

WIRELESS
LEADERSHIP
18 YRS

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RF design

engineering principles and practices

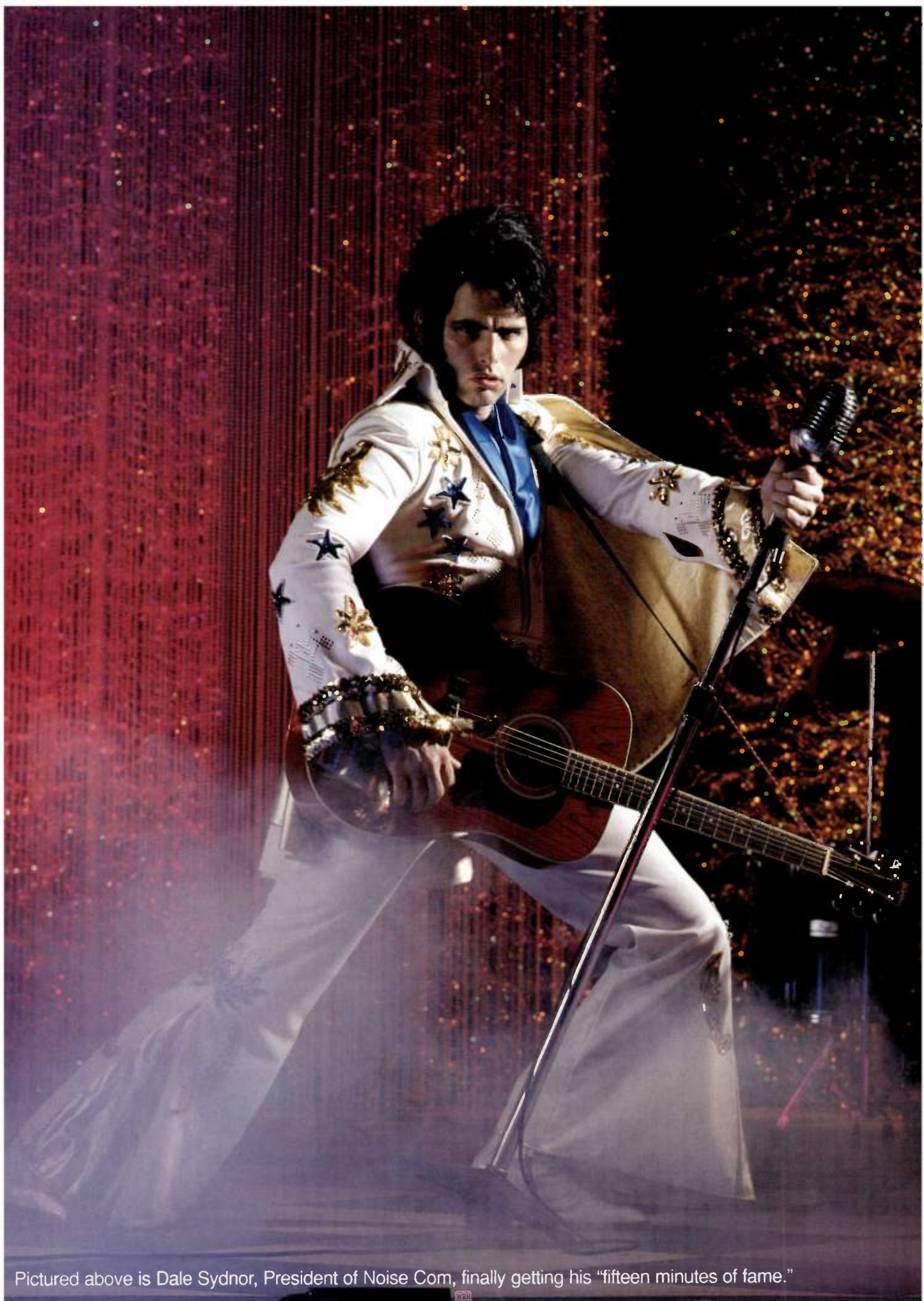
January 1997



***The future PCS
communicator will
travel far beyond
traditional
handsets***

RF design 97
Conference & Expo

**September 10-12, 1997
Santa Clara, CA**



Pictured above is Dale Sydnor, President of Noise Com, finally getting his "fifteen minutes of fame."

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To emulate multipath fading, the WIS Series can model wireless communication channels between base stations and mobile transceivers using Rayleigh, Rician, Log-Normal, Suzuki, and Nakagami fading statistics. The WIS Series is also the only CDMA test solution with an

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W201



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|----------------|----------|--------------------------|---------|--------------------|----------------------|-------------------|--|
| MODEL | LO (dBm) | FREQUENCY (MHz) LO/RF | I F | Conv. Loss | Isolation L-R L-I | S&A (qty. 1-9) | |
| JMS-1 | +7 | 2-500 | DC-500 | 5.75 | 45 45 | 4.95 | |
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| JMS-1H | +17 | 2-500 | DC-500 | 5.90 | 50 50 | 11.45 | |
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| JMS-2 | +7 | 20-1000 | DC-1000 | 7.0 | 50 47 | 7.45 | |
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| JMS-2MH | +13 | 20-1000 | DC-1000 | 7.0 | 50 47 | 10.45 | |
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Note: * 10-49 qty.

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|-----------|----------------|---------------|------------------------|------------------------|--------------------|
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| SKY-7G | 2000-7000 | +7 | 7.0 | 28 | 16.95 |
| SKY-60 | 2500-6000 | +7 | 6.2 | 28 | 14.95 |
| SKY-60LH | 2500-6000 | +10 | 6.2 | 28 | 16.95 |
| SKY-60MH | 2500-6000 | +13 | 6.2 | 28 | 17.95 |
| SKY-60H | 2500-6000 | +17 | 6.2 | 28 | 18.95 |
| SKY-53R | 2800-5300 | +7 | 5.7 | 28 | 14.95 |
| SKY-53LHR | 2800-5300 | +10 | 5.7 | 28 | 16.95 |
| SKY-53MHR | 2800-5300 | +13 | 5.7 | 28 | 17.95 |
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• IF: DC-500MHz min.



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| Model | *Freq. (MHz) | Gain (dB) | Max. Power Out (dBm, @ 1dB Comp) | Dynamic Range NF(dB) | IP3(dBm) | Price \$ ea. (10 Qty.) |
|---------|--------------|-----------|----------------------------------|----------------------|----------|------------------------|
| ERA-1 | DC-8000 | 11.6 | 11.7 | 5.5 | 26 | 1.80 |
| ERA-1SM | DC-8000 | 11.0 | 11.3 | 5.5 | 26 | 1.85 |
| ERA-2 | DC-6000 | 14.9 | 12.8 | 4.7 | 26 | 1.95 |
| ERA-2SM | DC-6000 | 13.1 | 12.4 | 4.6 | 26 | 2.00 |
| ERA-3 | DC-3000 | 20.2 | 12.0 | 3.8 | 23 | 2.10 |
| ERA-3SM | DC-3000 | 19.4 | 11.5 | 3.8 | 23 | 2.15 |
| ERA-4 | DC-4000 | 13.5 | ▲17.0 | 5.5 | ▲32 | 4.15 |
| ERA-4SM | DC-4000 | 13.5 | ▲16.8 | 5.2 | ▲33 | 4.20 |
| ERA-5 | DC-4000 | 18.5 | ▲18.4 | 4.5 | ▲33 | 4.15 |
| ERA-5SM | DC-4000 | 18.5 | ▲18.4 | 4.3 | ▲32 | 4.20 |
| ERA-6 | DC-4000 | 11.3 | 18.5 | 8.4 | 36 | 4.15 |
| ERA-6SM | DC-4000 | 11.3 | 18.0 | 8.4 | 36 | 4.20 |

Note: Specs typical at 2GHz, 25°C

▲ Typ. numbers tested at 1GHz. At 2GHz, Max. Pwr. Out may decrease by 0.4dB & IP3 by 3 to 4dB.

* Low frequency cutoff determined by external coupling capacitors.

⊕ Price (ea.) Qty 1000: ERA-1 \$1.19, -2 \$1.33, -3 \$1.48, -4, -5 or -6 \$2.95. SM option same price.

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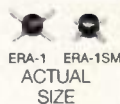
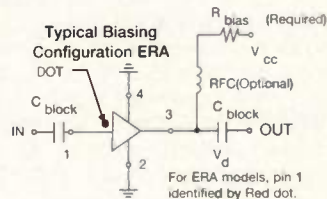
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|-------------|---|
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| 120x60 | .002, .047, .068, .1 μf |



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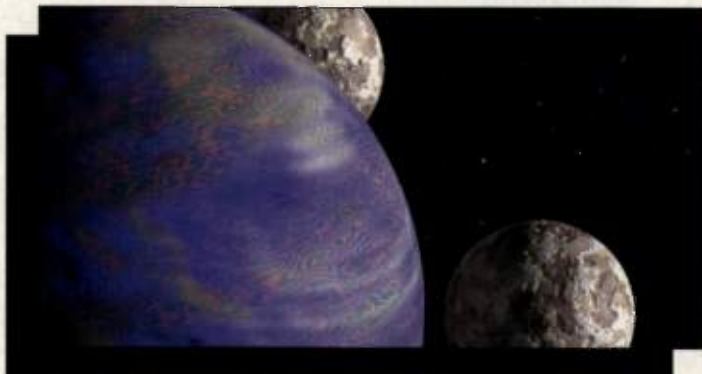
F 214 Rev B

featured technology — oscillators

22 Design methods of low-power crystal oscillators for wireless applications

The main considerations for designing low-power crystal oscillators for wireless applications focus on the power consumed by the oscillator and on its size. Included are different designs for the 170 MHz band, along with their main features and specifications.

—Jose Luis Jimenez Martin and Francisco Javier Ortega Gonzalez



cover story — p. 44

34 Design and validation of fault-tolerant, synchronized crystal oscillators

A new design and several verification techniques for synchronizing crystal oscillators achieve exact synchronization. Using SPICE, three simulation techniques are used to verify the design: oscillator startup, synchronization of out-of-phase oscillators and on-the-fly fault-injection.

—T. Kien Truong

cover story

44 PCS—working to make the link

The vendors at the PCS '96 trade show portrayed personal communications service (PCS) as the answer to all our future communications needs. But if you have been in the wireless industry for a while, you have a more realistic perspective of the technology, its implementation and its timetable.

—Ernest Worthman

tutorial

56 MRI basics and coil design principles

Magnetic Resonance Imaging (MRI) has become a critically important medical imaging technique during the past 20 years. The information includes a brief description of the relationship of these components. The balance discusses design details for high-field MRI RF receiver coils.

—David M. Peterson,
G. Randy Duensing, Ph.D. and J.R. Fitzsimmons, Ph.D.

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Coming in February

- CAD/CAE
- Resonator tutorial
- Cover story: DSP
- Product forum: Oscillators

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RF editorial

Come to RF Design '97 Conference & Exposition Sept. 10-12, 1997

By Don Bishop
Editorial Director

You're invited to join your engineering colleagues and vendor representatives in Santa Clara, CA, Sept. 10-12 for the *RF Design '97* Conference & Exposition. Those of us who are new to the magazine staff this year are looking forward to seeing you for the first time at our own conference, too.

Your participation, whether as a student in the *RF Design* Seminar Series of short courses, as a moderator or presenter during the technical sessions, as a visitor to the exhibition hall or in a combination of these activities, will allow you to take advantage of the focus that you and other readers bring to the science and art of designing RF components, products and systems.

Everyone has to strike some kind of balance between information-gathering and productivity. It's fine to read extensively, take courses and attend a lot of conferences, as long as the information you obtain helps you to fulfill your employer's and customers' requirements. Our intention is to use the combined resources of magazine publishing, instructional courses, the conference and exposition to provide the information you'll find the most useful. Actually, I'd like it to be compelling.

Every event needs performers and spectators. At engineering conferences, we call them presenters and attendees. Turn to page 86, and you'll find your invitation to submit an abstract for a technical paper to be presented at the conference and to be published in the conference proceedings. Some employers offer an incentive for this kind of participation; if you don't know whether yours does, it's time to ask. Whether

a tangible incentive is available or not, presenting a paper raises your visibility within your own company (and outside of it, too, though that's better whispered than shouted). It can help your company to promote its finished products as well as its product development or engineering service capabilities, which reflect well on the author. It helps you to meet other professionals who may help you to succeed with future projects. There are so many good reasons to submit a paper that I wonder why everyone doesn't.

Not really. Taking the time can be difficult. If it's difficult for you, there's still an important place for you.

Those who may not have the time to take advantage of submitting a paper still can benefit from hearing the presentations and from meeting their colleagues and vendor representatives. *RF Design '97* Conference & Exposition, along with the *RF Design* Seminar Series, gives you just that opportunity. Surveys that we've sent to you or perhaps to other readers confirm that, although there is some overlap between the RF and microwave disciplines, many engineers tend to work exclusively or primarily with RF designs, and many others look for such specialized information to help them with specific project assignments.

Take a moment to look at the suggested topics for papers. Send us an abstract for a subject you would like to cover, whether or not it is on the list. The conference's best ideas will come from you. Don't wait. You know there's an advantage in being first.

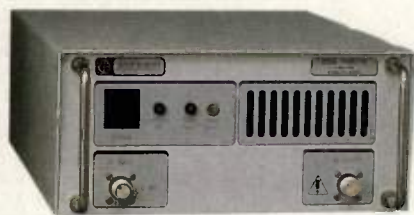


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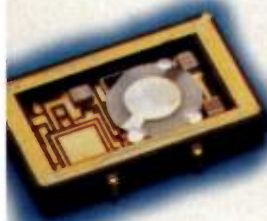
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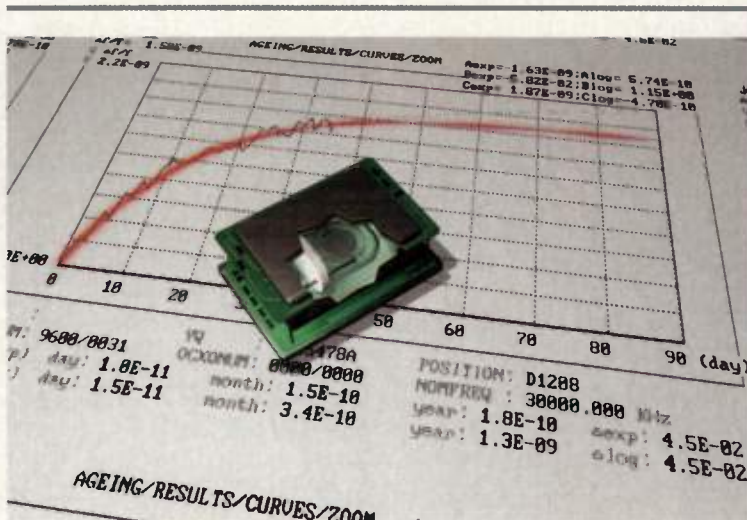
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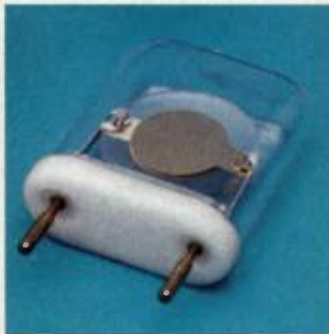


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|--------------|--------------|----------|------|----------|------|----|------|------------|------------------|----------------------|---------------|
| | | dB | ±KHz | dB | ±KHz | dB | ±KHz | | | | |
| 2 | TE5000 | 3 | 3.75 | 20 | 18.0 | - | - | 2 | 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 |

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21.4 MHz

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|--------------|--------------|----------|------|----------|------|----|------|------------|------------------|----------------------|---------------|
| | | dB | ±KHz | dB | ±KHz | dB | ±KHz | | | | |
| 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 |

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45.0 MHz

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|--------------|--------------|------|----------|------|----------|------|----|------------|------------------|----------------------|---------------|
| | | | dB | ±KHz | dB | ±KHz | dB | | | | |
| 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 | |

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70.0 MHz

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| | | | dB | ±KHz | dB | ±KHz | dB | KHz | | | |
| 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 |

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90.0 MHz

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| | | | dB | ±KHz | dB | ±KHz | dB | | | | KHz |
| 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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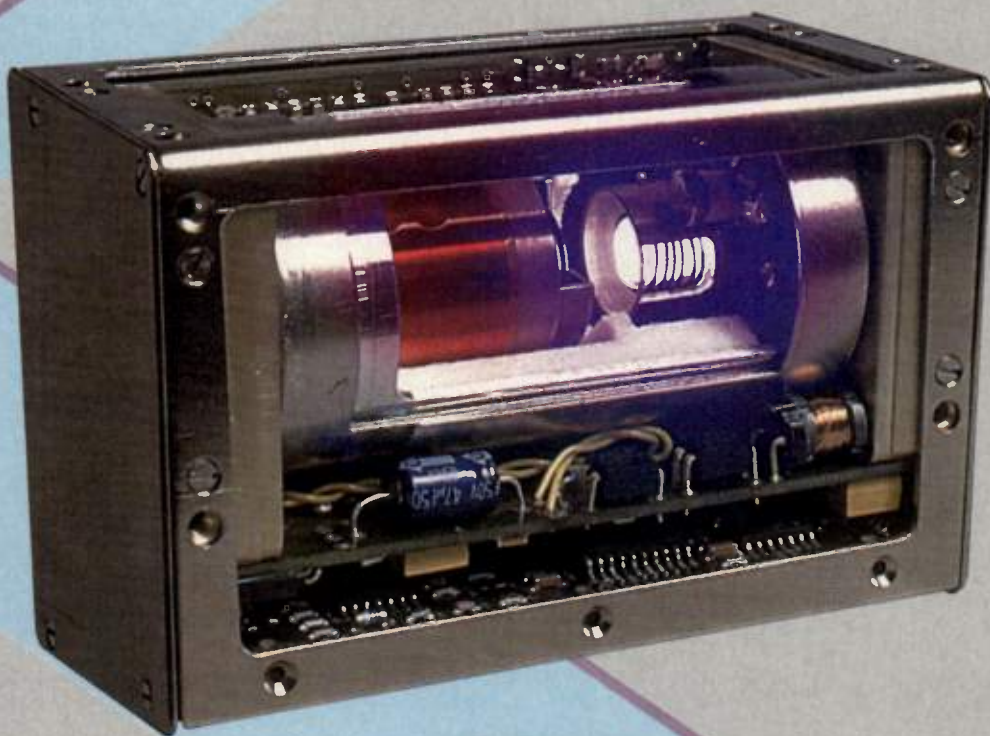
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| <input type="checkbox"/> | EMD40-1800L | Handsets | 1400-2000 | 8.0dB | 25dB |
| <input type="checkbox"/> | EMS-1X | Base Station | 10-1000 | 6.0dB | 30dB |
| <input type="checkbox"/> | ESMD-C1 | Base Station | 1-1000 | 6.5dB | 40dB |

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| <input type="checkbox"/> | ETC1.6-4-2-3 | Wireless | 500-2500 | 4:1 | 3 dB |
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| <input type="checkbox"/> | ETC9-1 | Wireless | 70-220 | 9:1 | 2.5 dB |

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|--------------------------|--------------|-------------|---------------|----------|----------------|
| Send me more info | Part Number | Application | Frequency MHz | Ratio | Insertion Loss |
| <input type="checkbox"/> | ESDC-7-2-75 | CATV | 5-800 | 7 ± 1 | 2.8 dB |
| <input type="checkbox"/> | EMDC-16-2-75 | CATV | 40-1000 | 16 ± 1.1 | 1.0 dB |
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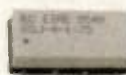
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RF letters

Letters should be addressed to the Editor, *RF Design*, 5660 Greenwood Plaza Blvd., Suite 350, Englewood, CO 80111. Letters published may be edited for length or clarity.

Weather conspiracy

Although I never expected conspiracy theories to penetrate to the hallowed editorial columns of *RF Design*, the Eugene Dusina letter published in the September issue fills a critical need that is little recognized in this country. Since the end of the cold war and the "demise" of the Russian bear, we have had no one to blame for lousy weather, hurricanes and picnic-ruining thunder-showers. The Dusina letter gives us hope.

R. Ellis
Las Vegas, NV

Praise for Weir

Just a short note to thank Jim Weir for the effort he put into writing the article on coaxial cables that appeared in the August issue.

I thoroughly enjoyed reading it and learned a few things as well. As manager of an ISO Guide 25 accredited calibration laboratory where we measure and generate signals up to 18 GHz on a daily basis, I can assure you that the non-mathematical common sense approach Mr. Weir adopted in writing the article, came as very welcome relief.

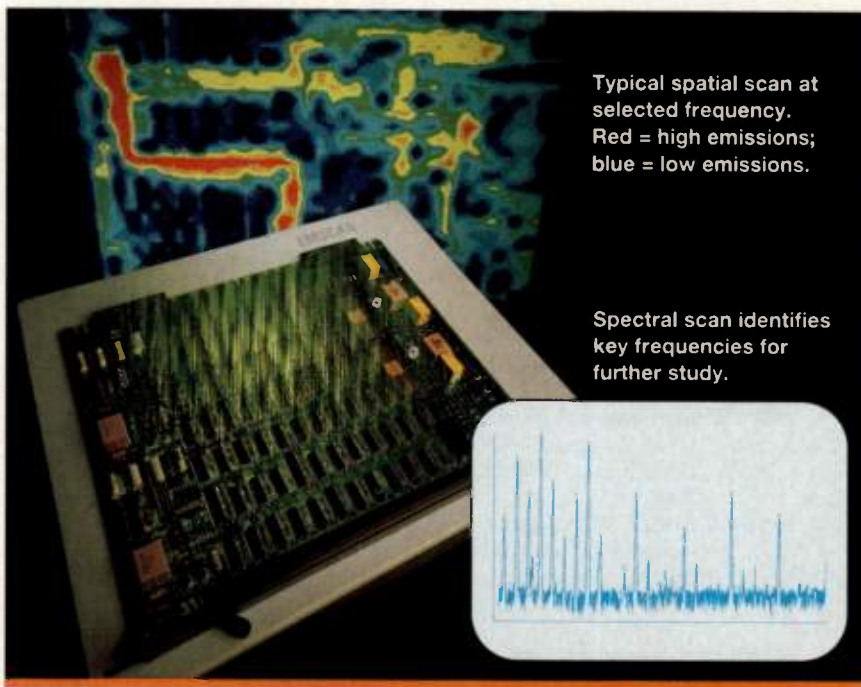
Keep up the good work. I will be forwarding the article to numerous metrologists, technicians and engineers for their benefit.

Eddie Tarnow
South Africa

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23-27 Nepcon West—Anaheim, CA. Information:
Reed Exhibition Company, 383 Main Ave.,
Norwalk, CT 06852-6059. Tel. 800-467-
5656; Fax 203-840-9656; Web site:
<http://www.nepcon.reedexpo.com>.

March 3-5 CTIA Wireless—San Francisco. Informa-
tion: Dobson and Associates. Tel. 202-463-
7905.

**13-19 CeBIT '97 World Business Center: Office,
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Hannover, Germany. Information: Mette
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17-20 European Design and Test Conference—
Paris. Information: Conference Secretariat,
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25-27 DSP World Spring Design Conference—
Washington, DC. Information: Dana Dowell,
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Web site <http://www.dspworld.com>.

**April 14-17 International Conference on Antennas and
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21-23 RF Design Seminar Series—Las Vegas.
Information: Intertec Presentations, 6300 S.
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80111. Tel. 303-220-0600; Fax 303-770-0253.

**22-24 International Wireless Communications
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**22-24 Convergence Tech and IC Expo for
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**23-26 Broadcast Technology—Jakarta,
Indonesia.** Information: Eileen Lavine,
Information Services, 4733 Bethesda Ave.,
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656-2942; Fax 301-656-3179.

**May 5-7 Vehicular Technology Conference
for cellular and mobile wireless communi-
cations—Phoenix.** Information: Wendy
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**6-8 Electronics Industries Forum of New
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**13-16 Computer and Communication Electronics
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**28-30 IEEE International Frequency Control
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tute of Standards and Technology, Time and
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ortegaw@boulder.nist.gov.

June 1-5 Supercomm—New Orleans. Information:
Telecommunications Industry Association.
Tel. 202-326-7300.

9-14 Asia Telecom—Singapore. Information:
Tom Dahl-Hansen, senior vice-president,
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730-6444; E-mail dahl-hansen@itu.ch.

**10-12 International Microwave Symposium and
Exhibition—Denver.** Information: Horizon
House. Tel. 617-769-9750.

**11-13 Virginia Tech Symposium on Wireless
Personal Communications—Blacksburg,
VA.** Information: Business Administrator,
Jenny Frank, Mobile and Portable Radio
Research Group, Virginia Polytechnic
Institute, 840 University City Blvd., Pointe
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24061-0350. Tel. 540-231-2958; Fax 540-
231-2968; E-mail hilda@vt.edu; Web site:
<http://www.ee.vt.edu/mprg/home.html>.

