

# *RF design*

*engineering principles and practices*

*December 1990*



*Featured Technology*  
**IC Applications**

*Reader Survey*  
**Attenuators and Switches**

**1989-1990 Article Index**



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*Cellular...  
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*The FM/AM-1200S is still the most feature intensive of all communications service monitors.*

**See for yourself.** The FM/AM-1200S includes a 1 GHz spectrum analyzer, 1 MHz scope, 2  $\mu$ V receiver, duplex capability, 150 watt power meter, RS-232 interface and numerous programmable functions—all standard in one self-contained, compact and portable unit. With **three new options**, the 1200S can now test cellular mobiles; CLEARCHANNEL LTR® compatible mobiles/bases; and duplexers, combiners and isolators.

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How many times have you wished for a convenient, inexpensive way to observe the frequency response of a cavity or duplexer? The new 1200S tracking generator option will do just that. By connecting two cables, you can observe the frequency response directly on the 1200S' CRT, providing for instantaneous monitoring of filter characteristics.

Where else can you get all this capability in one communications test set? No where! Compare and you'll agree...the FM/AM-1200S **is** the logical choice.

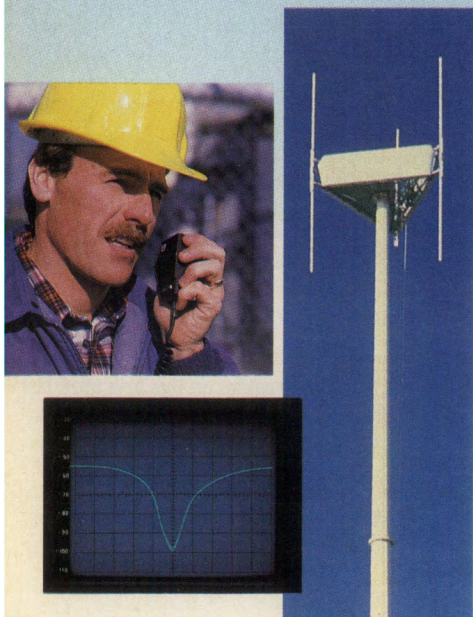
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$$t_{\text{Rise/Fall}} = 0.00000000007$$

$$t_{\text{Rise/Fall}} = 0.7 \times 10^{-9} \text{ seconds}$$

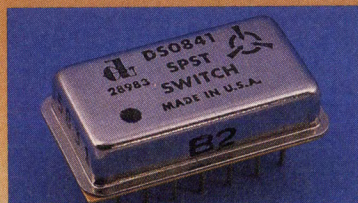
$$t_{\text{Rise/Fall}} = 0.7 \text{ nanoseconds}$$

$$t_{\text{Rise/Fall}} = 700 \text{ picoseconds}$$

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## DAICO Introduces The 700 Picosecond Ultra-Fast Switch

The new DAICO DSO841 offers incredible switching speed. It's much faster than a Schottky diode switch drawing less than 2mA. In fact, Schottky diode switches can draw as much as 100-200mA.



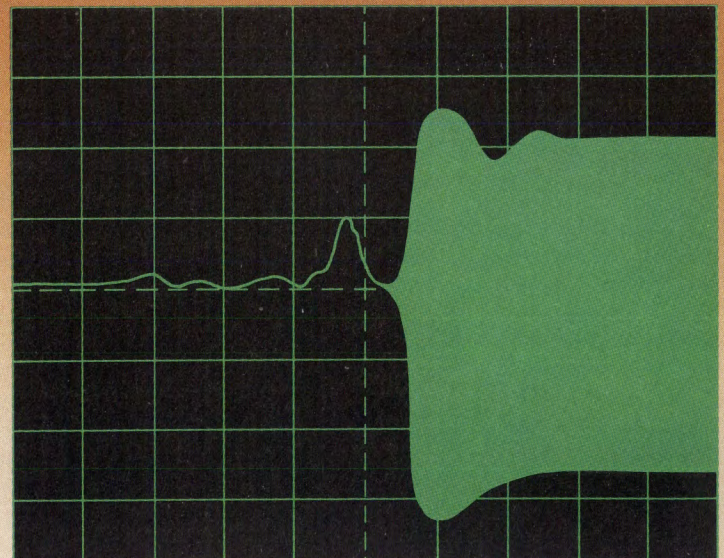
RF envelope rise time (10%/90% RF) for the DSO841 is typically less than 700 psec with 7 nsec switching speed (50% TTL to 90% RF).

DSO841 also offers excellent isolation of 70dB typical at 100MHz typical and 60dB typical to 200MHz.

The DSO841 is the ideal high speed, low transient, low power consumption switch for blanking and modulating applications.

### Operating Characteristics

PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
Current Drain		2	5	mA	AT +5V DC Supply
Switching Transients		56	100	mV	Peak Value
Transition Time		.7	1	nS	90%/10% or 10%/90% RF
Switching Speed		7	10	nS	50% TTL to 90/10% RF
Insertion Loss		1.7	2.3	dB	
Isolation	60	70		dB	10-100
	50	60		dB	100-200
Operating Frequency	10		200	MHz	
TTL Controlled					



### Scope Specifications

Ch.1 = 100.00 mvolts/div Timebase = 1.00 nsec/div Offset = 20.00 mvolts Delta T = 480.00

Call DAICO today at (213) 631-1143 for more information on this new ultra fast switch and our complete line of RF Attenuators, Switches and Phase Shifters.



