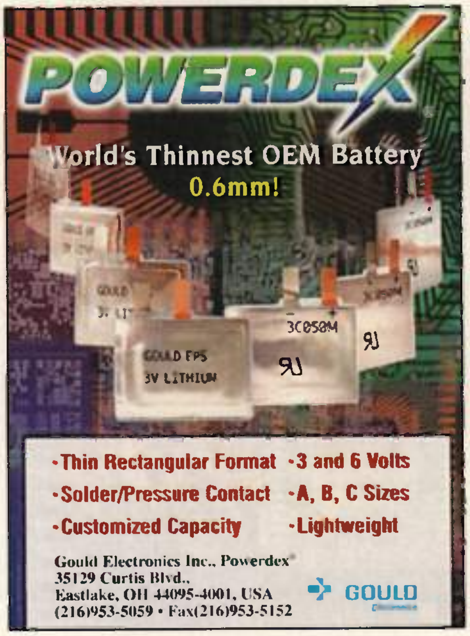

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Sr. MMIC Design: Design highly integrated GaAs MMICs for advanced cellular products.Circuits to be designed include power amplifiers, LNAs, mixers, IF amplifiers, buffer amplifiers.RF frequencies are 900 and 1800 Mhz.

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ZOS-75	37.5-75	-110	0.016	-26
ZOS-100	50-100	-111	0.026	-29
ZOS-150	75-150	-107	0.017	-26
ZOS-200	100-200	-106	0.015	-25
ZOS-300	150-280	-103	0.017	-27
ZOS-400	200-380	-100	0.021	-24
ZOS-535	300-525	-96	0.018	-27
ZOS-765	485-765	-96	0.033	-27
ZOS-1025	685-1025	-92	0.051	-25

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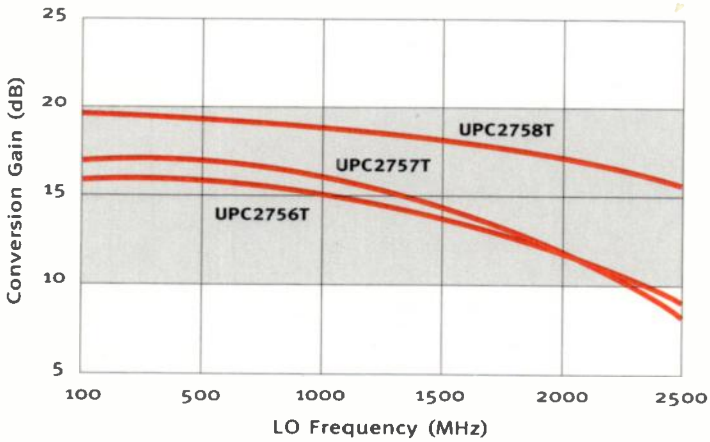
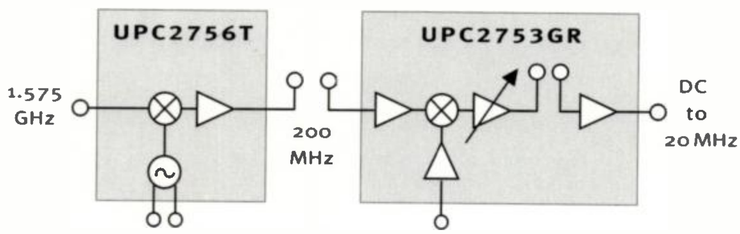
CONVERSION GAIN



Vcc = 3V LO Power = -5dBm IF = 200MHz

RECEIVER LINEUP



TYPICAL SPECIFICATIONS

PART	OPERATING VOLTAGE	Icc	CONVERSION GAIN	OUTPUT IP3
UPC2758T[1]	3V	11mA	17dB	+5dBm
UPC2757T[1]	3V	5.6mA	13dB	0dBm
UPC2756T[2]	3V	5.9mA	14dB	0dBm

1. Measured at 2.0GHz 2. Measured at 1.6GHz

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