**July 14-17 Image Processing and Applications—
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Conference Organizer, Institution of
Electrical Engineers, Savoy Place, London
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**August 18-22 IEEE EMC Symposium on Electromagnetic
Compatibility—Austin.** Information: John
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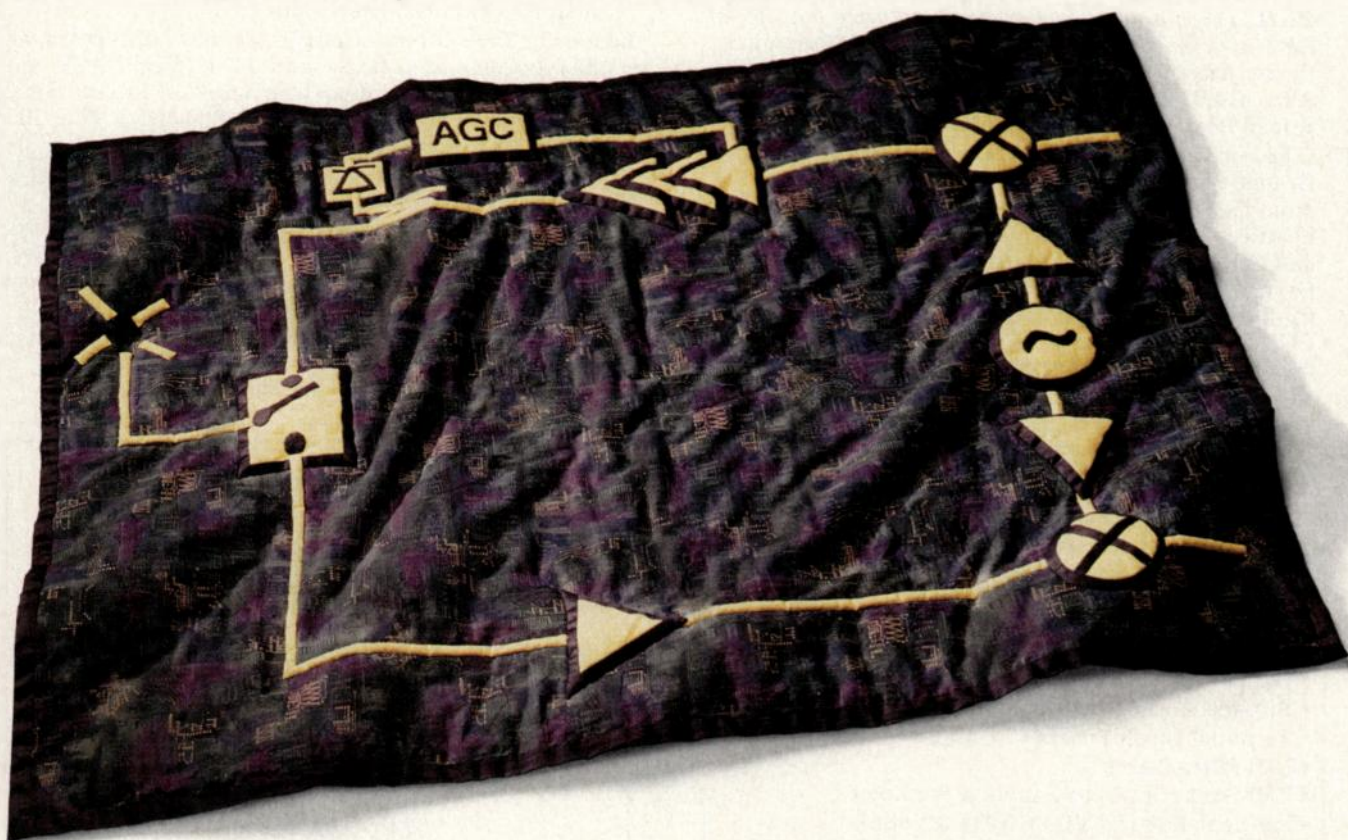
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Learning Tree International—*Wireless Networks and Mobile Communications*—Jan. 14–17, Feb. 18–21 Washington. Information: Learning Tree International, 1805 Library St., Reston, VA. Tel. 800-850-9197 or 703-709-9119; E-mail uscourses@learningtree.com; Web site <http://learningtree.com>.

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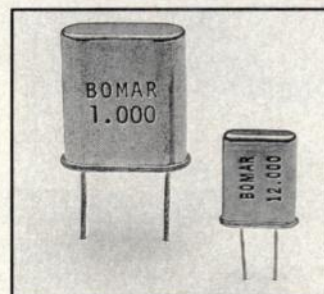
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UCLA—Digital Avionics Systems—Jan. 27–31;
HBT IC Technology for Communication Applications—
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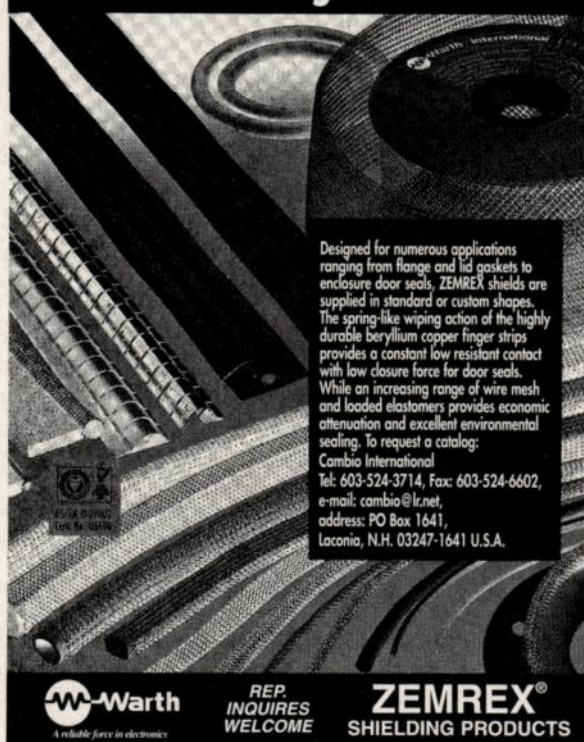
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Virginia Tech—Antennas: Principles, Design and
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St. Cloud, FL. Information: Kelly Brown, Northeast
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Memphis gets digital two-way system

Motorola has provided the city of Memphis with the Astro 800 MHz digital two-way communications system for better management of the city's public safety and other resources. Memphis' system is the first Astro 800 MHz public safety digital system in the state. The system includes 31 radio channels, 16 of which are dedicated to law enforcement and local government communications. The balance of the channels are assigned to fire department and other city agencies. The system provides communications coverage for some 410 square miles.

To efficiently manage communications traffic, the police and fire dispatch centers use 19 Centracom Series II Plus consoles with cathode-ray tube (CRT) displays that put the Astro digital system's features at the dispatchers' fingertips. Flashport software makes the system completely programmable. The system is compatible with those used by Memphis Light, Gas and Water; the Memphis International Airport; Federal Express; the city of Bartlett, Tennessee; Desoto County, Mississippi; and the Arkansas State Police.

NIST station signals increase power output

The Commerce Department's National Institute of Standards and Technology (NIST) will upgrade its radio station, WWVB, which broadcasts standard time and frequency signals. A new transmitter will increase the radiated power of the 60 kHz signal at least fourfold and will provide more reliable

coverage to the far corners of the continental United States, Mexico and Southern Canada.

WWVB signals can be used to set clocks to a few hundredths of a second. The signals also serve as a frequency reference with an uncertainty of one part in 10^{12} . The station's present output power of about 10 kilowatts broadcasts a signal strong enough to reach most of the continental United States, but it requires users at great distances from the transmitter to install bulky antennas for reliable service. The improvements, which will increase the station's power to well over 40 kilowatts, will make it possible to build automatic WWVB-controlled clocks into appliances and wristwatches. Resetting clocks after a power outage may become a thing of the past.

The new transmitting equipment is already on site at the station, located a few miles north of Ft. Collins, CO. NIST is a non-regulatory agency of the Commerce Department's technology administration. The agency works with industry to develop and apply technology, measurements and standards. News and general information on NIST is available on the World Wide Web at <http://www.nist.gov>.

Contracts:

Brady selected as strategy partner of Motorola—Brady U.S.A. has been selected as a global strategic partner of Motorola Indala, a manufacturer of radio frequency identification (RFID) systems. Brady will market these RFID products and services, specializing in the sale of auto ID systems designed for harsh environments.

RFID uses radio waves to transmit a unique identifier number instead of using light waves as are needed with bar codes, which are sensitive to environmental conditions. The low-frequency radio signal can be read through any non-metallic debris such as dirt, ice, or paint. RFID labels and tags are passive, meaning they require no battery. Applications are diverse, ranging from tagging valves in chemical plants to tagging cattle in feed lots.

Ericsson to use Teradyne's mixed-signal test system—Ericsson Components, Stockholm, Sweden, has selected Teradyne's A575 advanced mixed-signal test system for engineering characterization and production testing of its next-generation RF integrated circuits. Teradyne's mixed-signal microwave test system can test all of the Ericsson devices, which are becoming increasingly integrated with RF and digital functions on the same integrated circuit.

Three contracts for Neulink products—RF Industries' Neulink Telemetry Division, San Diego, has received three contracts, together valued at more than \$800,000, for wireless digital data products. Two of the contracts are for Neulink's 9600 transceiver modems. The first contract will use 1,000 modems in a supervisory control and data acquisition (SCADA) network. The second contract involves global positioning satellite (GPS) applications for location and tracking of recreational vehicles. The third contract is an add-on order for AM and FM and UHF receivers that will be used for the emergency alert system (EAS).

Business Briefs

DTI grant aids Bridlington manufacturing—The Department of Trade and Industry (DTI) awarded £200,000 in regional selective assistance to help K&L Microwave create a base in Bridlington, Yorkshire, United Kingdom and to bring 40 new jobs into the area over the next two years. The company manufactures electrical components for commercial and military communications systems.

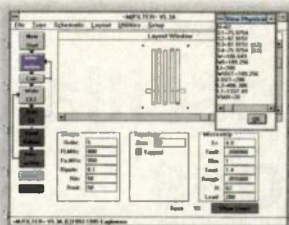
Services company created—Mobile Systems International (MSI) has created a wireless services company. MSI Services, Richardson, TX, will provide services to the cellular, paging and personal communications services (PCS) markets. Services will include site acquisition, gen-

eral construction, project management, network engineering, facility engineering, civil engineering, drive testing, system optimization and microwave relocation.

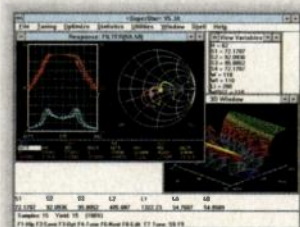
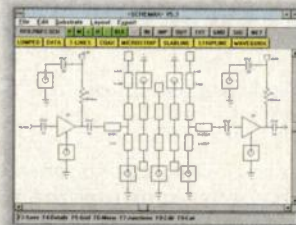
Rosenberger establishes facility—Rosenberger Hochfrequenztechnik, European manufacturer of coaxial connectors and microwave components, has opened a sales, engineering support and distribution facility in Lancaster, PA. Rosenberger of North America is a subsidiary of the privately-held, Tittmoning, Germany-based company. The Lancaster facility will warehouse and distribute products made overseas and will assemble custom coaxial cable assemblies.

DESIGN FROM START-TO-ART

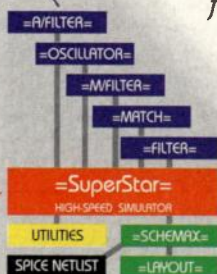
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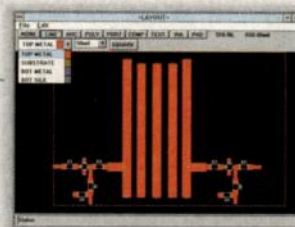


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Design methods of low-power crystal oscillators for wireless applications

By Jose Luis Jimenez Martin and Francisco Javier Ortega Gonzalez

The main considerations for designing low-power crystal oscillators for wireless applications focus on the power consumed by the oscillator and on its size. Included are different designs for the 170 MHz band, along with their main features and specifications.

The design of low-power crystal oscillators for wireless applications is of immediate interest for the RF equipment designer. Crystal oscillators are used for pagers [1,2,3] based on direct-conversion [4,5] and for superheterodyne receivers for personal commu-

nications services (PCS). A direct-conversion receiver for pagers is one of the best examples of a low-power crystal oscillator application. A classic direct-conversion scheme is shown in Figure 1. As with many other wireless systems, this kind of wireless receiver needs crystal oscillators with the following features:

- High stability in a wide temperature range (± 5 ppm, -10 to 60°C).
- Low-power consumption (< 0.5 mW typically).
- Low aging (< 1 ppm in a year).
- Low size (not many parts).
- Low pushing figure.

The design of crystal oscillators with these requirements and oscillator measurements taken over four topologies are analyzed as follows.

Tested oscillators

The oscillators shown in Figure 2 have been designed, constructed and

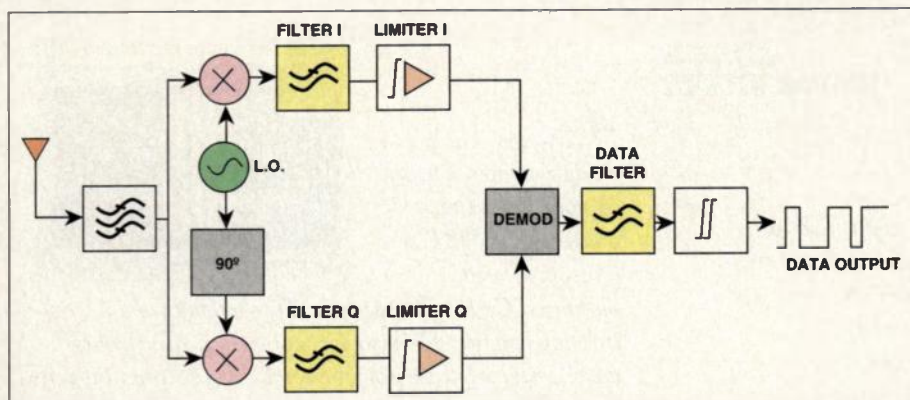


Figure 1. Direct conversion diagram block.

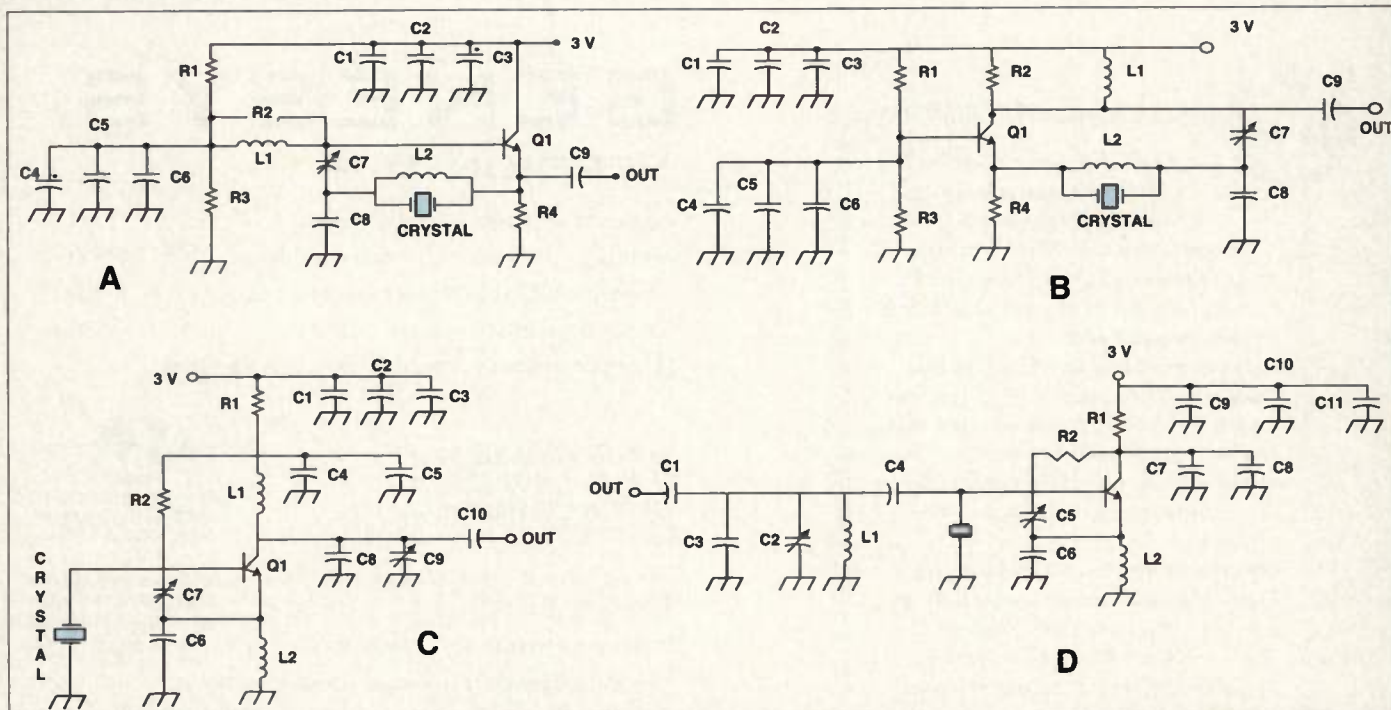
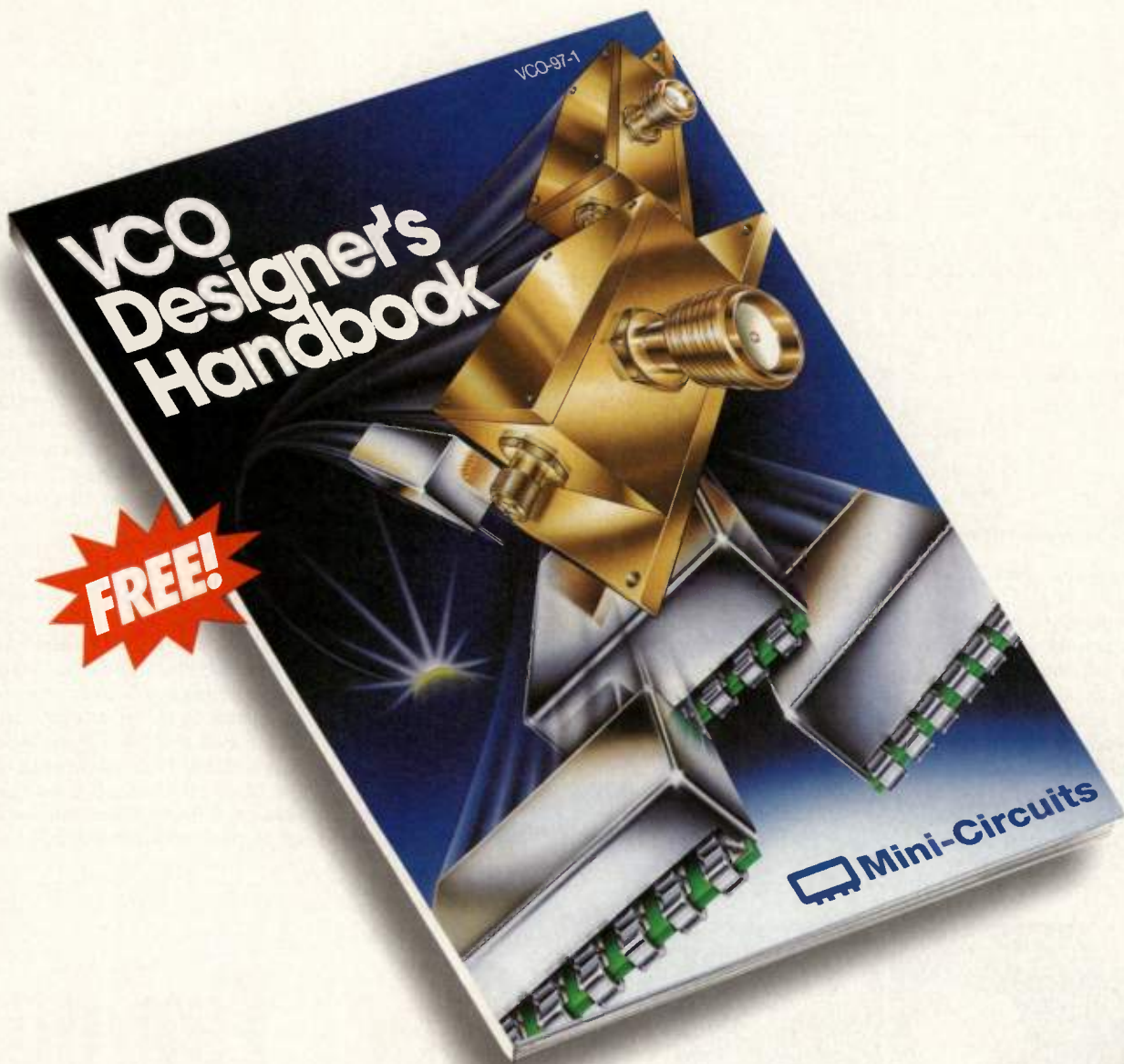
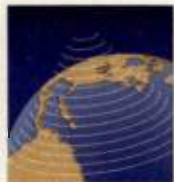


Figure 2. Topologies of the oscillators analyzed. a) common collector Butler oscillator, b) common base Butler oscillator, c) harmonic Colpitts, output at the collector, d) harmonic Colpitts, output at the base.



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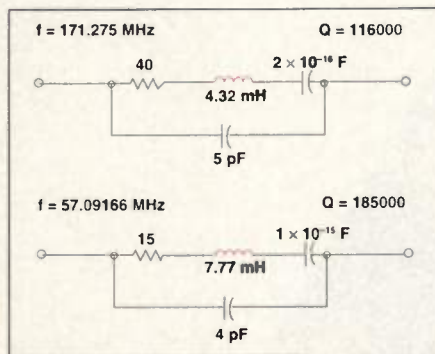


Figure 3. Equivalent circuits of the crystals used.

measured to work in the 171.25 MHz and 171.75 MHz frequency bands. Three prototypes of each oscillator have been constructed and measured to verify repeatability. Five quartz crystals of the desired frequency have been tested with each unit of the oscillators. This way, verification of the results are a result of the topology and not a result of the constructed circuit or quartz crystal.

The designed oscillators are the

following:

- Butler common-collector oscillator.
- Butler common-base oscillator.
- Colpitts tuned-to-harmonics oscillator, output at the collector.
- Colpitts tuned-to-harmonics oscillator, output at the base.

The objective is to compare the performance of these topologies using crystals working at the fundamental frequency and crystals working at the third overtone.

Figure 3 shows the equivalent circuit of the crystals. The crystal used for the Butler oscillator is cut to work in the series resonant mode (7th overtone). The crystal used for the Colpitts oscillators is cut to work in the parallel mode with a load capacitance $C_L = 8$ pF (3rd overtone). The series equivalent resistance is low.

Stability of the output frequency

Many wireless receivers exhibit an important deterioration of performance for frequency deviations of the local oscillator. In the direct-conversion re-

ceivers for pagers, this deviation gives an important increase of the errors of received data [6]. In this case, with a deviation of 4.5 kHz, there is a theoretical limit of 26 ppm until total failure of the receiver. For a variation of only 17 ppm, the reduction in sensitivity is 3 dB. When a carrier of 470 MHz is used (some paging services use this band), the previous values must be reduced to 9 ppm and 6 ppm, respectively.

Thus, the stability of the oscillator must be lower than 10 ppm. This stability is the result of the variations caused by temperature changes and by the aging of the crystal oscillator. Usually, the temperature range for commercial use is from -10°C to 50°C , and the "expected life" depends on the system. Nevertheless, a system for a commercial application never will be used both at extreme temperatures and at the end of its life simultaneously. It is estimated that a tolerance of 10 ppm for the -10°C to 50°C range will be enough. The temperature coefficient of the oscillator is defined in Equation 1.

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where:

- f = oscillation frequency
- X_n = parts of the crystal oscillator: XTAL, passive components (capacitors, resistors, coils), parts of the transistor's model and others.
- T_o = Reference temperature, 20 to 25°C typically
- $\partial X_n / \partial T$ = temperature coefficient of the oscillator's part (not normalized)
- $(1/f_o) \cdot (\partial f / \partial X_n)$ = sensitivity of the oscillator's output frequency (for different parts)

The "sensitivity" analysis of the output frequency vs. the most relevant passive components is shown in Table 1. The sensitivity of the Butler oscillators vs. the tuned tank (C7, C8, L1) and the inductance L2 (to tune the capacitance of the package) is shown. The sensitivity of the Colpitts oscillators vs. the feedback capacitors ((C6 and C7) or (C5 and C6)) and the coil L2 also is shown.

The sensitivity of the Butler oscillators is higher than the sensitivity of the Colpitts oscillators, but this sensitivity is not that high in any case.

The part elected to adjust the oscillator is the part

that affects the output frequency most. The temperature coefficient of the surface-mount (0805) capacitors used is ± 30 ppm/°C. For the surface-mount (1206) coils used, the temperature coefficient is 500 ppm/°C. For temperature changes suffered by the passive parts, a frequency change lower than 0.5 ppm was achieved for all the oscillators tested.

This leads to a negligible dependency of the output frequency on the passive components compared to dependency on the quartz crystal; therefore, maximum attention must be paid to the quartz crystal.