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### featured technology

#### 27 Circuits for Wide-band FM Demodulation

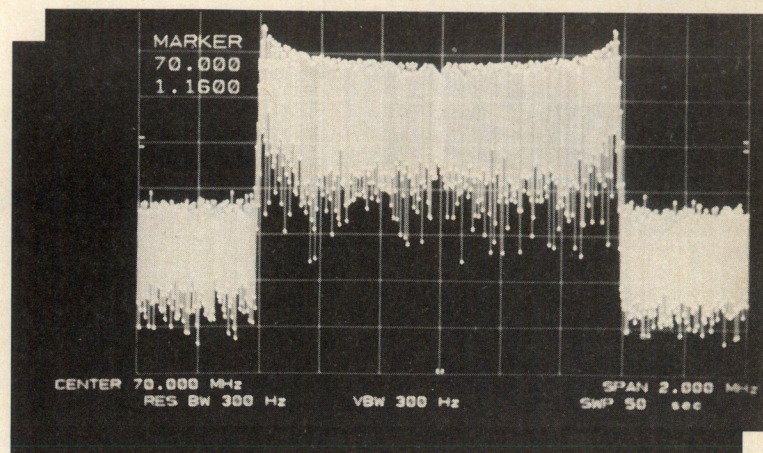
The integrated circuits discussed in this article can be used as wide-band discriminators in applications such as telemetry, high-speed data transmission, and FM over fiber. The designs presented give stable IF amplifier gains at 21.4 MHz and 70 MHz.

— Alvin Wong and Ali Fotowat

#### 47 Using Current Feedback Amplifiers

Current feedback amplifiers offer higher speed, nearly constant bandwidth at increased gains, and operate at higher frequencies than traditional op amps. This article describes these benefits as well as some applications of a specific current feedback amplifier.

— Al Little



### cover story

#### 52 A New 100 Hz to 26.5 GHz Spectrum Analyzer

Advantest America has released the R3271, a spectrum analyzer that covers the 100 Hz to 26.5 GHz frequency range. The analyzer has some interesting specifications and is ideal for detecting individual signals in a complex waveform.

### emc corner

#### 56 Spread Spectrum ASIC Eases Design of Low Cost Part 15 Systems

This article presents a custom-designed integrated circuit for spread spectrum applications. A tutorial on spread spectrum is a major part of this contribution.

— Raymond Simpson

#### 64 Crystal Delay Equalizers

This article describes a process of combining a number of equalizers into a single section, putting multiple crystals in parallel in each of the two branches of a half-lattice, thereby minimizing the number of added inductors.

— William Lurie

#### 67 Index of Articles: 1989-1990

This detailed list includes all articles and editorials published in *RF Design* for 1989 and 1990. It is arranged according to subject.

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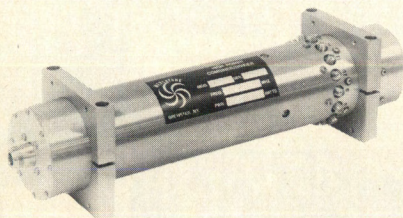
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## HIGH POWER 16 WAY COMBINER



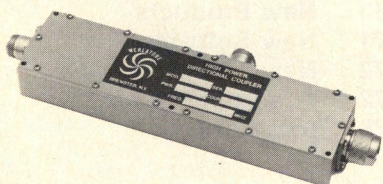
### TYPICAL SPECIFICATIONS

#### MODEL D2599

FREQUENCY RANGE ..... 0.4 - 1GHz  
INSERTION LOSS ..... 0.5db  
ISOLATION ..... 25db  
VSWR ..... 1.3:1  
POWER ..... 400 watts

The model D2599 features full power isolating terminations which maintain impedance match and isolation in "soft failure" modes.

## PRECISION DIRECTIONAL COUPLER



### TYPICAL SPECIFICATIONS

#### MODEL C2523

FREQUENCY RANGE .... 100-400 MHz  
COUPLING ..... 30db  
DIRECTIVITY ..... 35db  
VSWR ..... 1.1:1  
POWER ..... 750 watts

The model C2523 features exceptional coupling linearity vs input power and non-destructive precision stainless steel connectors.

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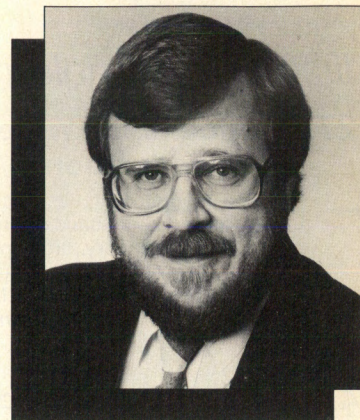
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decades ahead

## RF editorial

# RF-on-a-Chip

By Gary A. Breed  
Editor



When did electronics engineers start considering integrated circuit implementation for nearly any product? I'm not talking about mass-produced calculators and watches or development of ICs meant to be sold as commodities. I mean ICs that an engineer can specify to do the particular job he is working on.

Actually, it was nearly 20 years ago that digital product developers began to include programmable array logic (PAL) devices in commercial products. These were custom devices in that they could be either field or mask programmed to perform a specific set of logic instructions. Since then, we have seen plenty of ASICs (application-specific integrated circuits) with mix-and-match logic functions, plus universal gate array devices with thousands of individual gates that can be interconnected to do a huge number of different tasks.

More recently, digital designers have gained the ability to specify start-from-scratch ICs to meet their requirements, using some impressive computer-aided design tools to develop and debug the logic, lay out the circuit on a silicon substrate, and control the processing of the wafer. RF designers have seen some of these custom digital products for phase locked loops, direct digital frequency synthesis, and digital signal processing.

The next step combined analog and digital components, adding mainly op amps and comparators. Low-frequency signal processing, control devices, and data acquisition circuits became prime applications for this technology. At high frequencies, however, such circuits were power-hungry. The development in just the last couple of years of very high  $F_T$  PNP transistors has opened up a whole new set of applications, as complementary analog and digital circuitry can now operate into the hundreds of MHz, with its inherently lower power consumption.

This type of circuit has only recently reached wide acceptance.

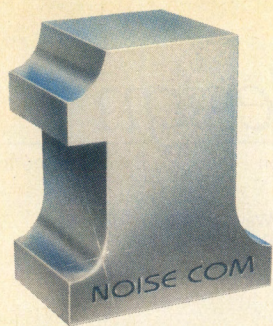
The final step in making full-custom RF ICs accessible is coming together right now. Advanced chip design software tools — like the ones used to design nearly every digital and analog IC — are incorporating true RF characterizations in their models. For analog functions, Spice-based routines are used in these CAD packages. The focus of recent efforts has been the development of Spice-compatible models for transmission lines, S-parameter device characterization, parasitics and discontinuities. With these tools, it is now possible to design ICs for analog or digital operation beyond 1 GHz.

GaAs ICs and MMICs are working their way toward a similar level of CAD modeling and development, albeit from a different starting point. These high performance RF ICs add even more options to an engineer's design arsenal.

Inevitably, technology moves ahead. RF has gone from spark to vacuum tubes, from vacuum tubes to transistors, from individual transistors to functional blocks of ICs and MMICs, and now we have the potential for completely integrated RF systems. We will certainly find some of the new technologies I noted last month implemented completely in silicon or GaAs.

It's hard to believe that some folks still think of RF as "old" technology!





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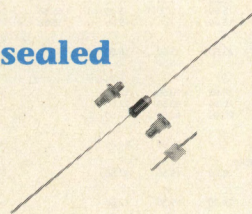
- Chips and five package options
- Flat output
- Operating temperature -55 to +125° C
- Storage temperatures -65 to +150° C
- Meet MIL-STD-750
- 50-ohm impedance
- Wide variety of packages

## Drop-in noise modules for BITE applications

### NC 500 Series

200 kHz to 5 GHz

- Drop-in TO-8 packages
- Economical solution to BITE
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- Storage temperature -65 to +175° C
- Temperature coefficient 0.01 dB/° C
- 50-ohm output impedance

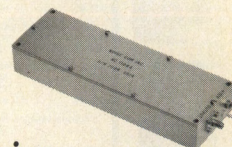


## Amplified noise modules

### NC 1000 Series

DC to 2000 MHz

- High power—up to 13 dBm
- Minimum crest factor 5:1
- Flat response—  $\pm 0.75$  dB
- Operating temperature -35 to +100° C
- Storage temperature -65 to +150° C

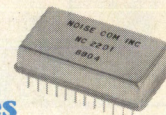


## Plug-in amplified noise modules

### NC 2000 Series

100 Hz to 300 MHz

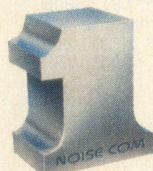
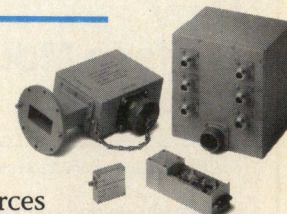
- 24-pin packages
- Minimum crest factor 5:1
- High-end roll off 6 dB per octave
- Operating voltage of +15 VDC, +12 VDC optional
- Storage temperature -65 to +125° C
- Operating temperature -55 to +85° C
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Fixed Attenuators, 1 to 20 dB:										
AT-50 (3)	50 (5W)	DC-1.5GHz	17.50	29.00	22.00	20.00	-	-	-	-
AT-51	50 (5W)	DC-1.5GHz	15.00	26.00	19.50	17.50	-	-	-	12.00
AT-52	50 (1W)	DC-1.5GHz	20.50	29.00	26.00	22.00	-	-	-	-
AT-53	50 (1W)	DC-3.0GHz	20.50	26.00	-	15.00	-	-	18.00	-
AT-54	50 (1W)	DC-4.2GHz	-	-	-	20.00	-	-	-	-
AT-55	50 (1W)	DC-4.2GHz	-	-	-	19.20 (10PC)	-	-	-	-
AT-75 or AT-90	75 or 93 (1.5W)	DC-1.5GHz	17.50	26.00	45.50	19.50	-	-	-	-
Detector, Mixer, Zero Bias Schottky:										
CD-51, 75	50 75	01-4.2GHz	64.00	-	-	64.00	-	-	-	-
DM-51	50	01-4.2GHz	-	-	-	64.00	-	-	-	-
Resistive Impedance Transformers, Minimum Loss Pads:										
RT-50/75	50 to 75	DC-1.5GHz	17.50	26.00	45.50	17.50	-	-	-	-
RT-50/93	50 to 93	DC-1.0GHz	17.50	26.00	45.50	17.50	-	-	-	-
Terminations:										
CT-50 (3)	50 (1.5W)	DC-4.2GHz	11.50	15.00	15.00	17.50	-	-	-	-
CT-51	50 (1.5W)	DC-4.2GHz	9.50	12.00	14.00	9.50	-	-	9.00	-
CT-52	50 (1W)	DC-2.5GHz	10.50	15.00	15.00	13.00	-	15.50	-	-
CT-53M	50 (1.5W)	DC-4.2GHz	5.60 (10PC)	-	-	5.60 (10PC)	-	-	-	-
CT-54	50 (2W)	DC-2.0GHz	14.00	15.00	15.00	17.50	-	-	-	-
CT-75	75 (25W)	DC-2.5GHz	10.50	15.00	15.00	13.00	-	15.50	-	-
CT-93	93 (25W)	DC-2.5GHz	13.00	15.00	-	15.00	-	15.50	-	-
Mismatched Terminations, 1:05:1 to 3:1, Open Circuit, Short Circuit:										
MT-51	50	DC-3.0GHz	45.50	45.50	45.50	-	-	-	-	-
MT-75	75	DC-1.0GHz	-	-	45.50	-	-	-	-	-
Feed thru Terminations, shunt resistor:										
FT-50	50	DC-1.0GHz	17.50	26.00	19.50	17.50	-	-	-	-
FT-75	75	DC-500MHz	17.50	26.00	45.50	17.50	-	-	-	-
FT-93	93	DC-150MHz	17.50	26.00	45.50	17.50	-	-	-	-
Directional Coupler, 30dB:										
DC-500	50	250-500MHz	60.00	-	84.00	84.00	-	-	-	-
Resistive Decoupler, series resistor or Capacitive Coupler, series capacitor:										
RD or CC-1000	1000 (1000PF)	DC-1.5GHz	17.50	26.00	19.50	17.50	-	-	-	-
Adapters:										
CA-50 (N to SMA)	50	DC-4.2GHz	17.50	26.00	19.50	17.50	-	-	-	-
Inductive Decouplers, series inductor, Bias T:										
LD-R15	0.1uH	DC-500MHz	17.50	26.00	19.50	17.50	-	-	-	-
LD-6R8	6.8uH	DC-55MHz	17.50	26.00	19.50	17.50	-	-	-	-
BT-50	1.8uH	15-500MHz	84.00	84.00	94.00	84.00	-	-	-	-
Fixed Attenuator Sets, 3, 6, 10 and 20 dB, in plastic case:										
AT-50-SET (3)	50	DC-1.5GHz	76.00	120.00	92.00	84.00	-	-	-	-
AT-51-SET	50	DC-1.5GHz	64.00	108.00	82.00	74.00	-	-	-	-
Reactive Multicouplers, 2 and 4 output ports:										
TC-125-2	50	1.5-125MHz	84.00	-	94.00	84.00	-	-	-	-
TC-125-4	50	1.5-125MHz	94.00	-	104.00	94.00	-	-	-	-
Resistive Power Dividers, 3, 4 and 9 ports:										
RC-3-50	50	DC-2.0GHz	84.00	84.00	94.00	84.00	-	-	-	-
RC-4-50	50	DC-500MHz	84.00	84.00	94.00	84.00	-	-	-	-
RC-9-50	50	DC-500MHz	-	-	-	104.00	-	-	-	-
RC-3-75, 4-75	75	DC-500MHz	84.00	84.00	-	84.00	-	-	-	-
Double Balanced Mixers:										
DBM-1000	50	5-1000MHz	61.00	-	71.00	61.00	-	-	-	34.00
DBM-500PC	50	2-500MHz	-	-	-	-	-	-	-	34.00
RF Fuse, 1/8 Amp., and 1/16 Amp.:										
FL-50	50	DC-1.5GHz	17.50	26.00	45.50	17.50	-	-	-	-
FL-75	75	DC-1.5GHz	17.50	26.00	-	17.50	-	-	-	-