Phase noise

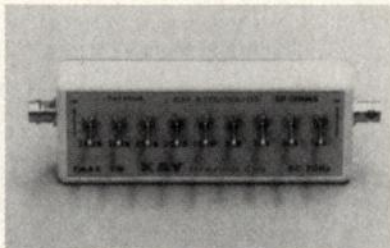
The main effects of phase noise in a receiver are the reduction of the sensitivity and the decrease of adjacent channel rejection. The second effect (adjacent channel rejection) usually is more important than the first (reduced sensitivity). A

TABLE 1

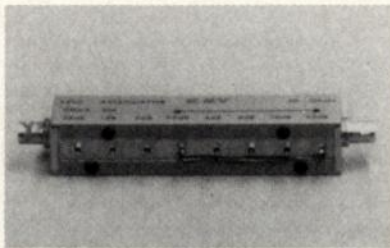
| BUTLER COMMON COLLECTOR | | BUTLER COMMON BASE | | HARMONIC COLPITTS OUTPUT AT THE COLLECTOR | | HARMONIC COLPITTS OUTPUT AT THE BASE | |
|-------------------------|-------------|--------------------|-------------|---|-------------|--------------------------------------|-------------|
| C7 | 0.58 ppm/% | C7 | 0.29 ppm/% | C6 | 0.14 ppm/% | C5 | 0.26 ppm/% |
| C8 | 0.17 ppm/% | C8 | 0.11 ppm/% | C7 | 0.31 ppm/% | C6 | 0.09 ppm/% |
| L1 | 1.47 ppm/% | L1 | 1.02 ppm/% | L2 | 0.049 ppm/% | L2 | 0.053 ppm/% |
| L2 | 0.025 ppm/% | L2 | 0.029 ppm/% | L1 | 0.017 ppm/% | L1 | 0.020 ppm/% |

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| 1/4550 | 50Ω | DC-500MHz | 0-16.5dB | .1dB Steps |
| 4560 | 50Ω | DC-500MHz | 0-31dB | 1dB Steps |
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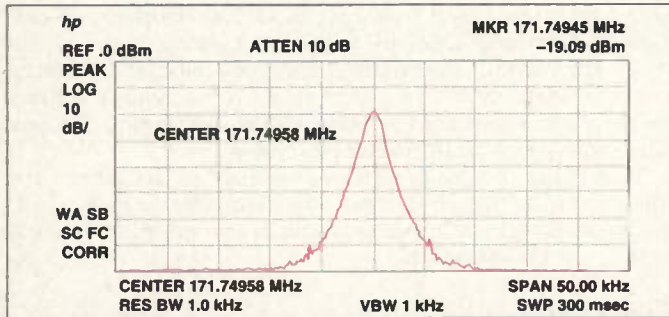


Figure 4a. Phase noise plot of the common-base Butler oscillator at 171.750 MHz.

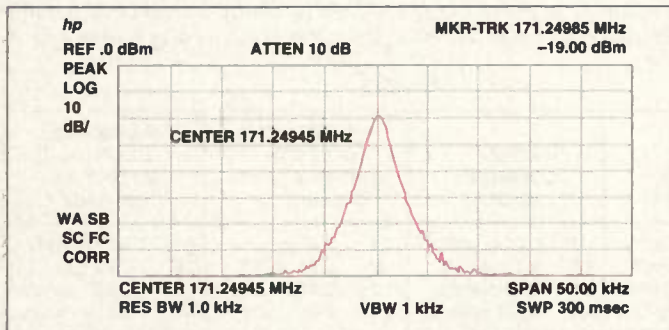


Figure 4b. Phase noise plot of the oscillator at 171.250 MHz.

direct-conversion receiver for paging needs an adjacent channel rejection of 65 dB for a channel separation of 25 kHz [7]. This means a phase noise less than -105 dBc at 25 kHz from the central frequency of the carrier. It is not too difficult to get this specification for a low-power crystal oscillator.

The phase noise of an oscillator is [8]:

$$\mathcal{L}(f_m) = \frac{1}{2} \left(1 + \frac{\omega_o^2}{4\omega_m^2 Q^2} \right) \left(1 + \frac{\omega_c}{\omega_m} \right) \frac{FkT_o}{P_{sav}} \quad (2)$$

where:

f_m = frequency deviation

f_o = fundamental frequency of the oscillator

Q = quality factor (loaded) of the oscillator

f_c = flicker cutoff frequency (of the transistor)

F = noise figure of oscillator (large signal mode)

k = Boltzman constant

T_o = noise temperature at the input of the oscillator

We will use this equation to analyze a typical example. If we are working with an oscillator of $f_o = 170$ MHz, $f_m = 2$ kHz, $f_c = 10$ kHz and $(kFT_o)/P_{sav} = -7$ dBc/Hz, we will need: loaded quality factor (Q) > 50. This value is lower than the value obtained for typical quartz crystals.

Simulation has proved that this quality factor is equal to the quality factor of the crystal multiplied by 0.5 or by 0.6. It is not necessary to pay special attention to this point. Figure 4 shows the phase noise obtained from the tested oscillators.

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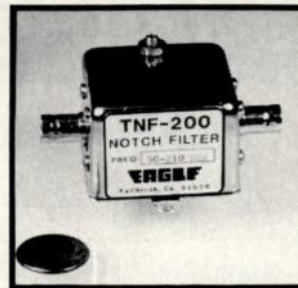
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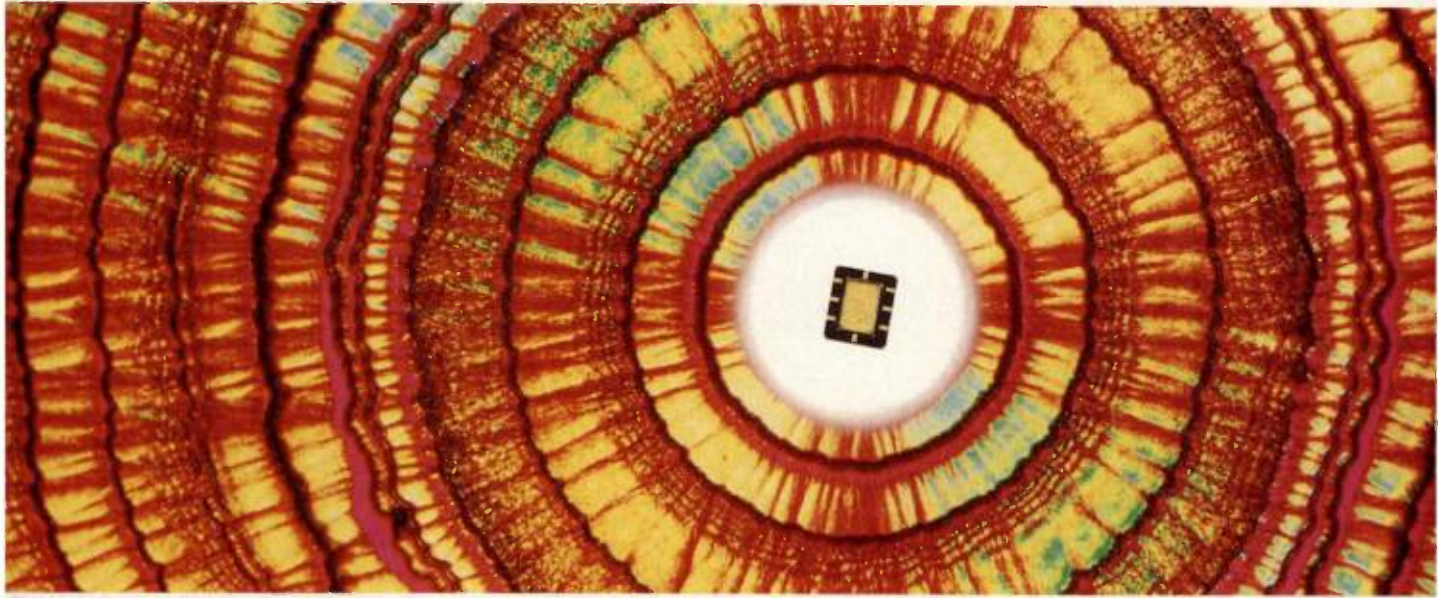
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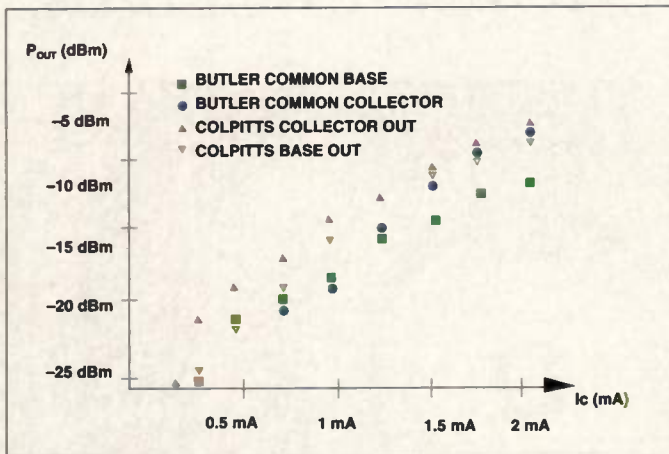


Figure 5a. Output power (of the oscillators) vs. bias current.

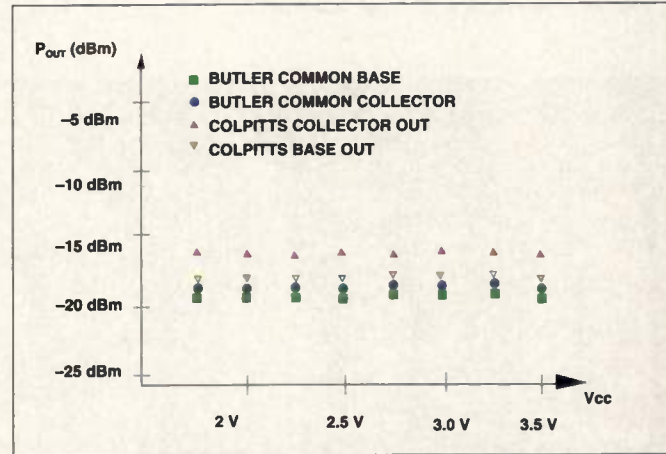


Figure 5b. Output power (of the oscillators) vs. supply voltage.

TABLE 2

| HARMONIC | BUTLER COMMON COLLECTOR | BUTLER COMMON BASE | HARMONIC COLPITTS OUTPUT AT THE COLLECTOR | HARMONIC COLPITTS OUTPUT AT THE BASE |
|-------------|-------------------------|--------------------|---|--------------------------------------|
| FUNDAMENTAL | -14.3 dBm | -15.7 dBm | -16.5 dBm | -18.4 dBm |
| 2° | -9.9 dBc | -19.7 dBc | -28.4 dBc | -29.2 dBc |
| 3° | -16.1 dBc | -30.0 dBc | -48.0 dBc | -44.1 dBc |
| 4° | -25.4 dBc | -38.4 dBc | -63.6 dBc | -67.0 dBc |
| 5° | -34.5 dBc | -49.2 dBc | -81.2 dBc | -88.4 dBc |

Power at the fundamental frequency

Most wireless receivers use downconverters based on active mixers using the classic "Gilbert topology," so it is possible to save power working in the linear region where the gain is proportional to the level of the oscillator [9].

Sharing the gain over the different stages of the receiver, it is possible to use oscillator power levels lower than -20 dBm. This means that small currents and voltages must be used to feed the oscillator.

Even so, the minimum current is limited by the minimum transconductance needed to start the oscillation. This effect is important in Butler oscillators working with quartz crystals in high overtones (7th). In this case, a high collector current is needed to compensate for the series resistance of the quartz crystal and to start oscillations. Suppose a typical oscillator efficiency of 2.5–5% and an output power of -17 dBm is needed for a direct-conversion power of 0.4–0.8 mW. This means a current of 0.3–0.15 mA is required if the receiver is operated at 3 V.

As a result, the effect of amplitude-limiting is caused by the limit of current, not by the voltage (V_{ce}) limit [10].

It follows that the output power depends on the collector current, not on the voltage of the battery. This is shown in Figure 5.

Table 2 shows the value of the output power of the four oscillators analyzed. *Fundamental* means "third overtone" for the Colpitts oscillators.

Power at harmonic and subharmonic frequencies

The harmonics and subharmonics of the signal generated by the oscillator cause interference with non-desired signals. This interference is limited by the attenuation presented by the RF stages. Usually, these stages are tuned to the desired frequency. This rejection may be expressed as:

$$\text{Attenuation} = \sqrt{1 + \left(\frac{2Q\Delta f}{f_o} \right)^2} \quad (3)$$

where:

Δf = frequency deviation

f_o = receiver frequency

Q = quality factor of the tuned tank (loaded)

For example, a maximum amplitude of -30 dBc is enough for a pager receiver based on a direct-conversion

scheme. From Table 2, it is easy to see that the Butler oscillators do not satisfy the specifications. This problem can be solved using a new filter, but this is bulky and more expensive, and more space is needed for the printed circuit board (PCB).

The Colpitts oscillators satisfy the specifications, but the safety margin is not high, especially for the second harmonic (the most dangerous). To decrease the level of this harmonic, increase the Q of the coil L_1 , which is the limiting factor of the Q of the tank. For example, this Q can be increased using high- Q , air-core inductors for surface mounting.

Power consumption

Wireless circuits usually are powered by batteries. For example, pagers consume about 2 mW, which ensure more than 800 hours of battery life. Pagers based on direct-conversion technologies consume even less power. In this kind of receiver, most of the power is consumed by the local oscillator; therefore, a reduction in the power consumption of the oscillator means an important increase in the batteries' life. Recognizing this fact, we will set a maximum power consumption of 0.1 mW for the oscillator.

This decision means a limit of 0.1 mA of current from a 3 V battery (or 0.34 mA for a 1.5 V battery). This current is not enough for a Butler oscillator. A Colpitts oscillator can work with this collector current. Experiments have determined that the Butler oscillators need 50% or 75% more current than the Colpitts oscillators.

Some authors recommend the use of auxiliary-bias circuits (using diode

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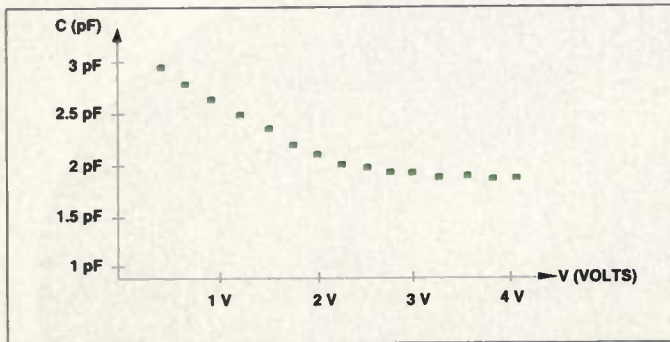


Figure 6. Changing of the parasitic capacitances of the transistor vs. voltage.

and transistors) to improve the behavior of crystal oscillator [10]. This practice usually is impossible to follow with low power wireless crystal oscillators because it increases the power consumption too much. In the oscillators shown, the bias point has been stabilized using resistors and negative feedback. This negative feedback is introduced by R4 in the Butler oscillators and by R1 and R2 in the Colpitts oscillators

Oscillator pushing

An important factor for an oscillator using batteries is oscillator pushing. The problem manifests itself as the battery voltage falling during its discharge period. The change of the output frequency of the oscillator can be expressed as:

$$\Delta f = P \cdot \Delta V_{cc}$$

where:

Δf = frequency deviation

P = pushing

ΔV_{cc} = change in battery voltage

The "pulling" mainly results from the change of the parasitic capacitances with the voltage applied to the transistor. For the Butler oscillator, this change of the capacitances produces a phase lag in the transistor. To compensate, it is necessary to change the frequency of the quartz crystal.

The same effect in the Colpitts oscillator can be compensated by adjusting C and C5 for base and collector outputs, respectively. The change of the capacitances is more important when working with low voltage, such as the voltage of most wireless crystal oscillators. (See Figure 6.)

To fight pushing in circuits without voltage regulators, it is necessary to select capacitor values as large as possible in the feedback circuit (compared to the parasitic capacitances of the transistor). This is easy to do with Colpitts oscillator working at the fundamental frequency of the crystal. It is more difficult to do with Butler oscillators working at high overtones, because the needed capacitors are small.

Table 3 shows the measured values of pushing and pulling figures of the tested oscillators. The best pushing figures belong to the Colpitts oscillator with output at the base. That is a logical result because the common collector Colpitts oscillator output is located in parallel with the quartz crystal. In the common-base Butler oscillator, the signal is extracted from the overtone-tuned tank; therefore, the changes of the load effect to the phase-lag introduced by the transistor changes the output frequency.

For the Colpitts oscillator with the output at the collector, the signal is extracted from a point where the RF impedance

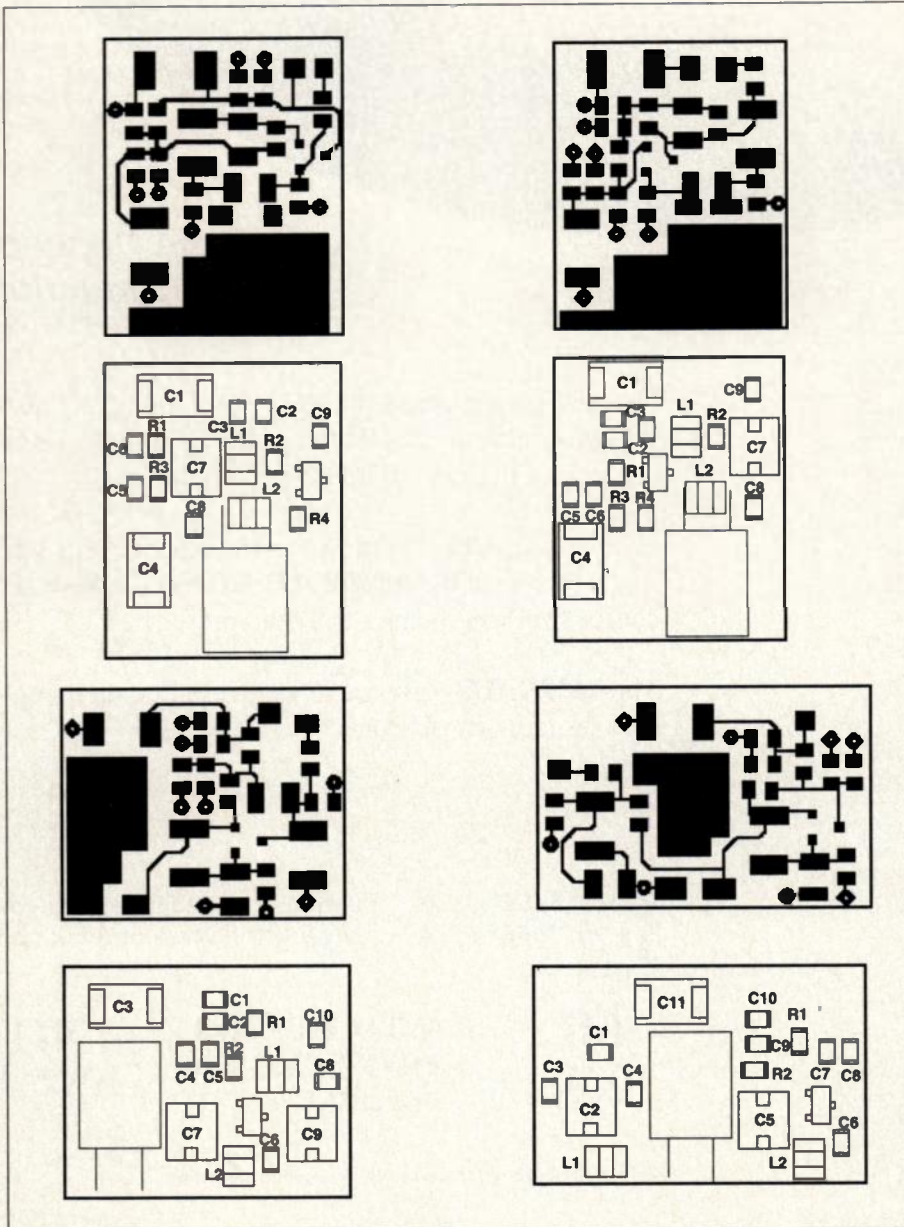


Figure 7. Oscillator's output.

nce is low; therefore, the changes of the load do not affect the oscillation frequency.

Parts and area

In wireless applications, it is important to get a small PCB. To do that, one can use 0603 surface-mount passive parts; small packages for transistors (SOT23 or smaller); or thin PCB lines (10 mils or thinner). The optimization of the layout must not overrule the basic principles of the RF design, that is, minimized coupling and good grounding. Figure 7 shows the layout of the PCB designed to test the crystal oscillators. Figure 8 shows the designed and tested prototypes.

Adjustment

The crystal oscillators must be adjusted to compensate for the tolerance of the quartz crystals. The lower the tolerance, the higher the cost of the quartz crystal. According to our tests, the Butler oscillators are the hardest to adjust. This negative effect is more important for the common-base Butler oscillator. This oscillator exhibits a critical adjustment. These difficulties have been related by others, especially for the common-base oscillators [11]. Usually, it is necessary to order at least two or three quartz

| TABLE 3 | | | | |
|--------------------|-------------------------------|--------------------------|--|---|
| | BUTLER COMMON COLLECTOR | BUTLER COMMON BASE | HARMONIC COLPITTS OUTPUT AT THE COLLECTOR | HARMONIC COLPITTS OUTPUT AT THE BASE |
| PUSHING PULLING | 0.9 ppm/V 0.83 ppm | 0.97 ppm/V 0.87 ppm | 0.3 ppm/V 0.08 ppm | 0.035 ppm/V 1.2 ppm |

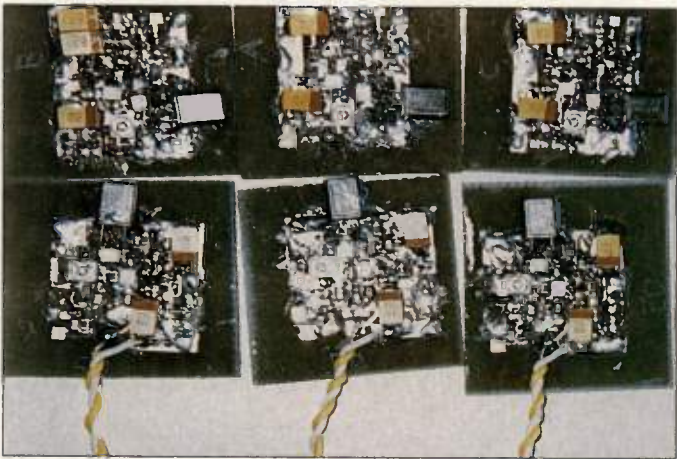


Figure 8. Photographs of the oscillators.



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crystals during the design process.

On the other hand, the Colpitts oscillators have been rugged and easy to adjust over a wide range of frequencies.

Table 4 compares the main features of the proposed and tested oscillators.

RF

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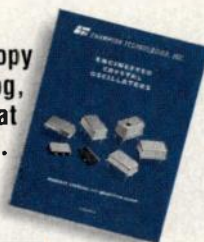
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| TABLE 4 | | | | |
|----------------------|-------------------------------|--------------------------|--|---|
| | BUTLER COMMON COLLECTOR | BUTLER COMMON BASE | HARMONIC COLPITTS OUTPUT AT THE COLLECTOR | HARMONIC COLPITTS OUTPUT AT THE BASE |
| STABILITY | GOOD | GOOD | GOOD | GOOD |
| PHASE NOISE | GOOD | GOOD | GOOD | GOOD |
| POWER AT FUNDAMENTAL | BAD | BAD | GOOD | MEDIUM |
| POWER AT HARMONICS | BAD | BAD | GOOD | GOOD |
| POWER CONSUMPTION | VERY BAD | VERY BAD | GOOD | GOOD |
| PUSHING | MEDIUM | MEDIUM | GOOD | GOOD |
| PULLING | BAD | BAD | GOOD | BAD |
| SIZE | GOOD | GOOD | GOOD | GOOD |
| ADJUSTMENT | BAD | VERY BAD | GOOD | GOOD |



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Design and validation of fault-tolerant, synchronized crystal oscillators

By T. Kien Truong

A new design and several verification techniques for synchronizing crystal oscillators achieve exact synchronization. The new design uses minimum part count for ultra-reliability. It is suitable for high-speed and on-chip integration. The new design improves upon existing architectures from capacitive coupling to feedback voting and finally, to full-fault tolerance. Using SPICE, three simulation techniques are used to verify the design: oscillator startup, synchronization of out-of-phase oscillators and on-the-fly fault-injection.

In commercial aviation, defense, aerospace and nuclear power plants where safety is critical, redundancy has been used extensively to improve the reliability of control and computing systems. In a fault-tolerant system, redundant computing channels depend on clock synchronization. All data distribution and comparing or voting are based on synchronous clock edges among channels. Crystal oscillators provide highly stable clock signals, which are required by the redundant processors. The simplest way to synchronize the redundant systems is to use a single crystal oscillator for a common

timing reference. This approach carries with it the penalty of a single-point failure that can "wipe out" the whole system.

In high-performance computers for real-time applications, many issues must be resolved in a practical fault-tolerant clock: continuous clocking without any temporary stoppage or slow-down to establish synchronization; high speed to support processing power; fault tolerance including arbitrarily malicious faults; complete fault coverage for all fault-containment regions; tight synchronization to subnanosecond range; startup reset synchronization so that the processors are not n-clocks apart; and minimum parts count for high reliability.

A crystal oscillator can produce many failure modes: amplitude error, frequency drift, phase shifting and jitters, duty cycle variation, glitches, runt-pulse transients and Byzantine fault.

A number of fault-tolerant clock designs have been described in literature. Most of them use phase-locked loops (PLLs) [1, 2, 3]. Some use standby switching spares [4]. Some periodically realign the oscillators after a free drifting period. Most designs require four redundant oscillators to tolerate a single malicious fault, and most have the drawbacks of being low-speed and loosely synchronized [5, 6, 7]. A patented design uses only three oscillators with minimum hardware to achieve fault tolerance and exact synchronization to less than 0.25 nanoseconds.

Independent oscillators

In Figure 1, the amplifier for the crystal oscillator is a complementary metal oxide semiconductor (CMOS) inverter that is biased at its mid-range to function as a linear amplifier for the clock signal. To oscillate, the circuit must satisfy the Barkhausen criteria that the phase shift around the loop be $n360^\circ$ and that the loop gain exceeds

unity at the resonant frequency.

We can model the quartz crystal as having a motional resonance arm R_n , L_m and C_m in parallel with the crystal-holding capacitance (CP) (the mounting electrodes). All circuit parasitic elements, on chip or off chip, are lumped into the loading capacitors and loading resistors.

In a series resonant oscillator, the crystal operates at its natural series frequency. In a parallel resonant oscillator, the feedback circuit introduces load capacitance to the crystal and causes the crystal to operate at the frequency at which the crystal reactance cancels the load-capacitor reactance. It follows that, when the load capacitance changes, the resonant frequency changes. The crystal reactance curve crosses zero at two frequencies that represent the series resonant and parallel resonant. Between the two frequencies, the reactance is inductive. Below and beyond that range, the reactance is capacitive. The oscillator operates in the inductive region between the two resonant frequencies. Within this bandwidth, the crystal phase varies by a large range with a small change in frequency. In other words, only an extremely small frequency shift is necessary to change the crystal's impedance to compensate for phase deviation around the loop.

Program 1 shows the SPICE deck of three independent phasing oscillators with a typical frequency deviation within 100 ppm among the crystals. The transistor parameters are from typical 2-micron CMOS technology. A triple-stage inverting amplifier is used to obtain higher gain for speeding up the simulation time for the startup transient. The CP is also commented out to speed up the simulation without affecting the fault-tolerant behavior of the circuit.

Even though crystal oscillator simulation is difficult because a quartz crystal has an extremely high-Q circuit

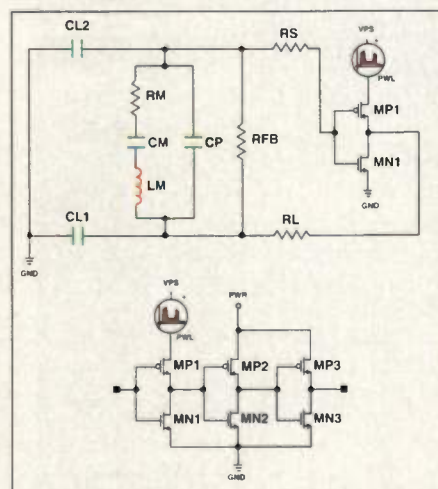


Figure 1. Crystal oscillator using CMOS inverter.

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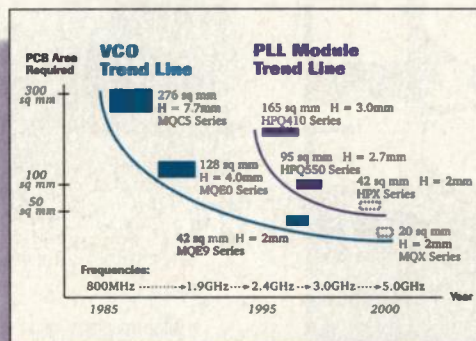
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PROGRAM 1

* Inverting amplifier

```
.SUBCKT INVAMP VCC VPS 1 4
MN1 2 1 0 0 NMOS L=2U W=2U
MP1 2 1 VPS VPS PMOS L=2U W=8U
MN2 3 2 0 0 NMOS L=2U W=5.43U
MP2 3 2 VCC VCC PMOS L=2U W=21.74U
MN3 4 3 0 0 NMOS L=2U W=14.76U
MP3 4 3 VCC VCC PMOS L=2U W=59.1U
RFB 4 1 22E6
.ENDS
```

* 24 MHz crystals

```
.SUBCKT XTAL24A 1 2
RM 2 4 10.5
CM 3 4 20.60E-15
LM 1 3 2.14E-3
*CP 1 4 4.5E-12
.ENDS
```

```
.SUBCKT XTAL24B 1 2
RM 2 4 10.7
CM 3 4 20.50E-15
LM 1 3 2.150E-3
*CP 1 4 4.5E-12
.ENDS
```

```
.SUBCKT XTAL24C 1 2
RM 2 4 10.0
CM 3 4 20.44E-15
LM 1 3 2.156E-3
*CP 1 4 4.5E-12
.ENDS
```

* Oscillator circuit

```
.SUBCKT OSCA AMPIN AMPOUT
XAMP VCC VPS AMPIN AMPOUT INVAMP
XTAL NXIN NXOUT XTAL24A
RS AMPOUT NXIN 200
RL NXOUT AMPIN 200
CL1 NXIN 0 10PF
CL2 NXOUT 0 10PF
VC5 VCC 0 DC 5V
V99 VPS 0 PWL ON 5V .1US 5V .105US 0V .110US 5V 1MS 5V
.ENDS
```

```
.SUBCKT OSCB AMPIN AMPOUT
XAMP VCC VPS AMPIN AMPOUT INVAMP
XTAL NXIN NXOUT XTAL24B
RS AMPOUT NXIN 200
RL NXOUT AMPIN 200
CL1 NXIN 0 10PF
CL2 NXOUT 0 10PF
VC5 VCC 0 DC 5V
V99 VPS 0 PWL ON 5V .11US 5V .115US 0V .120US 5V 1MS 5V
.ENDS
```

```
.SUBCKT OSCC AMPIN AMPOUT
XAMP VCC VPS AMPIN AMPOUT INVAMP
XTAL NXIN NXOUT XTAL24C
RS AMPOUT NXIN 200
RL NXOUT AMPIN 200
CL1 NXIN 0 10PF
CL2 NXOUT 0 10PF
VC5 VCC 0 DC 5V
V99 VPS 0 PWL ON 5V .12US 5V .125US 0V .130US 5V 1MS 5V
.ENDS
```

.ENDS

* TMR oscillators

```
XOSCA OSCINA OSCOUTA OSCA
XOSCB OSCINB OSCOUTB OSCB
XOSCC OSCINC OSCOUTC OSCC
```

* Transient analysis

```
.TRAN 1NS .6US 0.1US 10NS
.PROBE V([OSCOUTA]) V([OSCOUTB]) V([OSCOUTC])
```

* N-Well transistor model

```
.MODEL NMOS NMOS
+ LEVEL=2 VTO=0.825 UO=608
+ TOX=4.0E-8 NSUB=7.75E15 XJ=4.50E-7
+ LD=1.121E-7 DELTA=3.714 VMAX=49.89E+3
+ NFS=.105E12 CJ=323.1E-6 CJSW=929.9E-12
+ MJ=461.5E-3 MJSW=268.3E-3 PB=.44
+ CGSO=96.77E-12 CGDO=96.77E-12 CGBO=40.0E-12
+ UCRIT=50E3 UEXP=78.26E-3 NEFF=3.36 TPG=1
```

.MODEL PMOS PMOS

```
+ LEVEL=2 VTO=-0.703 UO=205
+ TOX=4.0E-8 NSUB=1.486E16 XJ=450E-9
+ LD=230.5E-9 DELTA=1.843 VMAX=40.76E3
+ NFS=0.01E12 CJ=804.9E-6 CJSW=749.1E-12
+ MJ=525.0E-3 MJSW=495.4E-3 PB=.958
+ CGSO=199.0E-12 CGDO=199.0E-12 CGBO=101.5E-12
+ UCRIT=70E3 UEXP=184.2E-3 NEFF=0.69 TPG=-1
.END
```


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n the range of 105, and the oscillator is free-running with no driving signal, complete analysis and simulation techniques in both time and frequency domains are known [8].

Transient simulation of a high-Q quartz-crystal circuit is a real challenge. Most designers rely on the frequency responses (small-signal analysis) to judge the ability of their circuits to oscillate. Transient response uses large-signal analysis. This involves nonlinearity operation and time-step estimation, which often results in divergence.

A few tricks can make the oscillation waveform come alive. In this circuit, the three oscillators are kick-started at different times with independent power supplies to simulate their independent phase relationship. The command V99 99 0 PWL 0NS 5V .1US 5V .105US 0V 11US 5V 100MS 5V introduces a voltage spike to the power supply to kick-start the oscillator. Note that in a real circuit, oscillation is started by random thermal noise in the circuit elements.

This noise is amplified by the inverter and is fed back positively through the crystal circuit. The amplitude of the signal increases exponentially as it goes through the loop again and again until saturation is reached. The frequency of the signal is controlled by the crystal. The crystal acts as a narrowband filter that passes only the frequency in the vicinity of the resonant frequency.

The CP is in parallel with the motional arm of the crystal and diverts the energy going through the crystal. This reduces loop gain and makes start-up slower. To show the transient start-up from time zero to saturation in a more reasonable time frame, we can comment out the CP temporarily.

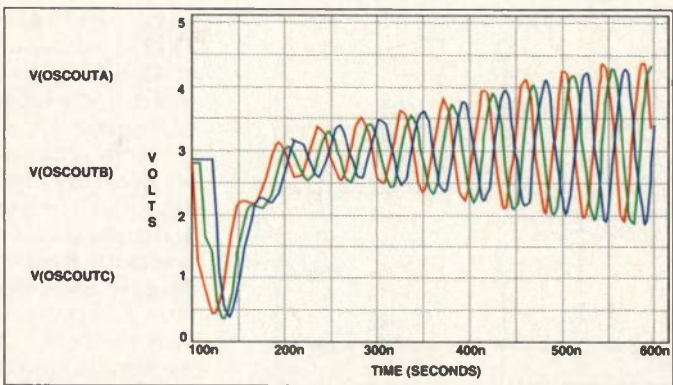


Figure 2. Independent oscillator startup transient.

Together with the higher-gain, triple-stage amplifier, the simulation time is decreased more than 20-fold from hours to a few minutes on a Sun Sparcstation. Figure 2 shows the transient response for the independent oscillator's start-up.

Capacitive coupling

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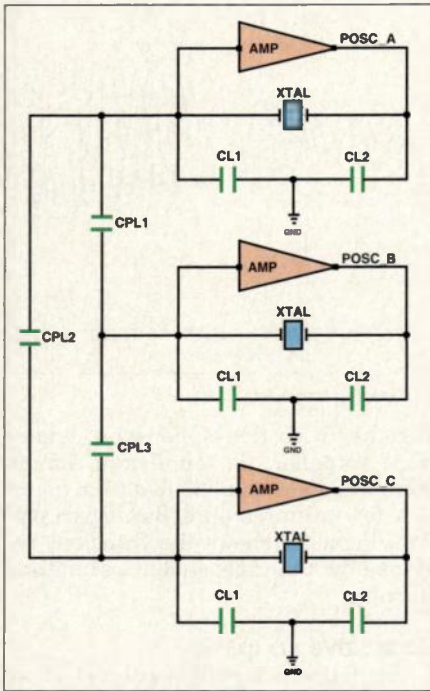


Figure 3. Capacitively coupled oscillators.

phase variation incurred in the loop can be tolerated and is translated into a small frequency shift. The flexibility of the crystal to yield to a driving signal operating close to its resonant frequency is exploited for synchronizing

PROGRAM 2

* TMR oscillators
XOSCA OSCINA OSCOUTA OSCA
XOSCB OSCINB OSCOUTB OSCB
XOSCC OSCINC OSCOUTC OSCC
CPL1 OSCINA OSCINB 15PF
CPL2 OSCINA OSCINC 15PF
CPL3 OSCINB OSCINC 15PF

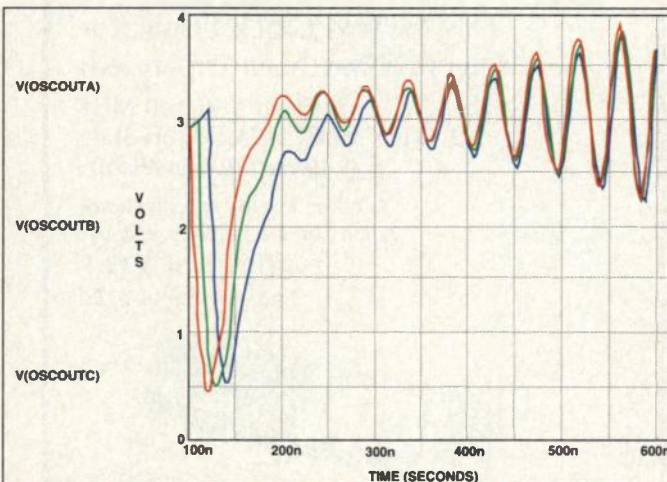


Figure 4. Capacitive coupling startup transient.

them mutually. Taking advantage of this frequency pulling, we can synchronize the oscillators mutually by coupling them capacitively. Figure 3 illustrates the schematic diagram. Because each oscillator forms a complete feedback loop, the coupling capacitors can be positioned at the input of the crystal network or at the input of the inverting amplifier.

The modification to the main body of the original SPICE listing shown in Program 2 results in the transient waveform of Figure 4. Note that the oscillators achieve synchronization after only a few clock cycles. This capacitive coupling design is not fault-tolerant; a failure in one of the coupling capacitors would cause the entire system to fail.

Feedback voting

To avoid the failure mode of the capacitively-coupled circuit, another way of driving the crystals with identical signals is to insert a majority voter in the feedback loop of each oscillator. The output of all amplifiers would drive the voters, and the identically-voted signal would drive the crystals.

A conventional digital voter would implement the two-out-of-three majority function $F = (AB + AC + BC)$ with three 2-input AND gates and one 3-

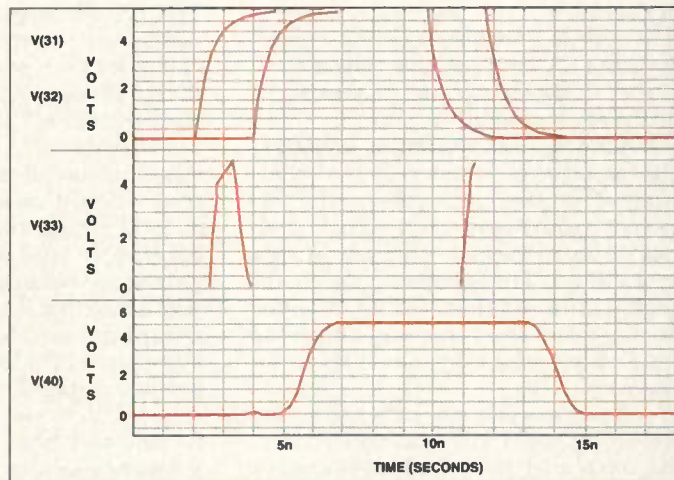


Figure 7. Voting with tight synchronization.

input OR gate (or a Boolean equivalence such as all NAND gates). To reduce the transistor count and logic delay, a transistor-level schematic for a voter that has only 12 transistors is shown in Figure 5. The majority function implies that unless all of the operational elements agree in value, the voted output will follow the faulty ele-

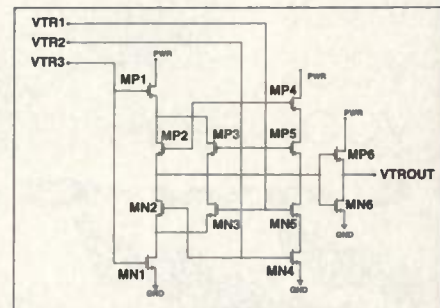


Figure 5. Majority voter schematic.

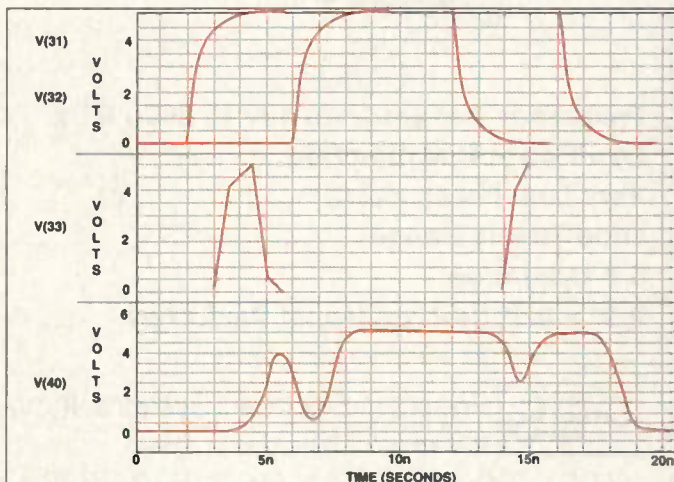


Figure 6. Voting with loose synchronization.

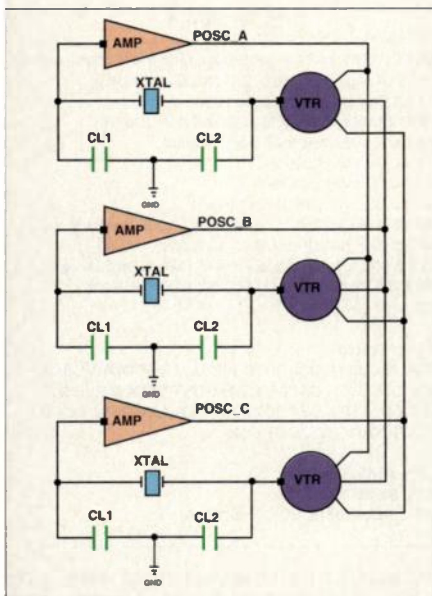


Figure 8. Oscillators with feedback voting.

ment; therefore, if two of the inputs are tightly synchronized (in exact agreement) to each other, the third input can fail without affecting the voted output. Otherwise, four oscillator channels and three-out-of-four voters will be required to survive a fault.

Although it has been proved that $3m + 1$ modules are necessary to tolerate m faults, loosely synchronous inputs to a majority voter are a problem. A tightly synchronized triple modular redundancy (TMR) design can tolerate any fault, including the Byzantine asymmetric fault.

A Byzantine fault sends different signals to different modules (or the signals are perceived differently by different modules). A Byzantine fault-tolerant algorithm makes no assumption about the behavior of the fault, no matter how arbitrary. This type of fault is devastating to a majority voter with loosely synchronous inputs, because the resulting outputs of the voters would be in disagreement. Within the displacement window (4 nanoseconds in this example) between two clock edges, where one clock is high and the other is low, the voter's output will follow the third input. Figure 6 illustrates the case in which a glitch in the rising-edge displacement window and an inverted clock in the falling-edge displacement window both cause havoc to the voted output. On the other hand, when two clocks are tightly synchronized (2 nanoseconds in this example), the faulty third clock effectively is fil-

tered out by the voter. Figure 7 shows the resulting waveform of the majority voter under tight synchronization. All real signals and circuits have limited bandwidth and inherent delay caused by parasitic and input and output loading capacitors. When the other two inputs of the voter are tightly synchronized or

when the skew is a fraction of the rise time (for example, a 1 nanosecond skew for a 5 nanosecond rise-time signal), the third input can fail within this window and still cannot pass the voter.

Figure 8 illustrates the schematic diagram for the TMR arrangement of oscillators, each with a majority voter in

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PROGRAM 3

```
* transistor-level voter
.SUBCKT VOTER VCC 31 32 33 40
MN1 34 33 0 0 NMOS L=2U W=8.5U
MN2 39 32 34 34 NMOS L=2U W=8.5U
MN3 39 31 34 34 NMOS L=2U W=8.5U
MN4 36 32 0 0 NMOS L=2U W=8.5U
MN5 39 31 36 36 NMOS L=2U W=8.5U
MN6 40 39 0 0 NMOS L=2U W=8.5U
MP1 35 33 VCC VCC PMOS L=2U W=12U
MP2 39 32 35 35 PMOS L=2U W=12U
MP3 39 31 35 35 PMOS L=2U W=12U
MP4 37 32 VCC VCC PMOS L=2U W=12U
MP5 39 31 37 37 PMOS L=2U W=12U
MP6 40 39 VCC VCC PMOS L=2U W=12U
.ENDS

* Oscillator circuit
.SUBCKT OSCA VTR1 VTR2 VTR3 AMPOUT
XVOTER VCC VTR1 VTR2 VTR3 VTROUT VOTER
RS VTROUT NXIN 200
* other components remain the same
.ENDS

.SUBCKT OSCB VTR1 VTR2 VTR3 AMPOUT
XVOTER VCC VTR1 VTR2 VTR3 VTROUT VOTER
RS VTROUT NXIN 200
* other components remain the same
.ENDS

.SUBCKT OSCC VTR1 VTR2 VTR3 AMPOUT
XVOTER VCC VTR1 VTR2 VTR3 VTROUT VOTER
RS VTROUT NXIN 200
* other components remain the same
.ENDS

* TMR oscillators
XOSCA OSCOUTA OSCOUTB OSCOUTC OSCOUTA OSCA
XOSCB OSCOUTA OSCOUTB OSCOUTC OSCOUTB OSCB
XOSCC OSCOUTA OSCOUTB OSCOUTC OSCOUTC OSCC
```

its feedback loop. Here, the voters provide identical driving signals for the crystals. With the crystal circuits designed to have a common pulling range, phase differences among the modules are compensated dynamically and continuously by the crystals within each clock cycle.

In a crystal oscillator, circuit-frequency stability predominantly is controlled by the crystal itself. The crystal frequency, on the other hand, is determined by its mechanical characteristics, such as thickness, elasticity,

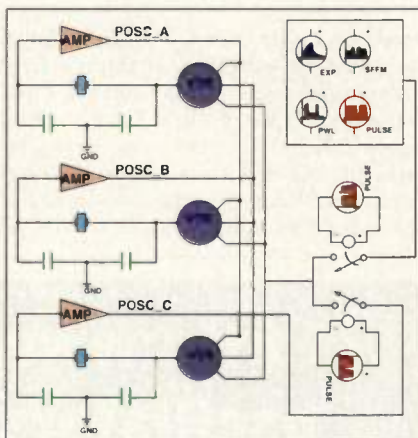


Figure 10. Verification by fault injection.

resonant area of the quartz, lead attachment and package sealing. Therefore, it is expected that drive energy, aging, temperature, humidity and other environmental conditions will have influences on the frequency stability of the circuit. One important characteristic of the crystal is its ability to oscillate at the frequency of a driving signal in the proximity of its own natural frequency. This operation allows all crystals in the system to be pulled to the same voted frequency, thus neutralizing the effect of small fluctuations in circuit components.

The addition of the voter subcircuit and modification as shown in Program 3 to the original SPICE listing results in the transient waveform of Figure 9. Note that the mutually voted oscillators immediately achieve synchronization after the first clock cycle.

Fault injection

Figure 10 illustrates the schematic diagram for the mutually voted oscillators with fault injection. Two voltage-

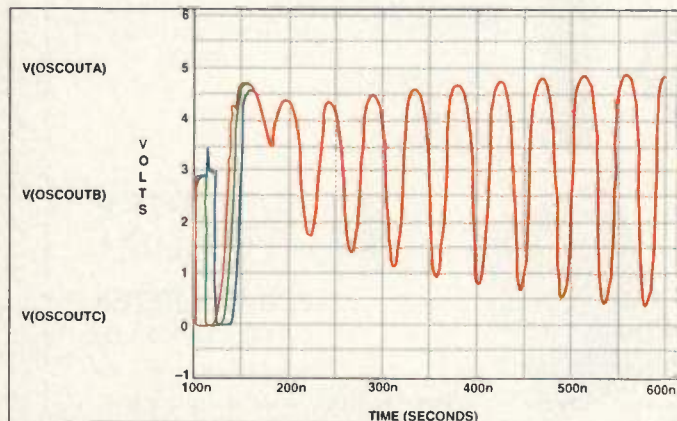


Figure 9. Feedback voting startup transient.

PROGRAM 4

```
* Fault injection logic block
.SUBCKT FILB 5 4 S1 5 4 8 0 SMOD ;fault free control switch
VS1 8 0 PULSE (5V 0V .6US 1NS 1NS 0.25US 2US)
S2 6 4 9 0 SMOD ; fault injection control switch
VS2 9 0 PULSE (0V 5V .6US 1NS 1NS 0.25US 2US)
* Selection of sources for Fault Injection
*VSA0 6 0 DC 0V ;stuck-at-0
*VSA1 6 0 DC 5V ;stuck-at-1
*VSA1 6 0 DC 2.5V ;stuck-at-midlevel
VNOISE 6 0 SFFM (2.5V 2.5V 24E6 200 1.2E6) ;noise burst
*VEXP 6 0 EXP (0V 6V 0.6US 0.1US 0.75US 0.1US)
*VSIN 6 0 SIN (2.5V 2.5V 24E6 0.5US 10E6 0) ;brown out
*VRAMP 6 0 PWL (0 0V 0.6US 0V 0.7US 5V .8US 0V)
.ENDS

* TMR oscillators
XOSCA OSCOUTA OSCOUTB AFGOUT OSCOUTA OSCA
XOSCB OSCOUTA OSCOUTB AFGOUT OSCOUTB OSCB
XOSCC OSCOUTA OSCOUTB AFGOUT OSCOUTC OSCC
XFILB OSCOUTC AFGOUT FILB

* voltage control switch
.MODEL SMOD VSWITCH
+ RON=1 ROFF=10E6 VON=5 VOFF=0
```

controlled switches are used with opposite controlling pulse waveforms. One switches the output of the C oscillator off; the other switches the arbitrary waveform generator on for fault injection, and vice versa. This way the injection time can be controlled precisely. The switches are a special kind of voltage-controlled resistor in which the resistance continuously varies between R_{ON} and R_{OFF} .

Predefined sources for fault injection include stuck-at-0, stuck-at-1, stuck-at-midlevel, noise-burst, slow-varying-signal that crosses the logic threshold decaying-oscillator to midlevel-logic and ramping-signal. Other kinds of fault injection waveform also can be defined. The injected waveform is fed back to all three channels to simulate a faulty oscillator C output.

Program 4 shows the fault-injection logic subcircuit and modification to the SPICE listing results in the transient waveforms of Figure 11 for a burst.