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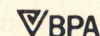
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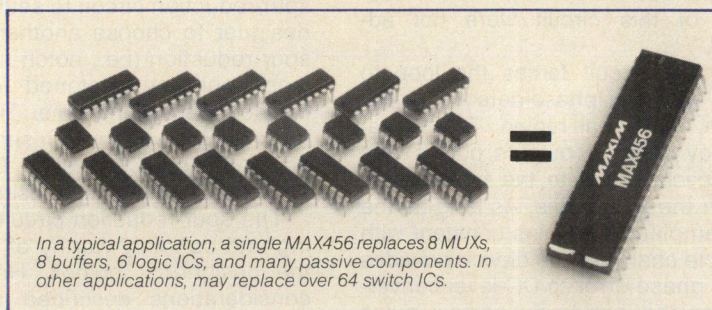


# 8×8 VIDEO CROSSPOINT SWITCH WITH BUFFERS -ONLY \$2.50\*/CHANNEL

## Connects Any Input to Any Output

Maxim's new **MAX456** is the **first** monolithic 8×8 video crosspoint switch that routes standard video signals (NTSC, PAL, SECAM). With a digitally controlled 8×8 switch matrix, control logic, and eight 35MHz output buffers together in a 40-pin DIP or 44-pin PLCC, the MAX456 significantly reduces component count, board space and cost over discrete designs. Applications include video surveillance, imaging, visual automation, and video editing.

## MAX456 Eliminates Over 20 Components

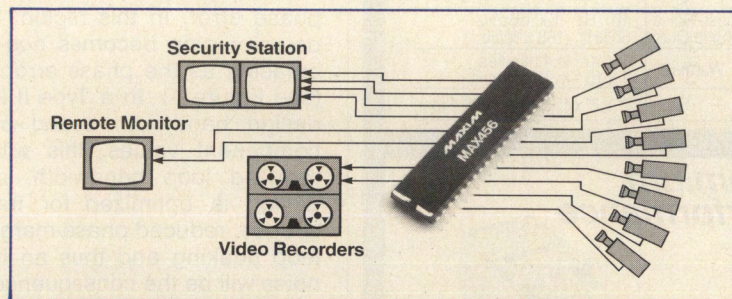


- Reduces Board Space up to 5X
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- Improves Reliability

## Build Larger Crosspoint Arrays

Each MAX456 buffer output can be disabled under logic control. With three-state outputs, multiple MAX456s can be paralleled to form larger switch networks.

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250V/ $\mu$ s Slew Rate  
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Buffer Disable Saves Power
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\* MAX456CPL, \$19.98 1000-up  
F.O.B. U.S.A. price.

# MAXIM

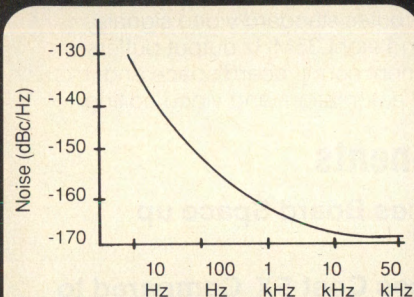
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0°C/+50°C:	±1 x 10 <sup>-8</sup>
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Aging:	5 x 10 <sup>-10</sup> /day, 1 x 10 <sup>-7</sup> /yr.
Phase Noise (Sine Out):	10 Hz: -130 dBc/Hz 50 kHz: -168 dBc/Hz
Warm-up:	1 x 10 <sup>-7</sup> in 4 minutes 1 x 10 <sup>-8</sup> in 10 minutes
Size:	2" x 2" x 1" (51 x 51 x 25.4 mm)

### Premium Performance



CO-705SL2

Frequency:	4-25 MHz
Output:	Sine or Logic
0°C/+50°C:	±1 x 10 <sup>-9</sup>
-55°C/+75°C:	±5 x 10 <sup>-9</sup>
Aging:	5 x 10 <sup>-10</sup> /day, 1 x 10 <sup>-7</sup> /yr.
Phase Noise (Sine Out):	10 Hz: -130 dBc/Hz 50 kHz: -168 dBc/Hz
Warm-up:	1 x 10 <sup>-7</sup> in 3 minutes 1 x 10 <sup>-8</sup> in 5 minutes
Size:	1 1/2" x 1 1/2" x 2" (38 x 38 x 51 mm)



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## RF letters

Letters should be addressed to Editor, *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111.