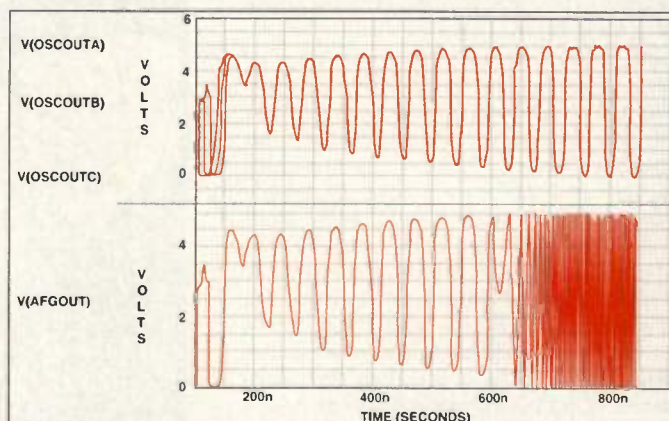
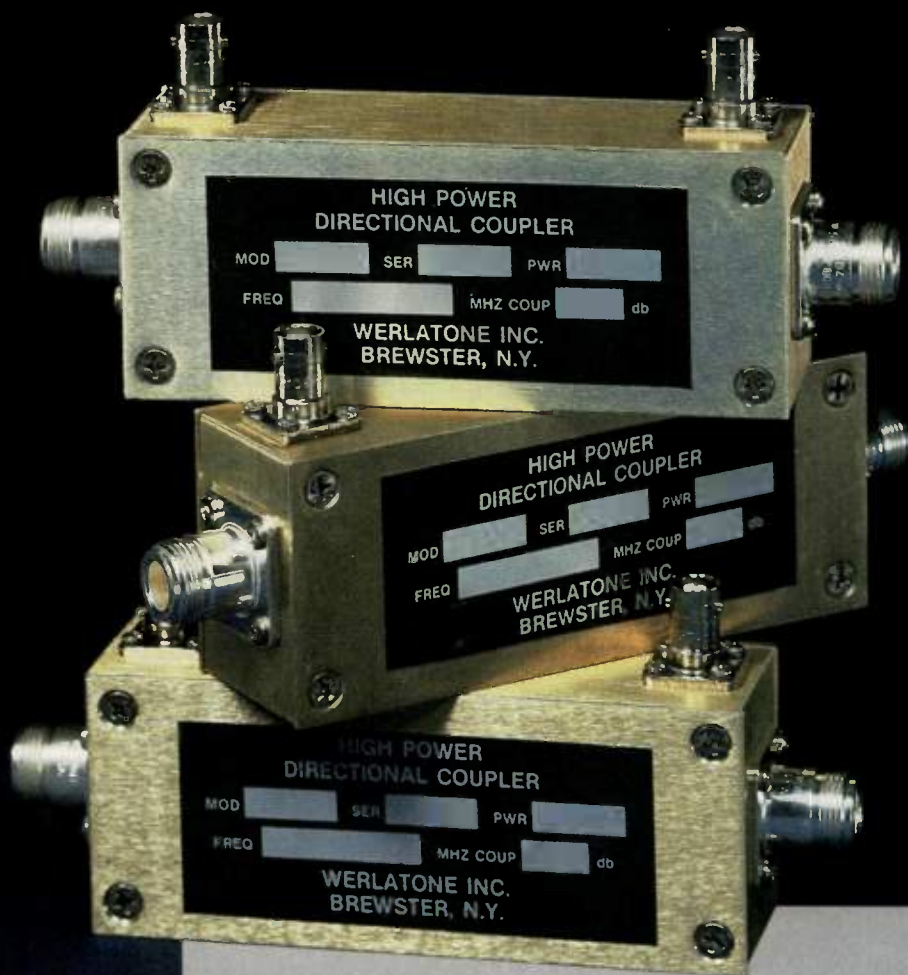


Figure 11. Fault injection with noise burst.

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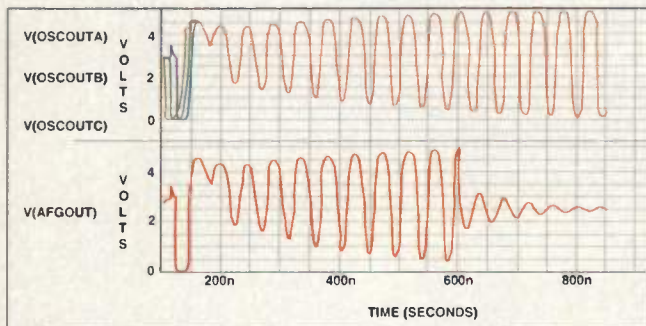


Figure 12. Fault injection with brownout oscillator.

noise injection and of Figure 12 for an oscillator brownout.

Overtone oscillators

For high-frequency applications (60 MHz or greater), we can make the crystal operate at a third overtone of the fundamental frequency by adding an inductor parallel to the input-load capacitor of the crystal. The newly-formed, inductive-capacitive (LC) tank forms a resonant circuit at just below the third overtone to suppress the fundamental frequency. Below this resonant frequency, the LC tank is inductive, and oscillation is not possible because there no longer will be a phase reversal in the feedback loop.

To model the third overtone crystal for simulation, we can add another branch of resonant arm R_x , L_x and C_x to the basic crystal model. Each series arm accounts for one resonance in the vicinity of which the other arm can be neglected. The new motional parameters should form a resonant circuit at the third overtone with:

$$C_N = \frac{C_1}{N^2}, L_N = L_1 \text{ and } R_N = R_1 N^2$$

where $N = 3$.

The frequency pulling range of a crystal at an overtone would be narrower than that of a fundamental frequency, because the pulling range is proportional to the motional capacitance of the crystal.

Fault-tolerant clock prototype

To be truly fault tolerant, the feedback distribution network has to be protected from short faults. For example, a short between two inputs of a voter would affect the two sourcing oscillators. Figure 13 shows the schematic for a fault-tolerant clock that has been built and tested [9].

The interconnection uses direct point-to-point wiring between the buf-

fers. There are n^2 interconnects (16 lines for the four channels). The isolation buffer network helps to tolerate multiple faults on the bus and helps to prevent a single module failure that can bring down the bus. Without the buffers, a bridge fault that shorts any two clock lines of two modules would fail both modules.

Note that some failures do not combine together to cause system failure. For example, if one oscillator fails s-a-1, and another one fails s-a-0, then the voted clock outputs still are correct. Also, there are certain multiple failures across all channels that will not cause system failure. To illustrate this latter case, a failure in any one input pin of each voter accounts for three simultaneous faults in three channels. Because of the replicated data paths and the isolation buffers, all three clock outputs still are correct.

Validation testing with various fault types injected to a single oscillator includes stuck-at-low, stuck-at-high, stuck-at-midlevel, open- and short-circuited, bridging, power-supply-failure, noise-coupling and random-signal, such as frequency sweep from subharmonic to overtones, frequency-burst, random-pulse-train and glitches.

Conclusion

For digital microprocessor applications where small frequency shift (less than 100 ppm or 0.01%) is inconsequential, off-the-shelf crystals can be used, and frequency pulling will compensate for the differences in crystal aging, temperature variation and circuit deviation (process and components). Given the skew caused by integrated circuit process variation and probe delay variation, the clock outputs are observed to be synchronized to within 0.25 nanoseconds in the prototype.

When a channel is faulty, the design purpose is to maintain synchronization in the remaining good channels and to continue to tolerate any additional fault that occurred in that channel; therefore, each oscillator channel is a fault containment region, with the voters and isolation buffers to quarantine the fault. One programmable logic device per channel would ensure physical and electrical isolation among the

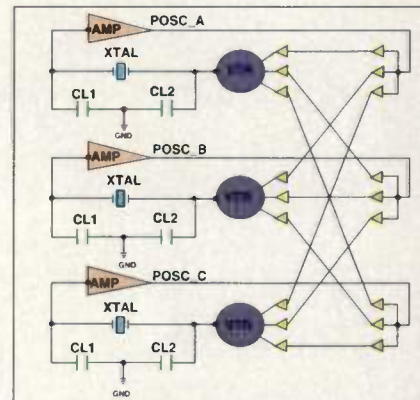


Figure 13. TMR fault tolerant clock.

channels. On the other hand, a single custom, application-specific integrated circuit (ASIC) using silicon on sapphire (SOS) technology or using CMOS with on-chip electrical isolation would make an ultra-reliable fault-tolerant clock suitable for many applications. **RF**

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About the author

T. Kien Truong is an engineer with Boeing Commercial Airplane in Seattle, where he works on integrated avionics research. An Iowa State graduate and holder of three U.S. patents, his experience ranges from antenna and RF to fiber-optic sensors and fault-tolerant computing. He can be reached at tkt@clr.iastl.ca.boeing.com.

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WRN

PCS—working to make the link

By Ernest Worthman
Contributing Editor

Personal communications service (PCS) has been touted as anything from a simple "follow-me" telephone to a full-featured digital voice and data transceiver. The vendors at the PCS '96 trade show portrayed PCS as the answer to all our future communications needs. But if you have been in the wireless industry for a while, you have a more realistic perspective of the technology, its implementation and its timetable.

Although many would like to carve out a uniquely identifiable niche for PCS, the reality is that both the con-

cept and some of the technology have been with us for quite some time in limited implementation and with different names. Technologies such as talk-back paging and cellular telephones loosely can be considered as PCS. In true PCS, the fact is that these technologies will be one or more of many components that make up the personal communications system infrastructure.

Modern PCS can be viewed as the technology with which the linking of places ends, and the linking of people begins. Additionally, PCS should be considered as "intelligent" communica-

tions, meaning it has more capability than mere signaling confirmations, two-way voice and simple data. Here are the issues that PCS faces.

The year 2525

It is 8:00 a.m., and you just have arrived at your office. Your personal communicator signals, and you answer. Up comes a miniature liquid crystal display (LCD) with a message that the temperature in the master bathroom has increased by 10°. You enter a code and lock onto the cameras installed throughout your home. You scan the master bath-

The future of DECT

Digital European cordless telephone (DECT) was originally planned as a replacement for analog cordless systems. After undergoing four years of refinements since it was issued by the European Telecommunication Standards Institute (ETSI), DECT can now boast additional features that include interoperability among various equipment and systems as well as new applications such as data transfer and connection via a wireless local loop (WLL). Numerous European countries already have reserved the frequency band of 1.88–1.9 GHz for DECT operation. DECT offers integrated services digital network (ISDN)-like speech quality, which outperforms analog systems and typically even other cellular phone systems, such as global system for mobile communications (GSM, formerly Groupe Speciale Mobile). The availability of a standardized and allocated frequency band, available data rate, interference-free operation and easy system upgrades are advantages that will continue to further the use of DECT worldwide.

DECT currently can be integrated into existing mobile communications systems such as GSM and digital cellular system (DCS) 1800, and can be hooked up directly to fixed-network

technologies such as ISDN, Ethernet and asynchronous transfer mode (ATM).

Future outlook

With the "anytime anywhere" outlook overtaking business and society, DECT system use is sure to expand. Rapidly growing and expanding applications include cordless residential and business handsets, base stations, wireless private branch exchanges (PBXs), WLL and wireless local area networks (WLANs). Expected soon are further feature advances such as generic access profile (GAP). GAP will enable interoperability among different vendors' equipment and devices. Another addition is the cellular telephone modem (CTM) proposal that will add mobile public access at traveling speeds as high as 70 kilometers per hour via public DECT base stations. These and other feature enhancements are sure to add the flexibility end-users demand.

Although initial sales have been in Europe, the Middle East, Southeast Asia and Australia also are opting for DECT. China's approval is expected soon. By some estimates, DECT phone sales are projected to outstrip sales of cellphones worldwide within three

years. According to the Giga Information Group, the DECT market will be a \$1.9 billion market by 1998, with silicon content accounting for 20% of that value. U.S. analysts at I.C.E. predict a compound annual growth rate of 74% for semiconductors in DECT systems with a compound annual growth rate of 85% for DECT systems.

The implementation of DECT in the United States is limited by incumbent microwave users that already occupy the 1.8–2 GHz band. With component manufacturers fully committed to this digital cordless standard, DECT looks to have a promising future worldwide—especially in locations where telecommunications infrastructures have not been installed already or in densely populated urban areas where implementing further wiring is cost-prohibitive.

With DECT hardware and software design and production trends leaning toward further integration and high volumes of key components, more wireless applications (remote control systems, surveillance systems and mobile-to-mobile systems, for example) will be able to take advantage of the DECT standard.

—Greg Ravenscroft
National Semiconductor

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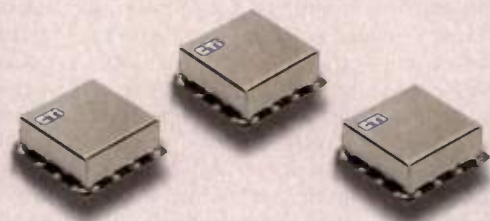
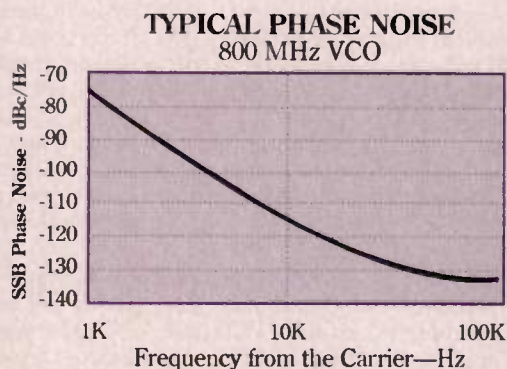
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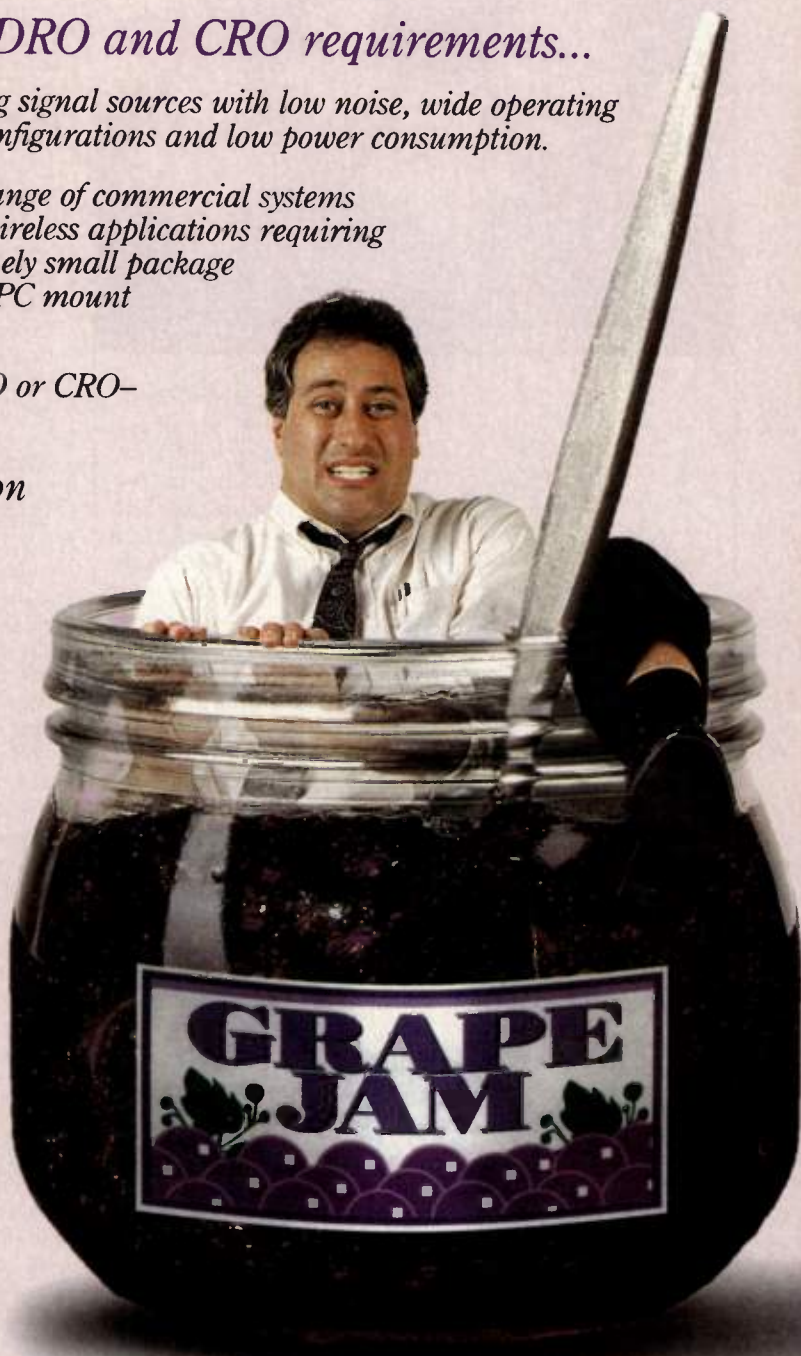
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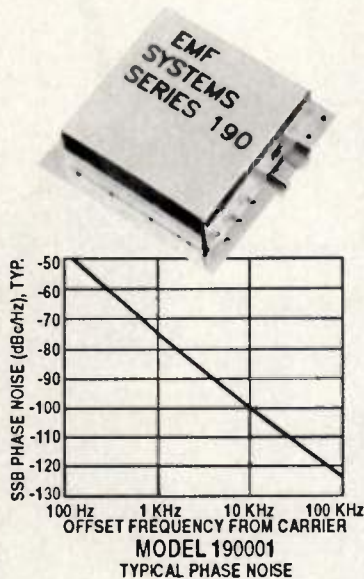
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The plot at left indicates the typical phase noise of EMF Model 190001.

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and find that you left the wall heater on. A few more key strokes, and you are tied into your home computer. A few quick commands, and the heater is turned off. Just to be sure, you scan the other rooms and find that the curtains in the den were left closed. You instruct the computer to open the drapes. Finally, you instruct the environmental monitoring system to disregard the alarm.

Next, you tie into the company's central data terminal and check your messages, meetings, E-mail and data that may have been sent since you left the office. Finally, you are ready to leave for a trip. You instruct your bank to pay the monthly bills electronically while you are out of town. Then, as you arrive at the airport, your communicator automatically links to the Iridium low earth orbit (LEO) satellite network.

Sound far-fetched? Not as far as you may think. This scenario is a typical example of intelligent communication linked to people rather than to places. The communicator you carry replaces today's wireline telephone, cellular telephone, laptop computer, pager and personal digital assistant (PDA).

Two elements of technology will be required to make this above scenario a reality. The first is the infrastructure, the second is the communicator.

The infrastructure

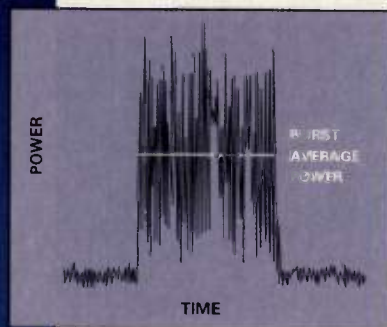
Assessing the principle needs for the infrastructure, the first issue is, of course, *spectrum*. International PCS is in the 1.7-2.0 GHz band. Although it might seem that there is sufficient spectrum at those frequencies to have the full offering of voice, image, data, full-motion video, 16-bit 44 kHz audio and other bandwidth-hungry technologies, lots of spectrum will be required. Fixed-link systems exist in this frequency band, too. Moreover, this spectrum will need to be globally common (rationalized), as some say overseas. Table 1, p. 52, shows typical data rates for potential PCS services. A bit of math will show that the bandwidth for audio and video can be quite demanding. Additionally, if services such as full-motion video are to be offered, the data rates and related bandwidth demands can be much higher.

Another issue the PCS infrastructure faces is real estate. PCS is a low-power, line-of-site technology. There will have to be many more antenna sites for PCS than for cellular and mobile radio, and they will have to be closer together. Sites will have to be mounted on buildings, power poles, light poles and other non-industry-owned properties or locations. Physical impediments to support power, cabling and rights-of-way will have to be considered along with the aesthetic perspective.

Reliability also becomes an issue. Because the system will be required to support "walk and talk" with high frequency and low power, propagation losses will be more of a problem with PCS than with cellular, specialized mobile radio (SMR) or satellite communications. Current wireless—especially cellular—systems have signal problems in areas such as parking garages, tunnels, terrain dead spots and even in some foliage. My experience with cellular service has proved it to be unreliable. I constantly experience dropped calls, fading, noise handoffs, out-of-area problems and co-channel interference. PCS systems are to be as ubiquitous as is predicted, they will require intense testing, flexible configuration and fail-safe redundancy before the users will believe in them and use them as their primary communications systems.

Yet another infrastructure issue is interconnect. Many other communications systems are in use today: satellite, public switched telephone networks (PSTN), computer database, digital European cordless telephone (DECT), cellular and

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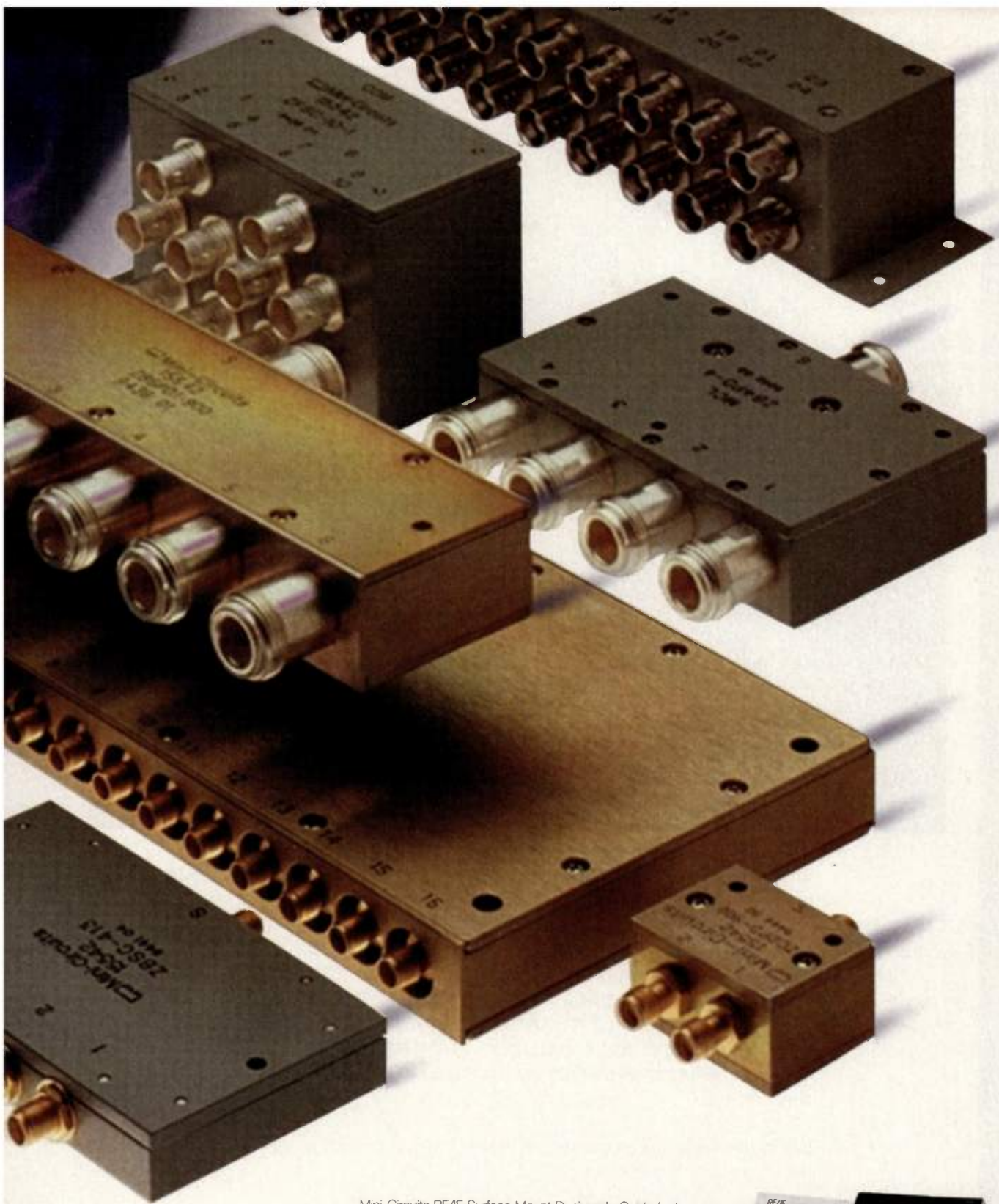
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SMR. For those systems to "talk" to each other often requires manual intervention or long waits and special equipment. A new level of interconnect hardware and software will have to be defined and implemented cost effectively, so the user can have access everywhere. Countries, states, local municipalities, territories, regions and governing bodies all have their own regulations and restrictions. So, in addition to the interconnect issues are the political ramifications of dealing with a worldwide political scene.

Next, transmission technologies will have to be either interfaced or standardized. Current wireless technologies implement time-division, multiple-access (TDMA); code-division, multiple-access (CDMA); and frequency-division, multiple-access (FDMA) as the most common. These complex transmission schemes are not compatible fundamentally and will require sophisticated conversion techniques to allow transparent hand-off from system to system.

Finally, intelligence will have to be built in and distributed across the system. This means that sophisticated software will be required to integrate the various components. Advanced intelligent networks (AINs) will become the backbone of the infrastructure. AINs will route calls based upon user profiles. They will accomplish tasks such as recognizing the difference between paging messages and data, and between voice and video. PCS subscribers will carry a credit-card sized smart device called a subscriber identification module (SIM) that contains the subscriber's data, billing information and caller identification (much like today's calling card). This card will allow the subscriber to access the network from any location and still be identified and properly charged when not using his or her own personal communicator.

The communicator

One of the major issues surrounding PCS is the communicator that will be used to access the network. As a society, we have become used to the traditional telephone look and feel as our standard communication device. If the AIN is to be realized, though, the communicator will have to be able to accommodate the various technologies (video, audio, images). For this, the standard telephone-type handset will not work. A more likely scenario will be a handset such as the one shown in Figure 1, p. 52.