### Reference Spur Reduction in PLLS

The article, "A Feedback Method for Reference Spur Reduction in PLLs," by John W. MacConnell and Dr. Richard W.D. Booth in the September 1990 issue of *RF Design* presents an interesting technique to be applied to phase-locked loop design. However, some significant consequences which may limit the usefulness of this circuit were not addressed.

First, the circuit forces the loop to operate with the phase-detector in the low-gain, non-linear region. The phase-frequency detector outputs pulses with a duty cycle equal to the phase error between the two inputs. As long as the output amplitude remains constant with duty cycle changes, the device provides a linear phase-error to DC-level conversion. For the extremely narrow pulse widths present for low-phase error inputs, internal propagation delay and especially finite slew rates result in a triangular-shaped output pulse, with both the amplitude and width varying with phase error. In this region, the phase-detector gain becomes non-linear, decreasing as the phase error decreases (see Figure 1). In a Type II loop with a design bandwidth based on nominal component values, this will result in reduced loop bandwidth, and if the design is optimized for transient response, reduced phase-margin. Closed-loop peaking and thus an increase in noise will be the consequence.

Furthermore, one of the implied advantages of the circuit is that it allows

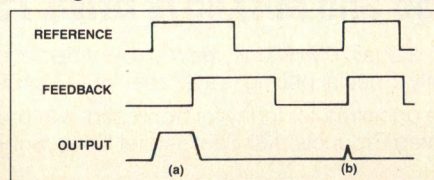


Figure 1. Slew rate resulting in a triangular-shaped output pulse.

In the October 1990 issue, another mistake was brought to our attention in Tom Cefalo's article, "Microstrip CAD Program." There were some parentheses omitted from equation 9c, on page 38 and the equation should read as follows:

$$a\left(\frac{W}{H}\right) = 1 + \frac{1}{49} \ln \left[ \frac{\left(\frac{W}{H}\right)^2 \left( \left(\frac{W}{H}\right)^2 + \left(\frac{1}{52}\right)^2 \right)}{\left(\frac{W}{H}\right)^4 + 0.432} \right] + \frac{1}{18.7} \ln \left[ 1 + \left(\frac{W}{18.1H}\right)^3 \right] \quad (9c)$$

In Figure 7 on page 42, the inductors should be 3.978 nH and not mH.

the more rapid loop settling times associated with wider loop bandwidths. However, this is only partially true. In order for the spur-reduction circuit to not degrade loop stability, its time constant and thus its settling time must be significantly larger than that of the primary loop (I would expect a minimum of ten times larger).

Consider the implications: If a designer needs a loop which must settle in one millisecond, with a 60 dBc spurious requirement, his/her design must provide a loop settling time of less than 100 microseconds to allow the spur-reduction circuit to settle. Were the designer to choose another method of spur reduction (i.e., notch filtering), the loop could be designed for the one millisecond settling time. The resulting narrower loop bandwidth should provide additional spur reduction (at least 20 dB with suitable circuit design).

The spur-reduction circuit is interesting and certainly worthy of inclusion in the designer's toolbox. However, the considerations described above may limit its utility to designs where gain variation and long settling times can be accommodated.

Jonathan McLin  
Electrical Engineer  
Motorola, Inc.

### Corrections

In our 1990/91 *RF Design* Directory Issue there were two mistakes in the design guide articles. The first appears in Richard Bain's article, "Noise Bandwidth Calculation," on page 87. Equation 2 should read:

$$B_n = \frac{1}{P_{ref}} \sum_{k=0}^{k=n} (f_{k+1} - f_k) \left( \frac{P_{k+1} + P_k}{2} \right) \quad (2)$$

Then in W.G. Beauregard's article, "Phase Difference Networks," on page 91 there is a line missing at the end of the page. The sentence should read "An unbalanced bridged tee network can be derived from the lattice network by an application of Bartlett's bisection theorem."



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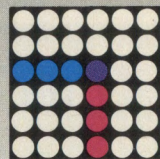
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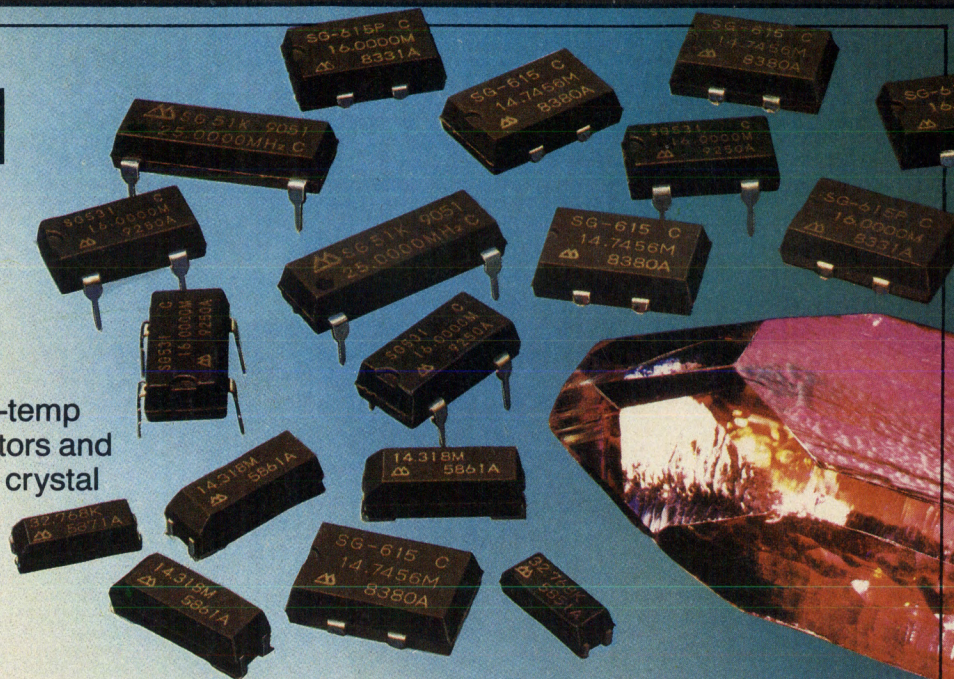


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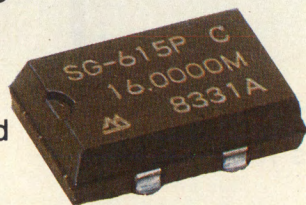
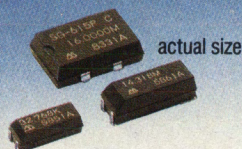
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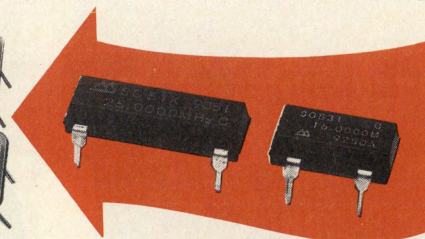
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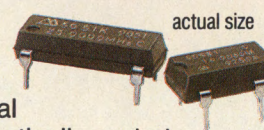
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# RF calendar

## December

**5-7 Ultrasonics Symposium**  
Honolulu Hilton, Honolulu, HI  
Information: LRW Associates, 1218 Balfour Drive, Arnold, MD 21012. Tel: (301) 647-1591.

**9-12 1990 IEEE International Electron Devices Meeting**  
San Francisco Hilton, San Francisco, CA  
Information: Melissa Widerkehr, IEDM, Suite 300, 655 15th Street, NW, Washington, DC 20005. Tel: (202) 347-5900. Fax: (202) 347-6109.

## January

**14-16 4th Annual International Superconductor Applications Convention**  
San Diego, CA  
Information: SCAA, 27692 Deputy Circle, Laguna Hills, CA 92653. Tel: (800) 854-8263 or (714) 362-9701. Fax: (714) 362-9803.

**15-17 ATE & Instrumentation West**  
Disneyland Hotel, Anaheim, CA  
Information: Tel: (800) 223-7126 or (617) 232-3976.

**22-24 Hyper 91, Microwave Technology Exhibition and Congress**  
Palais des Congres, Paris, France  
Information: B.I.R.P., 25 rue d'Astorg, 75008 Paris, France. Tel: 33-(1)-4742-2021. Fax: 33-(1)-4742-7568.

**28-31 Communications Networks '91**  
Washington Convention Center, Washington, DC  
Information: Michael Sullivan. Tel: (508) 820-8268.

## February

**5-7 RF Expo West 91**  
Santa Clara Convention Center, Santa Clara, CA  
Information: Kristin Hohn, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

**12-14 4th International Smart Card Exhibition and Conference**  
Novotel, Hammersmith, London  
Information: Elisabeth Beckett, Marketing Manager, Agestream Ltd., Towermead Business Center, High Street, Old Fletton, Petersborough, U.K. PE2 9DY. Tel: (0733) 60535. Fax: (0733) 45522.

**24-28 NEPCON West '91**  
Anaheim Convention Center, Anaheim, CA  
Information: Michele Filippi, Cahners Exposition Group, 1350 E. Touhy Ave., Des Plaines, IL 60017-5060. Tel: (708) 299-9311.

## March

**26-28 International Mobile Communications Expo**  
Anaheim Convention Center, Anaheim, CA  
Information: April Debaker, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

## Straight Answers to Tough Questions about the TESS Block Diagram Simulator

### How is the TESS simulator different?

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### How easy is it to run a simulation after drawing a block diagram in OrCAD?

You can run OrCAD/SDT under TESS 1.1. OrCAD's NETLIST utility produces the netlist which TESS simulates directly.

**Does TESS have models for commercial devices?** TESS blocks are meant to be general instead of representing specific devices. Parameters for each block are entered in the block diagram.

**What if TESS doesn't have the model I need?** The models are a basis set. Build up what you need with a few blocks. Then put the subcircuit in a library to use like a new model. The MODGEN option lets you add code to make new models.

### Can I specify NF, intercept, BW and flatness for an RF amplifier?

Essentially. The new RF amp model has intercept and sat parameters. Sum-in Gaussian noise to set NF. Add a bandpass filter to set the bandwidth and ripple.

**Is there a frequency limit?** Not really. The number of simulation points is the real issue ( $\propto \text{freq} \times \text{time}$ ). Settling time tends to decrease as freq increases. HF systems with slow sections can be shifted in freq to reduce the number of points.

**Can I measure stability, BW, noise and do Bode plots?** Use the built-in instruments and spectrum analysis to do testing just like in the lab. The sweep function generator, rms meter and  $\phi$ -det let you check bw, do Bode plots and excite loops.

### Can I enter an arbitrary freq response?

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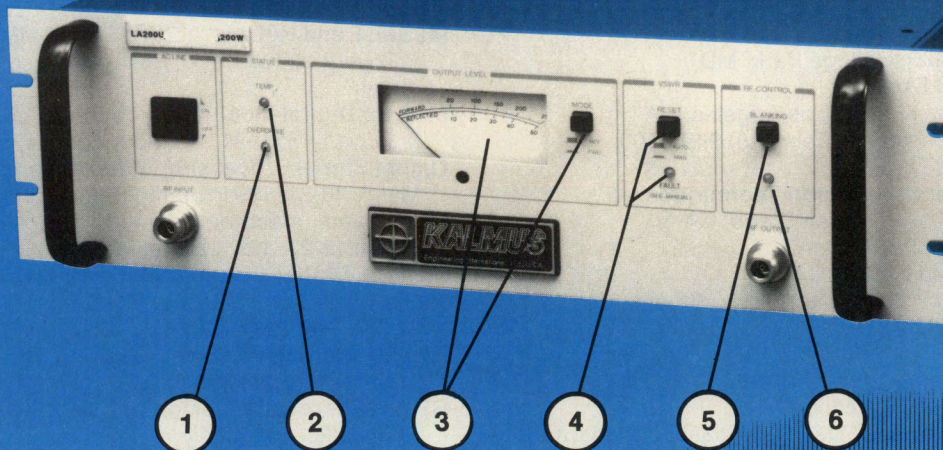
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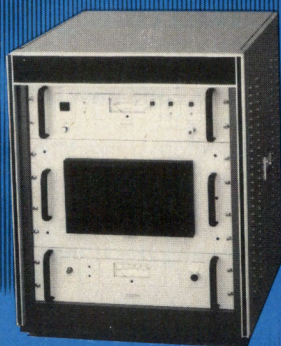
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## RF/Microwave Circuit Design II

December 10-14, 1990, College Park, MD

## RF/Microwave Circuit Design

January 28-February 1, 1991, Los Angeles, CA

Information: Besser Associates. Tel: (415) 949-3300.