Another possibility will be a variant of the personal digital assistant (PDA). PDAs offer the possibility of becoming a PCS terminal because they contain the necessary components to handle the technologies. Unfortunately, today's PDAs do not fit the mold of portable phones, and they are somewhat awkward

to carry. The ergonomics of handset design are fairly well defined, and at least for the near future, handsets will have to remain similar in size and weight to the current portable and cellular design. In terms of handset design, two of the issues that face developers are display-type power use and

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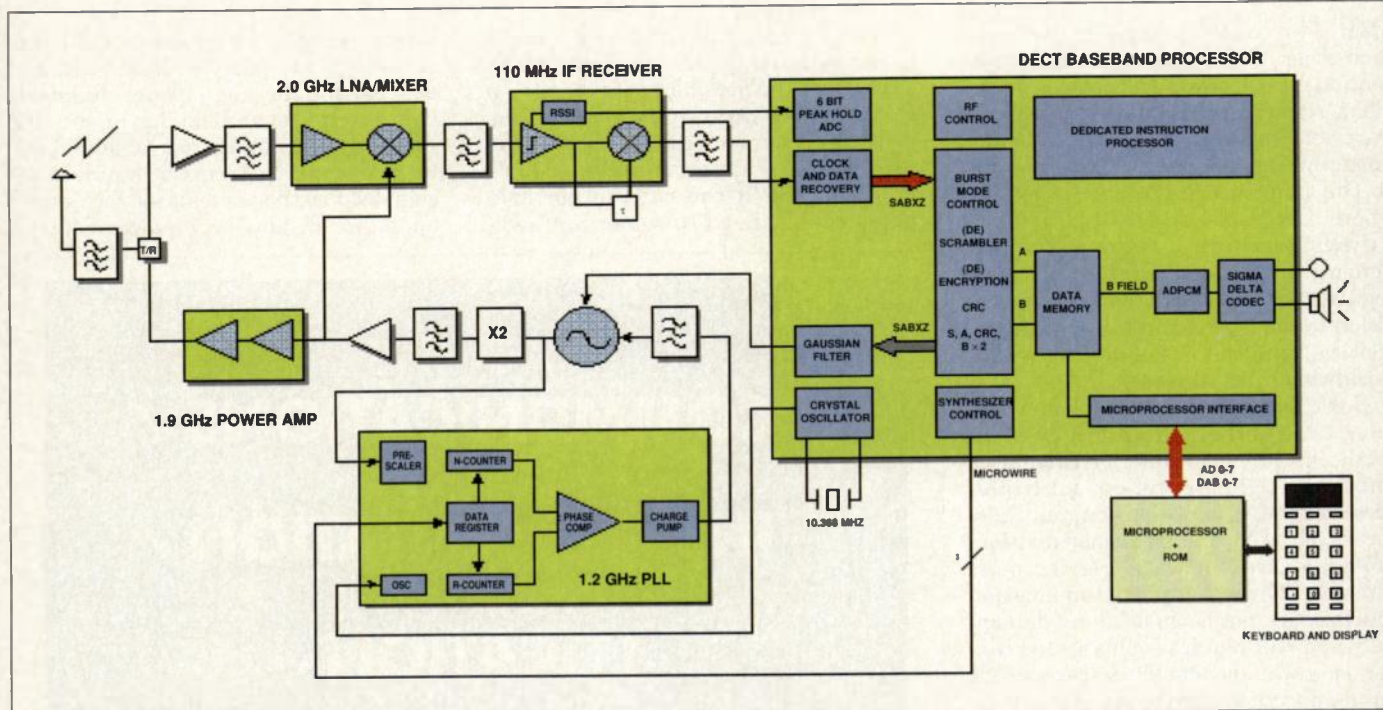


Figure 1. A complete DECT handset solution.

battery technology. Although a great deal of progress has been made in power-saving designs, one area that has made little progress is *video panels*.

Despite all of the hype promising new plasma and other focusing technology, the current rash of LCDs still leaves something to be desired. Active matrix types such as dual-scan or thin-film transistor (TFT) give better viewability than passive matrices, but require more power. Passive, low-power LCDs are inexpensive and easy to integrate, but offer poor viewability. Perhaps a good compromise will be low-current, light-emitting diodes (LEDs) once the blue LED becomes more popular.

The power issue also haunts portability. Battery science has developed new cell technology based upon nickel-metal hydride and lithium-ion (L-ion) chemistries. Both offer higher-power

density than the traditional nickel-cadmium, offering better power-to-weight ratios, but both are less stable and less forgiving than older battery chemistries.

For portable communicators to be accepted widely, convenience will be the issue. Portable communicators will have to have reasonably long use away from the recharger. They will have to recharge quickly and must not be sensitive to partial discharges or temperatures. Although the new battery chemistries look promising, and it is likely that L-ion cells will be the power source of portable products for the near future, there are some clouds on the horizon for L-ion chemistry.

First, L-ion batteries are expensive. It will take a while for the industry to have adequate supply. Second, L-ion cells require sophisticated charging

schemes to prevent overcharging and cell damage. Third, L-ion cells are more sensitive to temperatures in both discharging and charging than older chemistries. All of these issues will require "smart" communicators capable of monitoring internal components and power use. The present solution is to conserve as much power as possible by using various power-down schemes and sleep modes during idle or out-of-service times.

Finally, these communicators must be marketable, meaning they must be inexpensive and easy to use. For example the home video cassette recorder (VCR) has been around for more than 15 years yet more than 85% of the users cannot (or will not) take the time to understand how to use it for more than simple recording and playback.

When it comes to my cellular phone even I do not use it for anything other than holding a few stored memory numbers and everyday calling. Why? It takes too much time to learn all of the features. So it will be a real challenge to the PCS industry to come up with a communicator that will act like both a telephone and a computer, yet that will be easy to use.

Advanced intelligent networks

Clearly, AINs will be the backplan of the PCS network. AINs will be what

| APPLICATION | AVERAGE DATA RATES (KBPS) | PEAK DATA RATES (KBPS) | MAXIMUM DELAY (SEC) | MAXIMUM PACKET LOSS RATE (PACKETS) |
|--------------------|---------------------------|------------------------|---------------------|------------------------------------|
| E-Mail/paging | 0.01-0.1 | 1-10 | <10-100 | <10 ⁻³ |
| Computer data | 0.1-1 | 10-100 | <1-10 | <10 ⁻³ |
| Telephony | 10-100 | 10-100 | <0.1-1 | <10 ⁻⁴ |
| Digital audio | 100-1000 | 100-1000 | <0.01-0.1 | <10 ⁻⁵ |
| Video conferencing | 100-1000 | 1000-10000 | 0.001-0.01 | <10 ⁻⁵ |

Table 1. Data rates that PCS applications require and packet-switching must provide.

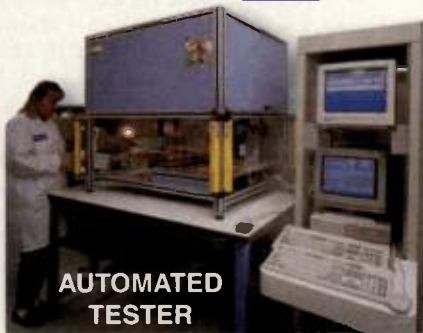
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allows the users to access the communications infrastructure without regard to location or information being processed. AINs will be software-driven for configurability.

The AIN consists of a three-layer model as suggested by Ashity, Sheika and Murthy (ASH93). The first level is the intelligence level, which will contain the necessary databases for the storage of information about the users. The next level is the transport level. This level handles the transport and is responsible for seeing that the information passes seamlessly from system to system and from user to user. The third level is called the access level. This level has the responsibility of integrating the different types of services. This level will be integrated closely with the transport level. It will contain the databases that track the user and will update the user databases as the user moves throughout the network.

AIN and PCS systems will pass data along this network using packet-switching technology and SS7 signaling. Packet-switching provides better reliability than circuit-switching and does not require a dedicated link with low bit error rate (BER). Also, packet-switched data can better compensate for lost or corrupt data than circuit-switched networks can. Table 1, p. 52, shows data rates that PCS applications require and that packet-switching must provide. The table also provides some data on delay and packet corruption rates. An interesting observation of the data is the delay that is acceptable with the different services. Notice that E-mail and data are the most tolerant of propagation and other delay elements,

whereas audio and video (real-time, high bit-rate applications) cannot tolerate much delay at all.

Other related technologies

As is commonly acknowledged, the only way to implement successfully many of the services PCS hopes to offer is to work in the digital world. Wideband RF signals cannot easily be digitized using high-speed analog-to-digital (A-to-D) converters. To try to digitize a 2 GHz signal using a lowly 4-bit sampling rate would require a frequency of 8 GHz—an intimidating challenge, even for ultra-high-speed A-to-D converters. A digital receiver, on the other hand, makes the job more manageable by inserting an analog downconverter into the channel before the A-to-D stage. Once processed, standard off-the-shelf (OTS) components can be used for the A-to-D converter and subsequent digital stages of the receiver. A digital receiver based upon this concept is cost-efficient and solves probably the most fundamental problem associated with spectrum loading and allocation.

Earlier, I presented a scenario of some of the potential of PCS. For this to be possible, one last piece of the puzzle must fall into place. That piece is the interface between communications and computers. Although there certainly are other possibilities, one promising technology is something called the VMEbus. Among other things, VMEbus offers A-to-D applications and interfaces for x86, reduced instruction set computers (RISC) and Motorola microprocessor-based systems. Currently, VMEbus is a prolific platform with a large number of offerings well-suited for interfacing the two technologies. It is possible that VMEbus will become the de facto standard for system integrators. This open architecture is one platform that system integrators can use either as off-the-shelf components, or to develop custom board-level components that can link computers and communications.

The future

PCS is a promising technology still in its infancy. Some PCS trial sites have had less than glowing success. Many of the technical issues need to be investigated further, and some do not have any immediate solutions. According to some FCC documentation, a number of PCS trials have been established merely to showcase the technology for customers. Not all licenses are being developed. Last year, the Canadian government had to "fire sale" licenses to generate interest.

Finally, PCS seems to have more of a following in other countries than the United States. Canada, for example, is embarking on some aggressive programs to bring PCS to large cities within the next two years. There is little doubt that PCS is the next generation of communications. But there is some doubt as to its timetable. It is unlikely that PCS will be as ubiquitous by the year 2000 as has been projected. Other technologies (e.g., cellular) are doing a reasonably good job of supplying today's wireless communications needs. The cellular infrastructure is far from bulging at the seams, and new innovations, especially in digital technology, continue to improve cellular services. There is a huge investment in current cellular technology, and it is not likely to die quietly while PCS takes over—unless, of course, the cellular companies become the PCS providers.

In the United States, the basic telephone infrastructure is aging but still adequate. The implementation of digital technology to the telephone infrastructure can give this old technology new life.

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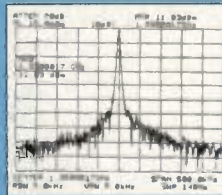
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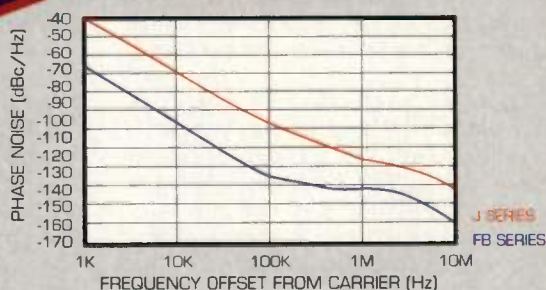
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MRI basics and coil design principles

By David M. Peterson, G. Randy Duensing, Ph.D. and J.R. Fitzsimmons, Ph.D.

Magnetic Resonance Imaging (MRI) has become a critically important medical imaging technique during the past 20 years. Improvements in technology continue to expand the scope and effectiveness of MRI. The system is composed of four major components: the static magnetic field, the pulsed magnetic field gradient systems, the image processing and display system and the radio-frequency (RF) transceiver. The following information includes a brief description of the relationship of these components. The balance discusses design details for high-field MRI RF receiver coils. Special-purpose coils are designed to optimize signal-to-noise ratio (SNR) from a given region of the body. The state-of-the-art coil system includes the use of four or more coils with four separate receivers. This method often is referred to as a phased array system, even though the signals are not added with fixed phase to one another.

The static magnetic field usually is denoted B_0 and is measured in Tesla (T). One Tesla is equivalent to 10,000 Gauss. This magnetic field causes partial alignment of the normally random magnetic spin moments of certain nuclei and, thus, of a bulk spin magnetization. Most notable of MRI sensitive nuclei is hydrogen, which normally is used in MRI because of a natural abundance in the human body. The spin is modeled as an infinitesimal magnetic dipole that can rotate but that cannot translate. In the presence of the strong static magnetic field, the bulk spin magnetization can be perturbed into an oscillation around the axis of B_0 . The oscillation frequency is linearly dependent upon the static field intensity. For hydrogen spins, this is 42.58 MHz/T [1]. A spinning magnetic dipole clearly will induce a voltage in a loop nearby, and this induction is the basis for signal collection in MRI. A graph of the resultant signal voltage would look like a sine wave damped by a decaying exponential with a time constant on the order of a second. The

Fourier transform would be a narrow line at a frequency related to the static field by the relation given above.

With only the static field and a means of perturbing spins, no spatial information is obtained because all hydrogen nuclei resonate at the same frequency in the same magnetic field, and negligible time delay occurs from different parts of the sample to the receiver loop. The typical method for obtaining spatial information is to change the net static field as a function of position. This method allows mapping of the observed frequency spread to the field distribution. Mapping normally is done in two or three directions to encode spatially the density of hydrogen within the sample. Many modifications exist for encoding properties other than hydrogen density, but they are largely irrelevant to RF coil design. The method commonly used for changing the net static field is to use windings that produce linear fields superimposed on the main field [1]. This creates a gradient in the static field. Consequently, these windings are called *gradient coils*.

Given the static field and gradient coils (and associated drivers and control circuits), the frequency spread of induced voltages can be related to position within the sample that results in an image of some region of the body. This image is produced by multidimensional Fourier transforms that are carried out after the received voltages have been sampled and stored in the computer. The entire frequency spectrum of interest usually is on the order of 10 kHz, an extremely narrow band, considering that the center frequency is about 100 MHz. This allows the use of single-frequency matching for coils because their natural bandwidth always exceeds the image bandwidth.

The quality of MR images depends on the SNR of the acquired signal [1]. The amplitude of the signal voltages is small and comparable to the amplitude of the thermal noise developed by the AC resistance of the body-loaded loops themselves. Additionally, increasing

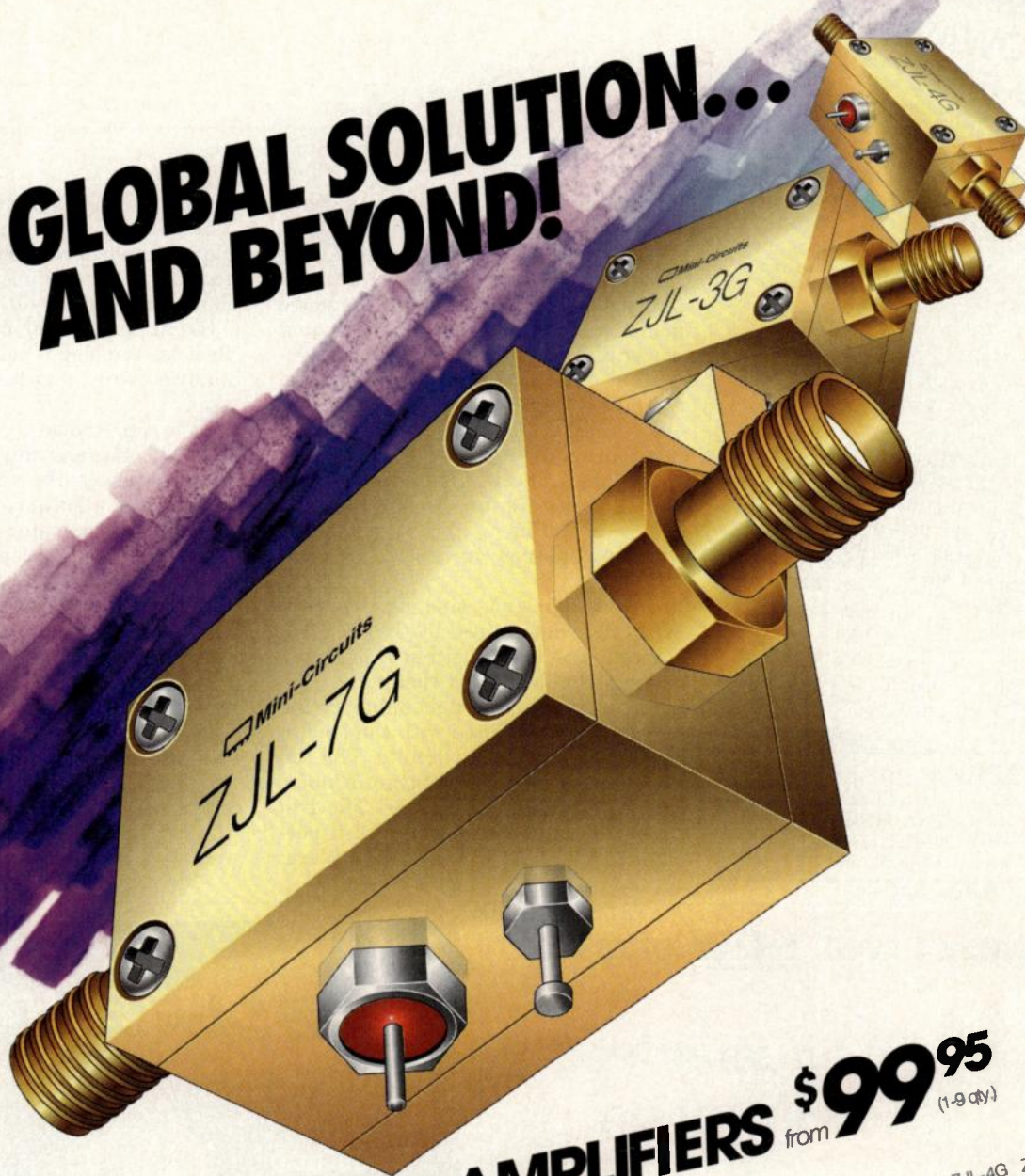
the spatial resolution tends to reduce the SNR, so that a constant battle is waged to obtain high-resolution, but clear, images of the interior of the human body. It has been shown that the obtainable SNR depends on the static field and, therefore, on the frequency of the spin oscillation. As a result, modern, whole-body MRI systems have static fields as high as 4 T (about 170 MHz oscillation for hydrogen). Because the signal transducer is the first element of the receiver chain, it determines the obtainable SNR and thus is critical to the quality of images.

Coil design principles

Coil optimization considerations include high-frequency effects, balancing, radiation, sample loading and multiple-coil interactions. The receiver coil must be sensitive to the spin magnetic moments in the region of interest. The coil's field direction, therefore, should be perpendicular to the static field. Furthermore, it should be as insensitive as possible to induced voltage or to noise from any sources other than the region of the body for which it is designed. These requirements are unusual and differ from the design of either transformers or antennas. The coil is a large inductor with considerable dimensions in comparison to a wavelength, yet no radiation is desired. Furthermore, the near field of the large inductor is loaded with a complex, imperfect conductor. The exact shape of the coil depends on the part of the body for which it is designed.

A simple circular loop is used for analysis. A standard communications antenna is a far-field radiator that transmits through the environment (mostly air), and its signal decays with distance. An MRI coil should be a poor far-field radiator and instead should couple inductively to the conductive sample; therefore, the power should be dissipated primarily in the near field close to the conductor. To optimize coils for high-field MRI systems, many techniques from antenna design are used,

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even though the goals are somewhat different.

To discuss the details of design, it is convenient to treat the coil as a transmitting source. The local RF magnetic field is related to signal reception, and power dissipation is related to noise reception. The perfect simple coil would produce a uniform RF magnetic field only in the region of interest, with no non-conservative electric fields and no radiation. The coil is usually a physically small antenna. (It fits inside a sphere of radius $\lambda/2\pi$.) We can use antenna techniques (in reverse) to optimize it.

The design and construction of an MRI coil is determined by the "load" on the coil. The load is either a phantom or the actual body part that will be imaged. All the factors that influence the coil must be evaluated with the load in place to have reliable results. The coil must be matched for optimum noise figure for the preamplifier. The coil should be electrically balanced if possible [2]. Imbalance will produce local

electric fields higher than necessary. This causes more power loss in the sample and may produce substantial radiation. To produce a local uniform field, it is necessary to make the source current in the coil uniform as well. Because the loop is relatively large, capacitors are distributed around the loop.

It is well-known that capacitance effectively shortens antennas and makes more uniform current. The standard for MRI coils is to have no more than a distance of approximately $\lambda/20$ between loop capacitors [3]. The capacitors also may have local electric fields that may interact with tissue and that may cause unnecessary losses; therefore, the capacitors should be located away from the tissue or shielded with the use of a thin copper foil.

Distribution of capacitance around the loop serves two purposes. First, as mentioned above, it maintains nearly uniform current around a large loop. If the current is uniform, then the only radiation is from magnetic multipoles

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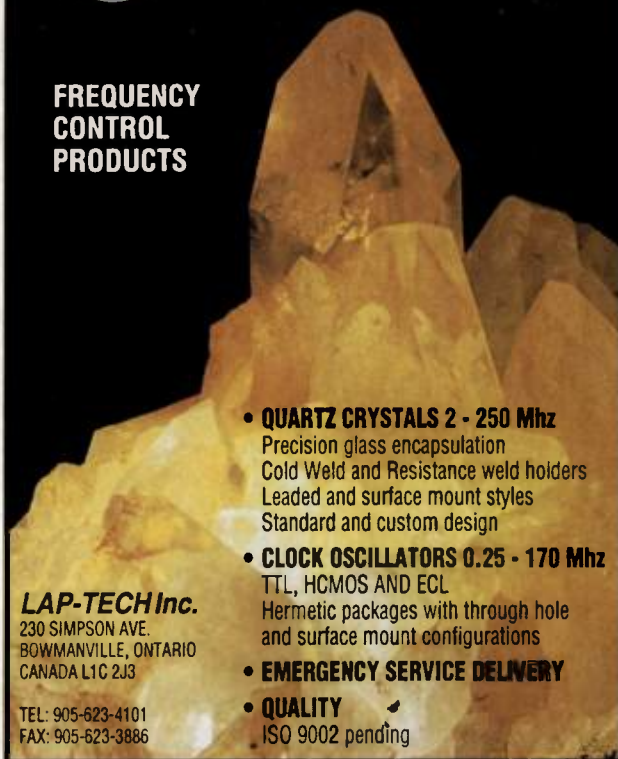
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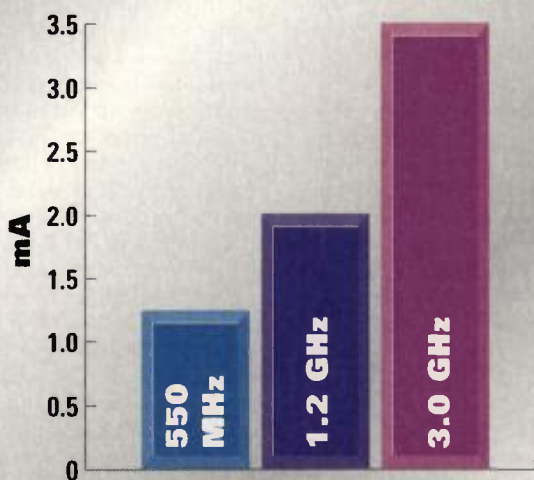
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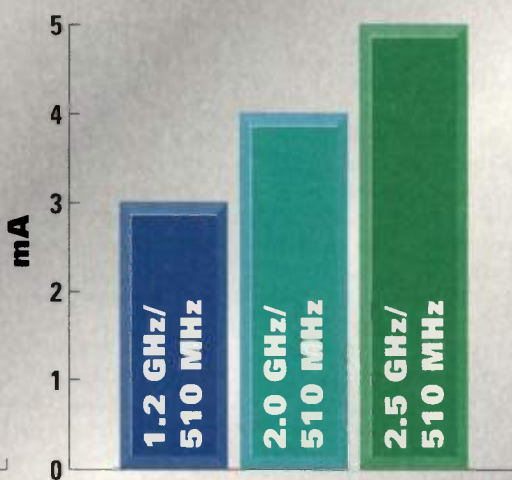
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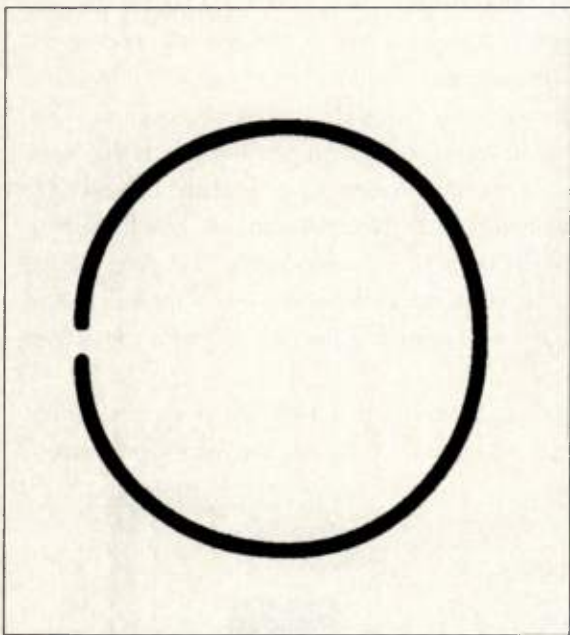


Figure 1. A simple circular loop with one gap.

generally dipole. Second, it reduces the maximum voltage present around the loop, which in turn reduces effects related to capacitive coupling with the coil. An important discussion centers on interactions between multiple coils at high frequency. Two coils are coupled if they share an impedance. At low frequency, the overwhelmingly dominant effect is mutual inductance between loops. This can be reduced or eliminated by proper geometry where the loops overlap each other such that the net mutual inductance is zero.