## Modern Power Conversion Design Techniques

February 25-March 1, 1991, San Diego, CA

April 29-May 3, 1991, Phoenix, AZ

May 20-21, 1991, San Rafael, CA

Information: e/j Bloom Associates, Joy Bloom. Tel: (415) 492-8443. Fax: (415) 492-1239.

## Digital Signal Processing Workshop

March 12-14, 1991, Campbell, CA

Information: Analog Devices, DSP Applications Department, Maria Butler. Tel: (617) 461-3672.

## Modern Microwave Techniques

February 25-March 1, 1991, Garmisch-Partenkirchen, Germany

## Far-Field, Compact and Near-Field Antenna Measurement Techniques

February 25-March 1, 1991, Garmisch-Partenkirchen, Germany

## Aspects of Modern Radar

February 25-March 1, 1991, Garmisch-Partenkirchen, Germany

## MESFET and Hetrostructure Based MMICs

February 25-March 1, 1991, Garmisch-Partenkirchen, Germany

## Modern Digital Modulation Techniques

March 11-15, 1991, United Kingdom

## Digital Signal Processing: Filtering and Estimation

March 18-21, 1991, United Kingdom

## Broadband Telecommunications

March 18-22, 1991, United Kingdom

## Modern Digital Communications for Space, Satellite and Radio

April 15-18, 1991, Italy

## RF and Microwave Circuit Design I: Linear Circuits

April 15-19, 1991, Italy

## RF and Microwave Design II: Non-Linear Circuits

April 22-26, 1991, Italy

Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: 46 (0) 122-17570. Fax: 46 (0) 122-14347.

## New HF Communications Technology: Advanced Technology

December 17-21, 1990, Washington, DC

## Introduction to Radar ECM and ECCM Systems

December 17-21, 1990, Washington, DC

February 20-22, 1991, Washington, DC

## Communication and Radar Signals: Detection, Estimation & Geolocation Techniques

January 9-11, 1991, Washington, DC

## Hazardous Radio-Frequency Electromagnetic Radiation: Evaluation, Control, Effects, and Standards

January 16-18, 1991, Washington, DC

## Mobile Cellular Telecommunications Systems

January 16-18, 1991, Washington, DC

## Fiber-Optics System Design

February 4-6, 1991, Washington, DC

## Cellular Radio Telephone Systems

February 25-27, 1991, San Diego, CA

## Principles of Digital Cellular Telephony

February 25-March 1, 1991, Washington, DC

## Microwave High-Power Tubes and Transmitters

February 25-March 1, 1991, Washington, DC

## Broadband Communication Systems

March 4-8, 1991, Washington, DC

## Satellite Communications: System Planning, Design, and Operation at Ku and Ka Bands

March 4-8, 1991, Washington, DC

## Radar Operation and Design: The Fundamentals

March 25-28, 1991, Washington, DC

Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

## Basic Network Measurements Using the HP8510B Network Analyzer

December 17-19, 1990, Boston, MA

January 14-16, 1991, Los Angeles, CA

## Microwave Fundamentals

January 8-11, 1991, Los Angeles, CA

## HP11776A Waveform Generation Language User Course

January 10-11, 1991, Los Angeles, CA

## Programming the HP 8510 Network Analyzer

January 17-18, 1991, Los Angeles, CA

Information: Hewlett-Packard Company. Tel: (800) 472-5277.

## TRIMMER CAPACITOR PROTOTYPING KITS



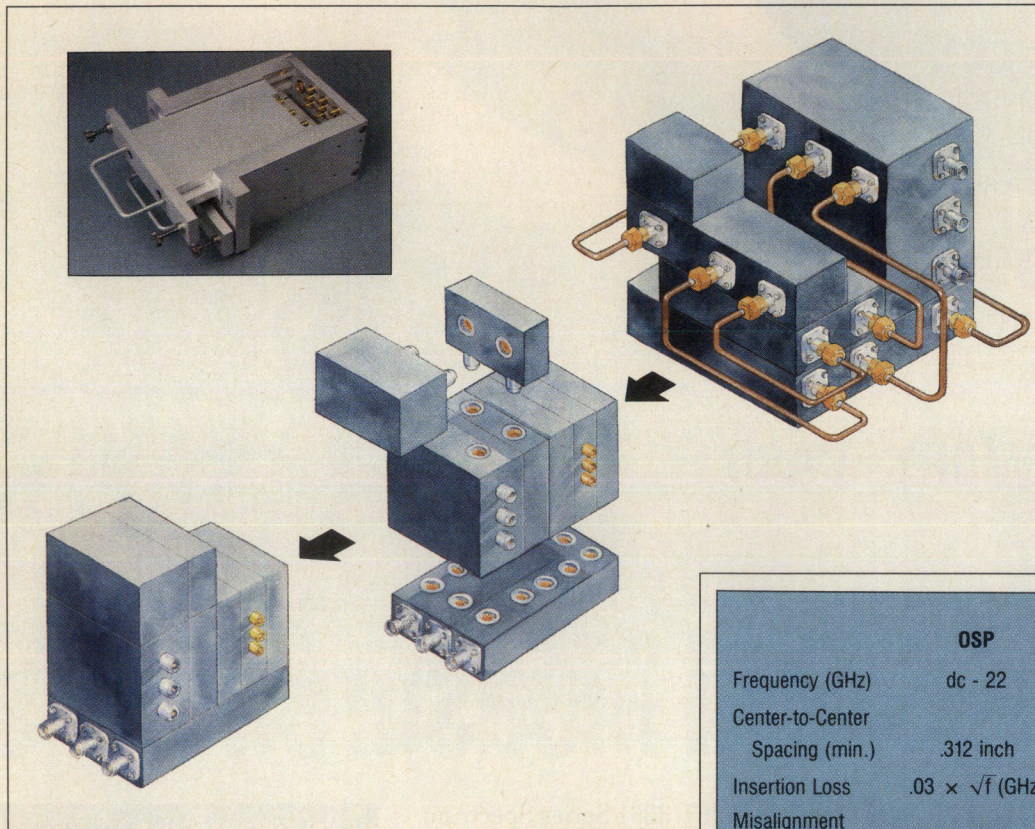
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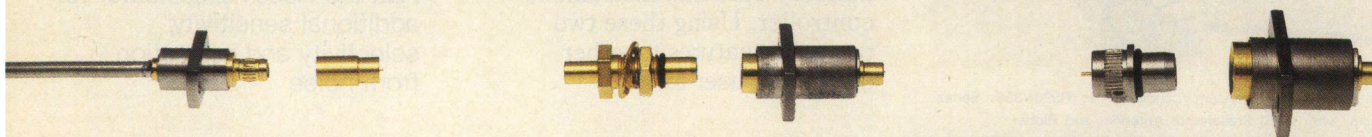
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Misalignment			
Rigid Mount	±.004 inch	±.0025 inch	±.0015 inch
Float Mount	±.020 inch	±.0200 inch	±.0065 inch
Durability	1,000 cycles	1,000 cycles	500 cycles