At higher frequencies, substantial coupling occurs through the sample. Because the sample is lossy, it is possible to have zero net mutual inductance but substantial mutual resistance [4]. This often is unavoidable and causes crosstalk and correlation of noise between channels.

Because the process of coupling to the sample is inductive, crosstalk can be reduced by overcoupling to the preamplifiers. MRI preamplifiers (GaAs FET) commonly are produced with input impedance as low as $2\ \Omega$. The coil is matched to $50\ \Omega$ with a matching circuit. In this case, the reactance x_m is inductive and equal in magnitude to x . When the preamplifier is attached, the impedance seen by the loop looking toward the preamplifier is higher than in the matched case, generally about 10 times greater. This reduces the current in the coil during reception without af-

fecting the SNR, and thus reduces inductively mediated crosstalk [5]. On the other hand, this increases capacitive coupling related to the voltage near the coil output because it increases this voltage relative to the matched case.

An increase in isolation is seen when using this matching scheme. Mutual resistance also may occur because of radiation effects that may be mediated by unbalanced currents to shields, between coils and through the sample. If the coils both have a given multipole character in the same physical direction, the radiation resistance increases over the sum of the two individual radiation resistances.

For multiple coils, where each is attached through coaxial cable to a low impedance preamplifier, there is considerable opportunity for interactions with the cable shields. Aside from balancing, discussed in detail below, there are other ways to eliminate these effects. A simple method used in conventional antenna design is to wrap the cable on a ferrite core. This makes a large inductive reactance that blocks this unbalanced current flowing on the outside of the coax.

For MRI applications, a non-ferromagnetic, narrowband equivalent is produced by again wrapping the coax into an inductor, but then a capacitor that is resonant with the shield inductance is placed from one end to the other. The capacitor produces a high impedance that blocks this unbalanced current, which reduces coupling to other elements of an array and eliminates loss associated with the shield current, including radiation losses. Be careful to shield the resonant loop formed by the shield and capacitor so other means of coupling are not produced.

Matching and balancing

It is conventional to transform RF coils for MRI to $50\ \Omega$ nominal impedance. The primary concern for these coils is SNR, and it is assumed that the attached low-noise preamplifier has an optimum noise figure when the source impedance is $50\ \Omega$. On modern MRI systems, the preamplifier input impedance is not $50\ \Omega$, i.e., the coil is not

power-matched to the receiver. Usually, the input impedance of the preamplifier is as low as possible, typically about $2\text{--}3\ \Omega$. The impedance mismatch is used to decouple inductively the coil from its surroundings, including other coils [5]. The impedance mismatch does not reduce the SNR even though it reduces the power delivered to the preamplifier. This was discussed above. The circuits discussed below all can be used for this purpose by carefully selecting components.

Associated with impedance-matching is the problem of coupling the coil to coaxial cable with a grounded shield. Early literature discussed the problem of balancing the coil because of its electric field interactions with the sample, which were viewed as lossy paths to ground [6]. Other related concerns, including coupling to other coils and to cable shields, as well as radiation effects, will be evaluated in the context of impedance-matching. Most of the techniques used at lower frequencies have positive benefits for high-frequency coils, too. In the following section, a model of a coil is used to examine the relevant issues.

The typical MRI coil for high-field systems can be viewed, to a reasonable approximation, as a lossy inductor. The resistance comes from many sources: including components and conductor; inductively-coupled sample resistance; capacitively-coupled sample resistance; and radiation resistance.

If the system is too large, a lossy inductor may not be an adequate model because of non-uniform current on the coil structure. Generally, the coil should fit inside a sphere of diameter $\lambda_{2\pi}$ to be considered to be a small antenna and, therefore, to be represented using lumped elements [7]. The electrical size may differ from its physical size, making such a description difficult. Nonetheless, models of high-frequency MRI coils using lumped circuit analysis are adequate for many purposes. As was mentioned above, distributing capacitors around a loop forces the loop current to behave as a electrically small antenna.

First, consider a single-turn loop that has a single gap as shown in Figure 1. Measurement of the input impedance at the gap for a particular frequency will give a resistance and inductive reactance, assuming the small size as described above. Conventional matching techniques produc-

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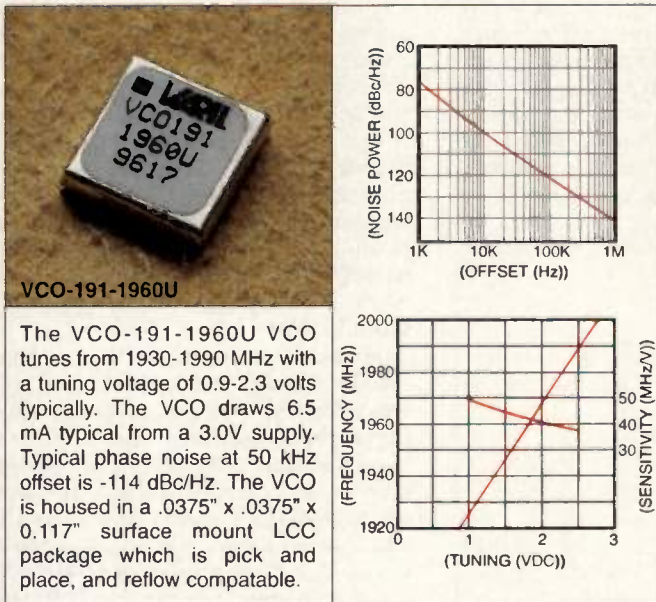
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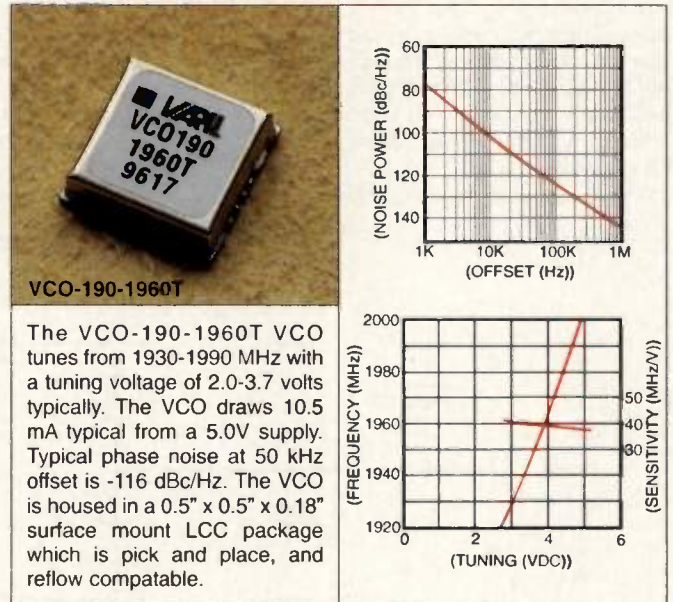


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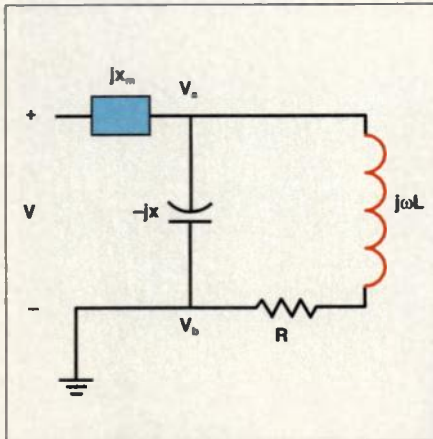


Figure 2. A schematic of a matched (to 50 Ω) circuit model of the loop of Figure 1.

the circuit model shown in Figure 2. Because the match is only adequate at a single frequency, reactive components are used in the simplest configuration, using two or three reactances.

It is interesting to examine the voltages at the gap of the loop. For this case $V_a = V(1 + jx_m/50)$ and of course $V_b = 0$. The voltage will decrease continuously from V_a to V_b around the loop. The loop will couple capacitively to anything at ground or virtual ground, with the bulk of the effect taking place from the top half of the loop.

Cable shields may have unbalanced current caused by this effect. A large, lossy dielectric, such as a person, will have current induced from this effect. Additionally, the current path through the sample may cause an electric dipole to be formed that could cause substantial radiation.

One method of reducing this capaci-

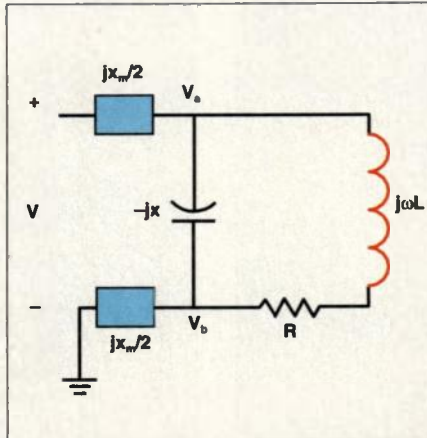


Figure 3. A more balanced version of the coil of Figure 2.

tive coupling effect is shown in Figure 3. This method results in the voltages $V_a = V(1 - jx_m/100)$ and $V_b = V(jx_m/100)$. The potential difference is the same, but some of the voltage is now antisymmetric around the center line of the coil. There still remains a voltage V , which is unbalanced. The improvements this circuit makes depend on the details of the coil system, but several points can be made.

First, the antisymmetric part of the voltage will not produce a net coupling to a cable that lies on the coil center line. Second, the antisymmetric part of the voltage that locally couples to the lossy dielectric produces half as much power loss as does the situation described above. Finally, antisymmetric current paths that pass through the dielectric tend to produce an electric quadrupole instead of an electric dipole. This greatly reduces radiation from

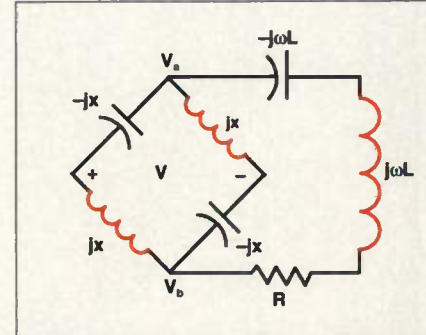


Figure 4. The component matched using a discrete component balun.

this effect. For the case of using the preamplifier for inductive decoupling most of the voltage is unbalanced because x_m is relatively low (typically 2 Ω or less).

Figure 4 shows a balun used for matching the coil as well as for balancing [3]. This circuit produces the voltages $V_a = V(1/2 - jx/100)$ and $V_b = V(1/2 + jx/100)$. This result is the same as above, except the unbalanced voltage has become symmetric and half as large in amplitude. This produces half as much power loss in the sample and half as much radiation from an effective electric dipole that has been halved. Coupling to a central cable still can take place.

Finally, in Figure 5, a circuit we have developed produces entirely balanced voltages. In this case, for the choice of $y = 50$, $V_a = -jV(1 + jx/100)$ and $V_b = jV(1 + jx/100)$. The circuit preserves the ability to decouple inductively the circuit using the low-impedance preamplifier. An additional $+90^\circ$ or -90° phase shift is required in comparison to the matching circuits in Figures 1–4 to produce high impedance from the loop perspective.

Conclusion

MRI coil development has evolved rapidly through the application of radiowave theory. A thorough understanding of these principles and technique becomes even more critical as the coil size to wavelength ratio increases for high-field systems and as multichannel receivers continue to proliferate. Based on principles of MRI coil design, a new circuit for MRI coil matching provides complete balance and allows preamplifier-based decoupling. Although techniques have been proven at lower frequencies, challenges still arise because coils are approach-

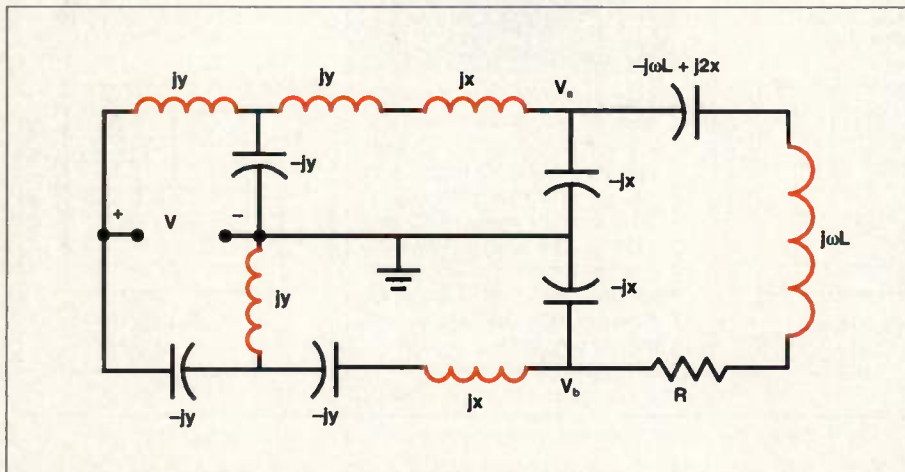


Figure 5. A new circuit for completely balanced matching of the coil.

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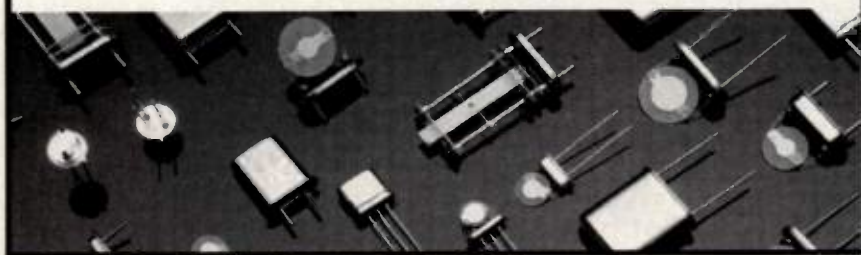
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INFO/CARD 59

λ at the higher frequencies. This will require more initial engineering design work to eliminate susceptibility to far field radiating effects and losses. Future explorations will include the optimization of coils with shields, correlation of noise from different coils related to common far-field sources and investigation of the human body as a secondary radiator.

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RF

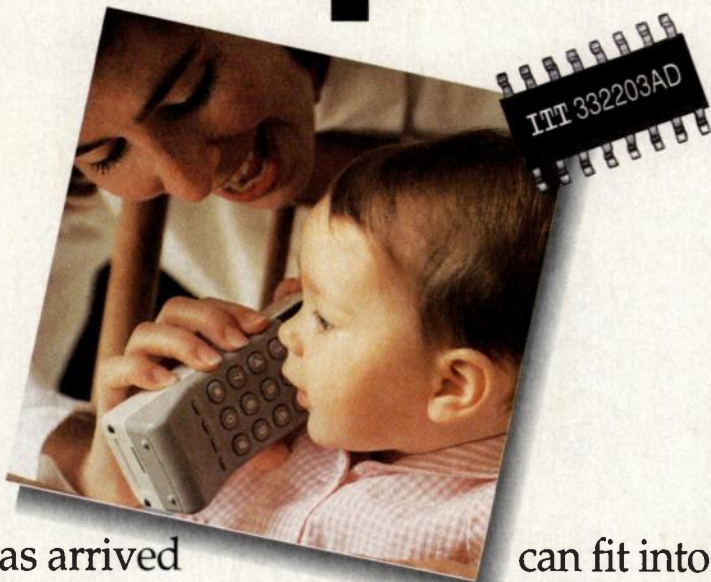
About the authors

David M. Peterson is an RF engineer for the University of Florida Brain Institute. He designs and improves RF coils and hardware for research on high-field MRIs.

G. Randy Duensing, Ph.D., is an assistant scientist for the University of Florida Department of Electrical and Computer Engineering and is president of Applied Resonance Technology, where clinical MRI coils are designed and produced for worldwide distribution.

J.R. Fitzsimmons, Ph.D., is a professor for the University of Florida Department of Radiology and is the technical director for the 3T high-field program at UF.

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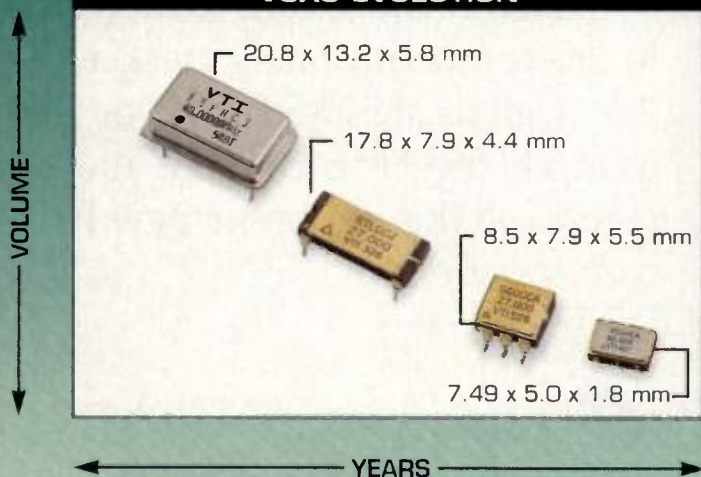


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NOT JUST PRODUCTS... MINIATURE VCXO SOLUTIONS

Manufacturers look to die technology to keep up with RF transistor trends

This month marks the return of the product forum to RF Design. Each month, this section will give the companies manufacturing the products used by RF engineers the opportunity to offer their opinions regarding today's marketplace and trends in the industry without editorial interpretation. This month, the product forum highlights the transistor market.

Hewlett-Packard

One of the challenges facing portable wireless manufacturers has been the tradeoff between an efficient, high-performance discrete design compared with an easier, smaller RFIC design. In addition to traditional discretes, HP is taking a self-biased transistor approach yielding simpler, smaller designs while still maintaining performance, efficiency and flexibility.

An example of this direction is a self-biased Silicon RF bipolar transistor in the miniature SOT-363 package. This product achieves performance and efficiencies approaching those of a pure discrete transistor, while removing the complexities and board space of a biasing circuit. The grounded emitter design allows control of collector current, allowing for H_{fe} variation and stabilizing current over temperature. Finally, it eliminates the cost, parts count and space required for as many as eight additional DC components.

This concept can be extended to SOT-363 packaged self-biased GaAs PHEMT products in development. In this case, they have the added advantage of requiring only a single positive supply. Their simple design and efficient die layout allow these silicon and GaAs parts to be priced below traditional wideband RFIC products, creating an innovative tool to meet aggressive system specifications and timetables.

NEC

NEC has 10 small-signal bipolars in the SOT-23 package style, with f_T s from 0–15.5 GHz. These are also available in the SOT-323 package style, which is 40% smaller than the SOT-23. NEC

has introduced the smallest transistor package, the 19 package, which is 40% smaller than the SOT-323. It measures 0.8×1.6 millimeters, and seven die types are available in this package style.

To give design engineers additional options in reducing board real estate, NEC has introduced a new series of dual-chip transistors in a six-pin SOT-363 package, (1.25×2.0 mm). These dual-chip transistors are available in two-pin configurations. These two-pin configurations are suitable for two-stage cascode LNA circuits or oscillator and buffer amplifier applications.

M/A-Com

The world of silicon high-frequency transistors can be divided as a first cut by power level; i.e., small-signal, low-power devices and large-signal, high-power units. This first cut segregation by power level is useful because it provides some insight into the applications and markets served by these different devices. In general, small-signal devices are used only as in the "receive" side of any given radio architecture. This receive function emphasizes parameters such as noise figure, associated gain, phase noise, f_T , F_{max} , S-parameters, low voltage and low current. This functionality contrasts markedly with high-power transistors where the primary, if not only, function is the "transmission" of signals. In this mode, parameters such as P_{1dB} , P_{out} gain, IMD, power-added efficiency, high voltage and high current become the driving factors in device performance.

In the realm of high-power transistors, silicon remains the only viable material and device technology for both radar and cellular base station applications. The only question at this point is whether the ubiquitous silicon bipolar transistor will remain the workhorse device or whether it will be supplanted by other silicon-based high-frequency structures. Leading candidates that have demonstrated performance and that realized a significant level of market acceptance, especially at the 900

MHz and 1.9 GHz cellular bands, are power MOSFETs, particularly vertically isolated LDMOS structures. Another possibility, based upon recent high-frequency, high-voltage developmental structures, is the SiGe HBT. Although adapting SiGe to the higher voltage requirement of power transistor applications tends to negate some of the raw frequency capability of the material, enough remains to provide significant frequency and power advantages over both standard BJTs and power MOSFETs, especially as the marketplace drives to higher and higher frequencies.

Motorola Semiconductor

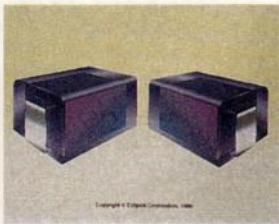
The RF market has become increasingly competitive. Cost, size, efficiency and availability have become the significant drivers of designs. Today's dynamic wireless portable marketplace is making special demands on both the required RF solutions and the companies that provide these systems. It is readily apparent that there is no one RF approach to the various wireless markets. The needed approach is to offer a wide selection of topologies, technologies and packaging that allows use in almost any of the RF design applications such as analog cellular, GSM cellular, DCS, PCS, cordless phones, RF modems, cable modems and two-way pagers.

To meet the ever-increasing demands of today's communication systems, new die technologies such as silicon FET and GaAs are being used, either separately or in combination. Addressing the need to reduce size as well as cost, these advanced technologies, which include silicon LDMOS, submicron bipolar, BiCMOS, GaAs MESFETs and PHEMTs, are being coupled with surface-mount packaging to produce cost-effective solutions in the wireless portable marketplace. To enable customers to compete in today's competitive wireless markets, it will become commonplace to see a variety of topologies, technologies and packages used together to form a total RF system.

RF

RF product/services showcase

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The EC3225 and EC4532 series of surface-mount inductors have an induc-

tance range of .012–220 μ H and 0.1–1000 μ H, respectively. Other features include a tolerance of $\pm 5\%$, $\pm 10\%$ or $\pm 20\%$ and an operating temperature range of -25°C to 85°C .

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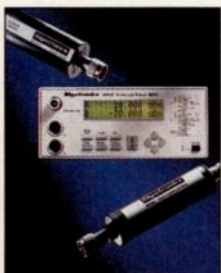
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INFO/CARD 157



Mmicad Layout, a software package designed to automate circuit layout

and artwork creation, features schematic capture with bidirectional links to layout, the Mmicad simulator, supports the use of as many as 64 layers and allow designs to be performed with as many as 20 levels of cell hierarchy.

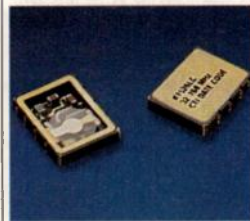
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The K1526 series of voltage controlled crystal oscillators (VCXOs) is compatible with today's

phase-locked loop (PLL) applications found in digital access radio, wireless base stations and telecommunication transmission equipment. Available for frequencies from 2–55 MHz, the ceramic package measures 0.56" \times 0.36" \times 0.16".

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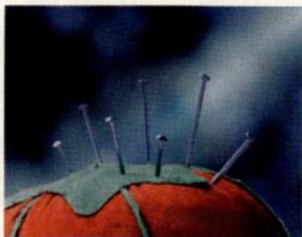
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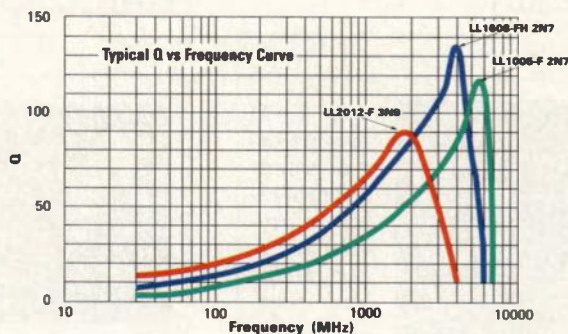
Such a micro size may seem impossible, but not to Toko engineers. The LL1005 (inductance range 1.0 to 27.0 nH) is merely the latest addition to Toko's micro-miniature line of ceramic multilayer chip inductors available from Penstock. With an 0402 footprint and an 0.05 profile, *the LL1005 is the world's smallest—by far.*

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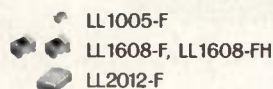
- Available 0402, 0603 and 0805 packages assure wide flexibility
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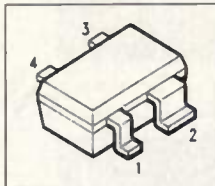
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The BFP 405, BFP 420 and BFP 450 are three Siemens grounded emitter transistors (SIEGET) featuring a grounded emitter configuration, high-transition frequency and high reliability. Designed for high-gain, low-noise and low-power consumption applications, they operate as high as 12 GHz.

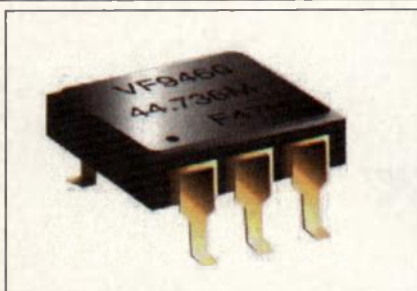
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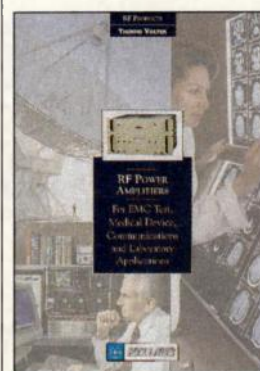
The VF946 s-type crystal oscillators feature HFFX crystals which achieve greater pulling on higher frequencies with low phase noise.

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The SA038WA-008 surface acoustic wave (SAW) filter takes up less than half the board space of commonly used SAW filters. The filter features a miniature LCC package and is designed for digital audio broadcasting (DAB) applications.

Toko America
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The RF power amplifier catalog features EMC test amplifiers, communications amplifiers, pulsed amplifiers, dual band amplifiers, distributed tube amplifiers and

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The UPB 1004GS receiver-on-a-chip for GPS applications combines a double RF/IF downconversion block and

PLL frequency synthesizer in a 30-pin SSOP package. Designed to convert a 1575.42 MHz input to an IF as high as 5 MHz, it delivers 92.5 dB typical total conversion gain. Features include spurious signal rejection, high sensitivity and low power consumption.

California Eastern Labs
INFO/CARD 168

RF products

Surface-mount VCXO offers power-saving enable and disable control

Designed for use in high volume portable products, the VX-8000 series surface-mount voltage-controlled crystal oscillator (VCXO) features a J-lead package profile height of 0.185" and a flat pack profile height of 0.150". The VX-8000 series has enable and disable control, which allows the user to stop clocking unused complementary metal oxide semiconductor (CMOS) parts, thus helping to reduce power consumption. They are available at any frequency from

1-160 MHz, which includes 155.52 MHz for use in synchronous optical networks (SONETs) and asynchronous transfer networks (ATMs) and operate at either 3.3 or 5 V. Linearity is $\pm 10\%$ or better. Stability is within $\pm 25\%$ higher than the operating temperature range of 0°C to 70°C. The VX-8000 series is priced as low as \$12 each in quantities of 100,000 depending on frequency, package and configuration. **Raltron Electronics**
INFO/CARD 169



In-line EMI/RFI shielded enclosures

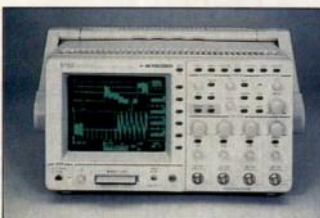
Allowing products to undergo accurate, efficient testing without removing the product from the manufacturing line, these enclosure designs test in-line electromagnetic interference and radio frequency interference (EMI/RFI) of wireless and cellular products. Designed using welded metal construction, the



in-line enclosures meet the specific automation and throughput needs of the manufacturer. The enclosure is designed to interface with the manufacturer's programmable logic control (PLC) used to control a complete automated test sequence. **Lindgren RF Enclosures**
INFO/CARD 170

Four-channel digital oscilloscope

Model 5150, a four channel digital storage oscilloscope, features a 50 GS/s



sampling rate for repetitive waveforms and a 200 MS/s sampling rate for non-repetitive waveforms. Other features include: automatic set-up of timebase; 40K of internal memory with a built-in Personal Communication Memory Card Industry Association (PCMCIA) slot that provides an additional 1 MB of memory per card; and as many as 16 automatic measurements. Stored signals can be compared, manipulated, displayed and transmitted in a variety of formats, while selected portions of measured waveforms can be compared directly with waveforms stored in memory. The 5150 is priced at \$4,995. **B+K Precision**
INFO/CARD 171

SP8T GaAs MMIC switch

The non-reflective HMC 183QS24 single pole eight throw (SP8T) gallium arsenide (GaAs) monolithic microwave integrated circuit (MMIC) switch operates from DC to 2 GHz, covering cellular, personal communications service (PCS) and dual-band designs. Designed to replace discrete positive intrinsic negative (PIN) diode or multiple SP4T MMIC switch designs with a single easy-to-interface device, the switch features



isolation of 40 dB at 1 GHz and 32 dB at 2 GHz, insertion loss of 1.3 dB at 1 GHz and 1.7 dB at 2 GHz and switching times typically about 25 nanoseconds. **Hittite Microwave**
INFO/CARD 172

Power entry filters designed for EMI

A complete line of power entry filters designed for general-purpose broadband



electromagnetic interference (EMI) filtering includes power entry modules, printed circuit board (PCB) mount power filters and power line filters. Power entry modules are rated for voltages as high as 10 A and are available with a variety of terminal styles. Miniature PCB-mounted power filters are rated for AC or DC currents as high as 3.6 A and are standard for two-wire cord systems. Power line filters are rated for AC or DC currents as high as 30 A and will filter differential mode and common mode noise. **Spectrum Control**
INFO/CARD 173

CABLES AND CONNECTORS

Adapters for waveguide to coaxial connectors

Two series of right-angle-launch and

end-launch adapters from WR22 waveguide to 2.4 millimeter coaxial connectors feature a connection that offers 30–40 dB improved connection repeatability over conventional flanges. The J236 series of right-angle-launch adapters features a voltage standing wave ratio (VSWR) of 1.15 typical

across the entire waveguide band. The J237 series also offers low VSWR with the added convenience of an in-line design.

Maury Microwave
INFO/CARD 174

Shielded stereo jack for audio applications

The ST-3200 stereo jack, designed for audio applications, has a 3.5 millimeter high profile for PCB layout requiring 6.5 millimeter height off the board. The threaded or non-threaded bushing provides grounding protection. The ST-3200 is shielded to provide electromagnetic interference and radio-frequency interference (EMI/RFI) protection. Both three and five positions are available.

Kycon Cable & Connector
INFO/CARD 175

Coaxial D connectors for data, computer applications

A 75 Ω coaxial D connector system allows coaxial cables to be used with normal data lines with only a single I subminiature plug and socket. Designed for data and computer applications, they are priced less than \$5 each.

Connect-Tech Products
INFO/CARD 176

DISCRETE COMPONENTS

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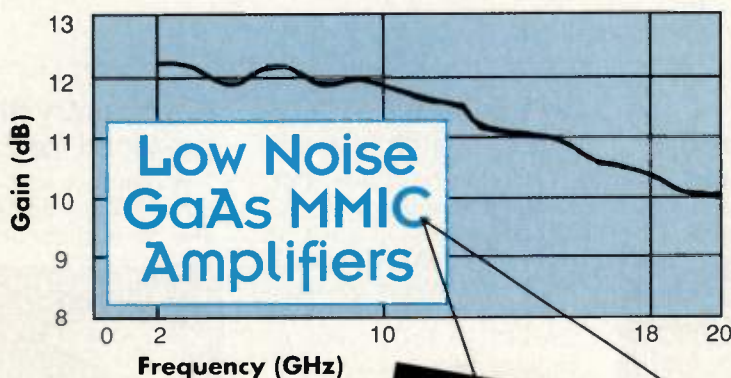
Type ESR polymer aluminum capacitors offer lower impedance at frequencies of 100 kHz and higher when compared to tantalum and conventional



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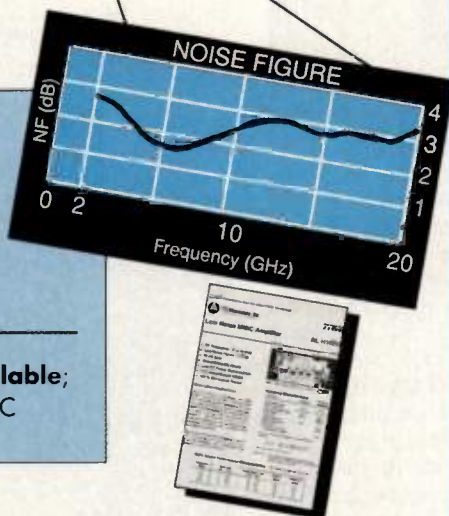
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Cornell Dubilier
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**Quarter-watt resistor
provides high performance**

The WSL-1206 chip-style resistor offers an ohmic value (0.007–0.2 ohms), inductance (0.5–5 nH) and the precise stability ($\pm 1\%$ tolerance, ± 75 ppm/ $^{\circ}\text{C}$ temperature coefficient) normally associated with only wirewound resistors. Designed for battery monitoring in wireless applications, it also features a small size (EIA 1206) and is priced at 18 cents each in quantities of 10,000.
Dale Electronics
NFO/CARD 178

**Surface-mount inductors
or RF applications**

The SML 44 series of high-current inductors features a small height (4.5 millimeters) and range from 10 μH @ .3 A to 680 μH @ 0.160 A. These inductors, designed for SMPS and RF applications, use a ferrite bobbin mounted to a ceramic substrate and are priced at 43 cents each in quantities of 10,000.
Howanda Electronics
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SUBSYSTEMS

**SM antenna
designed for WLAN**

Designed for 2.4 GHz industrial, scientific and medical (ISM) band applications, such as wireless local area network (WLAN) and wireless modem, the Jova Comm miniature antenna features a gain of 1.8 dBi (typical), 50 Ω impedance and a voltage standing wave ratio (VSWR) of ≤ 2 . The antenna is priced at \$5.95 in quantities of 10,000.
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NFO/CARD 180

**Downconverter module
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Model DA4 series four-channel downconverter module operates over IF and LO frequency bands in the range of 0.5–18 GHz. Standard designs include input limiter protection diodes and an IF amplifier. This unit can be used for basic direction-finding equipment or as an integrated receiver front

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Miteq
INFO/CARD 181

**GPS antenna
designed for surveying**


Designed to provide superb tracking

capability, the DM C146-20 broadband antenna has a minimum gain of -4 dBic at 5° above the horizon. It operates in the 1.2–1.625 GHz global positioning system (GPS) frequencies and has a small radome with a low profile.
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
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2000 features a maximum power input of 10 W in 50 Ω systems and is priced at \$94.95 each in quantities of fewer than 10.

Mini-Circuits
INFO/CARD 183

Fixed channel up/down frequency converters

Models PUL 070D, PDL 070D data grade frequency up/downconverters feature an input or output frequency of 7 or 140 MHz and an output or input frequency of a customer-specified L-band channel within the 950–1,450 MHz frequency range. Both models feature an input return loss of 9 dB nominal and an output return loss of –13 dB nominal, and measure 1.75" \times 1.9" \times 1.2".

Quintech Electronics and Communications
INFO/CARD 184

Subminiature SAW IF filters for GSM

A surface acoustic wave (SAW) filter for global system for mobile communications (GSM, formerly Group Speciale Mobile) applications is designed for use with the new generation GSM chip set architectures requiring higher-frequency intermediate frequencies (IFs) in the 240 MHz range. The filter features a center frequency of 246 MHz with a passband of ± 80 kHz. The rejection bandwidth is 25 dB minimum at ± 40 kHz and 45 dB minimum at ± 80 kHz. Typical insertion loss is less than 6 dB. Designed for use as a IF bandpass filter for personal communications service (PCS)-1900 and digital communications service (DCS)-180 applications, it measures 9.1 \times 4.8 \times 1.1 millimeters.

Toko
INFO/CARD 185

Varactor series meets wireless design needs

The MA4ST200 series of high quality factor (Q), high capacitance ratio varactor diodes enables applications such as cellular phones, active pagers and automotive wireless systems to seek the highest quality signal and lock onto the desired frequency. Designed for voltage control oscillators (VCOs) and voltage-tuned filters used in battery-operated wireless systems as high as 2.5 GHz, the diodes typically have a Q of 400 \pm

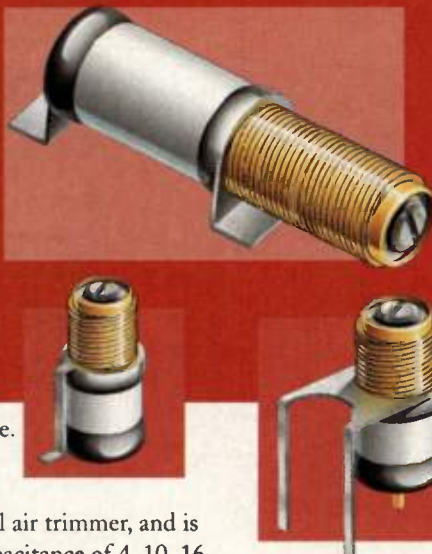
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INFO/CARD 62

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V/A-Com
INFO/CARD 186

**Separate filters
isolate frequency bands**

The PTD87/39NF diplexer uses one bandpass filter for the 1,375-1,400 MHz frequency band and uses another bandpass filter for the 1,427-1,452 MHz personal communications services (PCS) frequency band. The diplexer features a passband insertion loss of less than 1 dB, an ultimate stop band attenuation of 80 dB minimum and an RF power capability of greater than 15 W.
Penny Technologies
INFO/CARD 187

SIGNAL SOURCES

**Product combines
TCXO, OCXO technologies**

A family of products based on resonator-thermostat (RT) technology incorporates a directly heated quartz crystal, a temperature sensitive element and a thermocontroller circuit




sealed in one enclosure. The family features the performance of an oven controlled crystal oscillator (OCXO) while maintaining the size, power consumption and warm-up time of a high-end temperature controlled crystal oscillator (TCXO). Specifications include 90 nW power consumption, phase noise of -155 dBc/Hz @ 10 kHz offset and a frequency range of 8-25 MHz for the RT, OCXO and 2-105 MHz for the TCVCXO.
Valpey-Fisher
INFO/CARD 188

**VCXO designed
for PLL applications**

The K1526LC series of surface mount (SMT) voltage controlled crystal oscillators (VCXOs) is designed for phase-locked loop applications used in clock recovery, signal tracking and local refer-

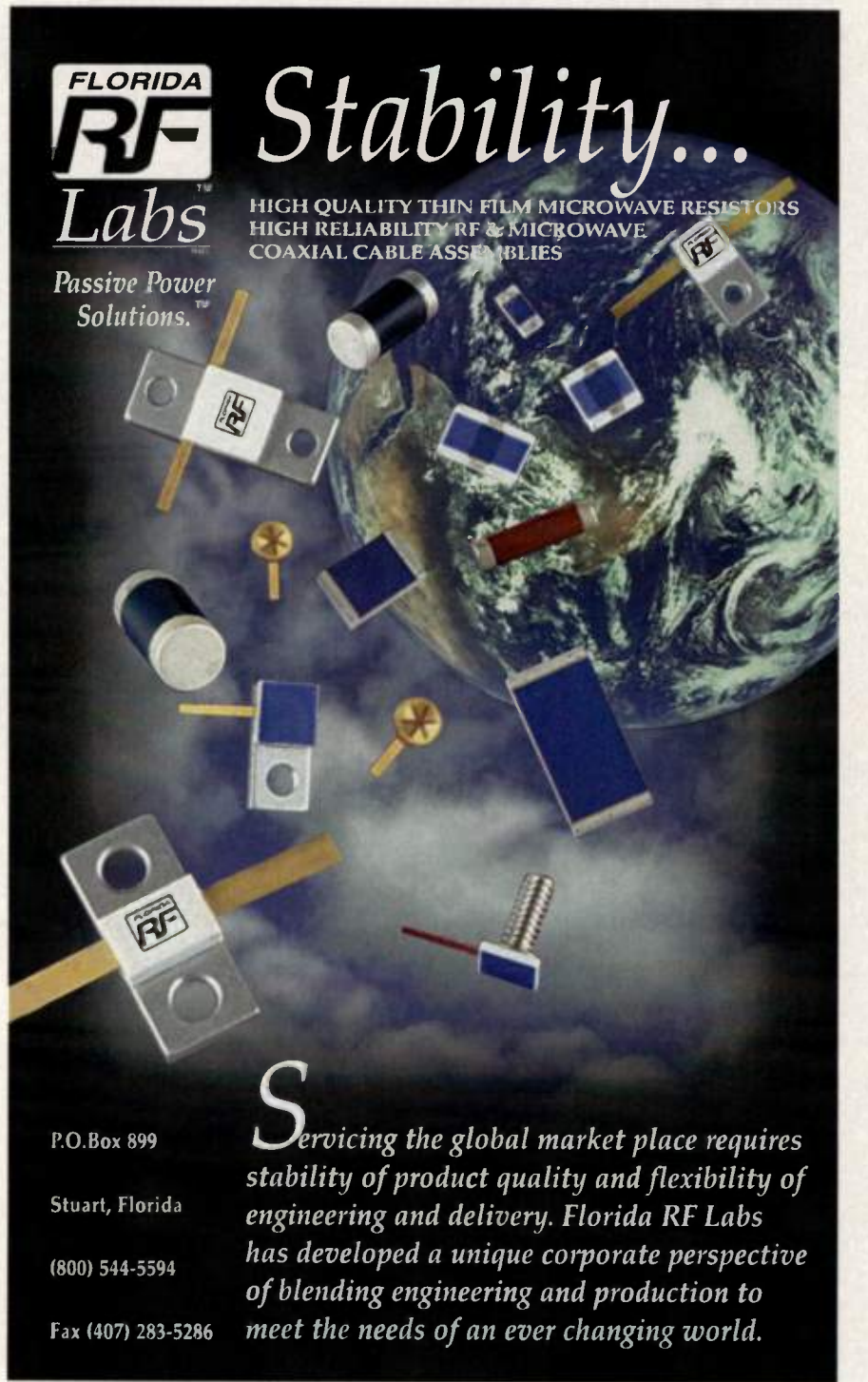
ence clocking circuits. The VCXO delivers a TTL/CMOS compatible output on frequencies ranging from 2-40 MHz. Requiring 5.0 V input, it provides a typical pull range of ± 120 ppm over a control voltage range of 0.5-4.5 V.
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RF software

Analog simulation, schematic entry integrated

Intusoft's series of analog and mixed-signal design tools integrate analog and mixed-signal capability into OrCAD Capture, Protel Schematic and Viewlogic Workview Office schematic tools. The software packages are called ICAP for OrCAD, ICAP for Protel and ICAP for Viewlogic. Each package includes the IsSpice4 native mixed-mode (analog and digital) circuit simulator, SPICE model libraries with more than 8,000 analog and digital parts; the IntuScope graphical waveform analyzer, the SpiceMod Spice modeling program, a set of symbol libraries and the Intusoft Integration Module, which allows engineers to run SPICE simulations directly from the OrCAD, Protel or Viewlogic schematics and to cross-probe the results in real time.

Intusoft
INFO/CARD 160

Real-time development environment available

Hyperception's real-time integrated development environment (RIDE) is a visual environment optimized for the design, implementation and analysis of real-time digital signal processing (DSP) algorithms and systems. Its power lies in its visual nature and support for industry-standard plug-in DSP boards. Hypersignal RIDE allows DSP applications to be generated quickly with little or no software coding required. RIDE's support of several different DSP chip families from different semiconductor companies, as well as its support for different DSP board vendors, makes it suitable for many real-time DSP projects. The device-independent approach allows different types of DSPs to be used in the same design. The ability to move designs from one DSP technology to another in the same environment means that users do not have to learn multiple tools, and they can upgrade their designs in the future for more performance.

RIDE was created by combining DSP hardware with Hyperception's hypersignal block diagram application and Windows DSP board drivers. These board drivers are installed from the Windows control panel just like other peripheral devices. The driver handles all communications and control of the DSP hardware from the P environment. The block diagram appli-

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ation does not require information about what DSP hardware is being used. The RIDE driver links DSP object files, downloads code, data and parameters to the DSP memory, controls the execution of the DSP and monitors activity on the DSP.

The user interface is the same for both simulated and real-time DSP block functions. This allows for convenient conversions between design simulations and real-time implementations.

Hypersignal RIDE is available for \$3,995. An automatic C code generator can be purchased with RIDE for \$5,000. Bundled packages including all hardware and software also are available.

Hyperception
NFO/CARD 161

Software automates immunity testing

Amplifier Research (AR) SW1000 testing software permits computer control of AR power amplifiers, signal generators and other equipment for a range of immunity testing requirements, including IEC-1000-4-3. The software uses an IEEE-488 communications link with a power meter or field monitoring system to level by power or field strength.

The SW 1000 software is designed to operate as a stand-alone program for Windows 3.1 and 95. It was developed under the National Instruments (NI) Labview environment and can be run in combination with more than 500 instrument drivers. The software permits use of as many as eight digital output channels for control of test instruments.

Amplifier Research
NFO/CARD 162

Evaluation software demonstrates capabilities

Labview Evaluation software on CD or Windows, Power Macintosh, Sun parc stations and Hewlett-Packard 9000 series workstations is a free graphical instrumentation software package. Users can access details about Labview add-on toolkits, technical support and customer education. Applications involve instrument control, data acquisition, analysis and presentation or test and measurement industrial automation. On-line tutorials guide users step-by-step through building a custom virtual instrumentation system.

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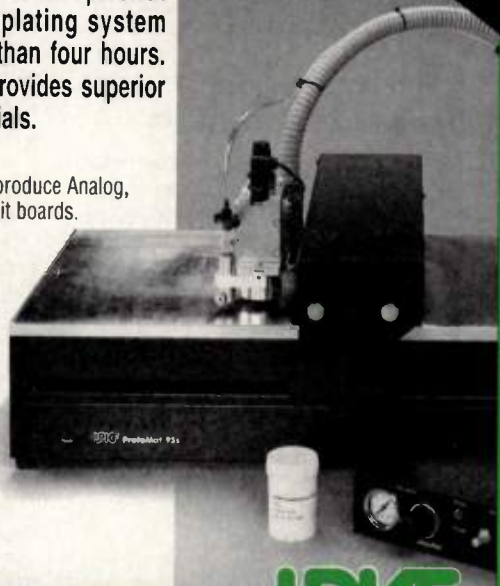
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RF literature

TIA publishes interim standards

The Telecommunications Industry Association (TIA) published two interim standards. TIA/EIA/IS-96-B, speech service option standard for wideband spectrum systems, forms a standard for service option 1, which provides two-way voice communications between a base station and a mobile station using the dynamically variable data rate speech code algorithm. The transmitting speech code takes voice samples and generates an encoded speech packet for every traffic channel frame. The receiving station generates a speech packet from every traffic channel and supplies it to the speech code for decoding into voice samples.

TIA/EIA/IS-126-A is the mobile station loopback service options standard. It provides the basis for a loopback of primary traffic information bits through a mobile station. It also provides the means for a base station to supply a known data stream on both the forward and reverse traffic channels, so that a mobile station's receiving and transmitting performance can be measured. The service option also provides a convenient means of setting up calls and generating traffic for system testing, and it allows for normal operation of signaling messages and secondary traffic.

TIA
INFO/CARD 164

Components catalog features terminal blocks

A 45-page catalog of electronic components featuring Magnum terminal block line is available. Complete specifications, illustrations, photographs and ordering information for products are included. The company's wire-ready option is available on all single-row terminal blocks. Other products offered are double-row terminal blocks, edge board connectors and Mag-Master electronic and field wiring interfaces.

Bussmann Circuit Components
INFO/CARD 165

Products designed to eliminate EMI, RFI

A 44-page catalog details a line of power products designed to eliminate electromagnetic interference (EMI) and radio frequency interference (RFI). This line of power products includes high-

current, single-line filters, power arrays, power-entry filters, power-line filters, multisection filters, custom assemblies and power distribution systems.

The catalog provides features, applications, order numbers, performance specifications, temperature characteristics, insertion loss curves, circuit diagrams and line drawings for each of the product lines. Also covered in the brochure is the company's EMI filtering experience and electromagnetic compatibility (EMC) testing services, application guidelines for using EMI filters, EMI measurement guidelines and a design inquiry form.

Spectrum Control
INFO/CARD 166

Modulator and demodulator products described

A free brochure from Stanford Telecom presents modulator and demodulator application specific integrated circuits (ASICs) and board-level assemblies for hybrid fiber and coax (HFC) upstream community antenna TV (CATV) systems. The four-page brochure describes the company's capabilities in the interactive cable transmission and reception segment.

Information is included on the STEL-1108 bipolar phase shift keying (BPSK) and quadrature phase shift keying (QPSK) digital modulator ASIC

chip used in subscriber modems and set-top boxes.

Stanford Telecom
INFO/CARD 167

Catalog details signal components, subsystems

Technical Research and Manufacturing's 100-page signal-processing components and subsystems catalog detail an array of products. The catalog is arranged in nine sections comprising power dividers, directional couplers, hybrids, mixers, transformers, phase comparators, modulators and beam forming networks.

Technical Research and Manufacturing
INFO/CARD 168

Filter products catalog released

Micro-Coax's *Filter Products Catalog* features the company's family of In-4 Cable filters, which combine Chebyshev and other filter types with semirigid or flexible cables to produce performance in the same space as the cable alone. The free 38-page catalog includes technical information, specifications and application information about all of the products.

Micro-Coax
INFO/CARD 169

On line:

ASIC products on web page—Qualcomm's web page features a line of synthesizer, forward error control, voice compression, automatic gain control and code division multiple access (CDMA) application specific integrated circuit (ASIC) products developed for advanced communications systems.

Included in the ASIC web site are an overview of each product, worldwide sales information, new product releases and the option to download any technical data sheet or application note. The ASIC products can be found within the Qualcomm web site at <http://www.qualcomm.com/ProdTech/asic>.

Qualcomm
INFO/CARD 170

Home page redesigned—Anadigics' web site provides corporate in-

formation, company news, information on products, links, industry news and engineering tools. The web site also features Java applet engineering tools to help designers evaluate and model receiver performance in communications systems. The site is located at <http://www.anadigics.com>.

Anadigics
INFO/CARD 171

Free catalog available—Time Motion Tools' catalog features telecommunication, coaxial, test and measurement meters, including the Fluke 7-300. This low-voltage (as high as 300 volts AC or DC) meter automatically decides between measuring volts, continuity or ohms. The company's web site is located at <http://www.timemotion.com>.

Time Motion Tools
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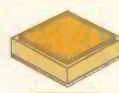
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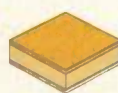
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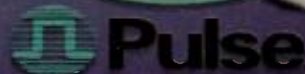


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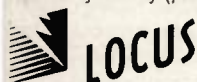
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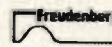
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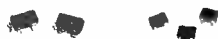
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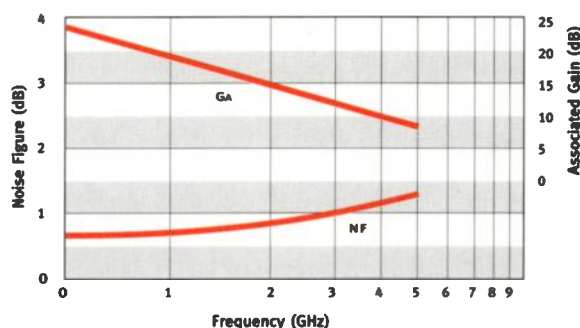
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