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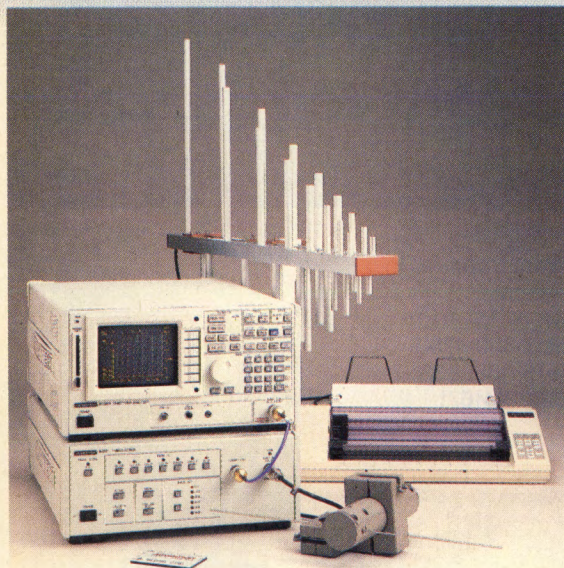


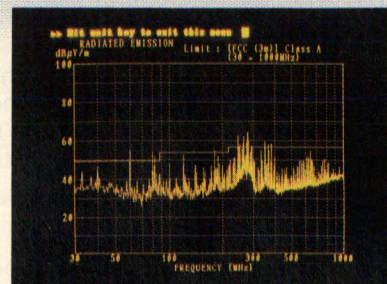
Photo: R2542B EMI Receiver System, connected to R3261/3361 Series Spectrum Analyzer with R3551 Preselector Antenna, and Plotter.

The R3261/3361 Series Spectrum Analyzers are ideally suited for RF signal analysis and offer superior functions and capabilities for EMI measurements.

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## RF Expo East: The Best Yet

The most successful RF Expo East ever was held in Orlando, Florida at the Marriott Orlando World Center. One hundred twelve exhibiting companies, five full day special courses, and 53 technical sessions at the fifth annual Expo East combined to produce an outstanding show. Attendance was up this year with approximately 870 registered attendees and 523 exhibit personnel for a total attendance of 1393, an increase of 19.1 percent over last year's Expo East. The increase can be attributed to several factors: an excellent technical program, good exhibiting companies, location, and promotion of the show at some local companies.



*Steve Russell explains the finer points of DSP Demodulation.*

Dr. Frederick H. Raab, this year's technical session chairman, put together a well rounded program. For the first time, a fourth track was added to accommodate the large number of submittals for this year's Expo. The four receiver sessions drew large crowds that left standing room only. The three tutorials as always, drew a large crowd. The second session on PLLs and Synthesizers, drew another large crowd for its papers. "Designing with Direct Digital Frequency Synthesis," "Direct-digital Waveform Generation using Advanced Multi-mode Digital Modulation," and "Optimum PLL design for Low Phase-noise Performance" generated some good questions. Of interest, was the session on RF Systems for Research in Particle Physics. The papers came from three different research facilities — two in the United States and one in Germany. In all cases, the papers were well attended, and response from the attendees was positive.

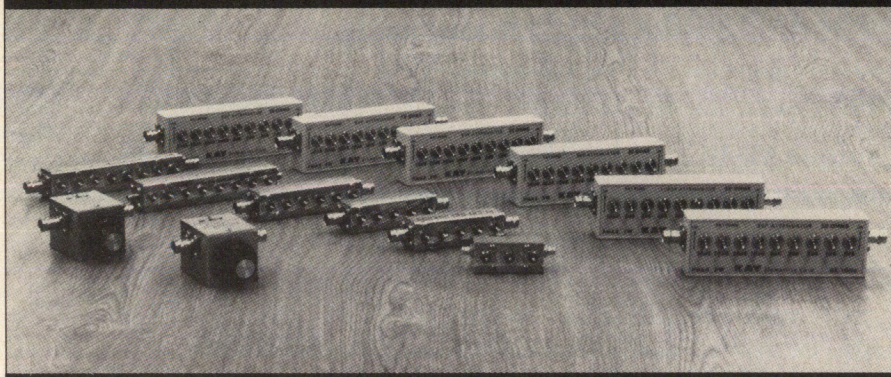
The design tutorials covered RF circuit design, computer-aided filter design, oscillator design and CAD techniques.

There were a number of new products being exhibited at the show. TTE featured their new line of miniature duplexers that cover the frequency range of 20 kHz to 100 MHz. Phillips Components also received excellent response

to their cellular chip set Mark II demo board.

RF Expo West will be held February 5-7, 1991, at the Santa Clara Convention Center in California. Next year's Expo East will be held in Florida again at the Stouffer Orlando Resort from October 29-31. Come join us for the best that the RF industry has to offer in technical presentations and expositions.

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<b>847</b>	75Ω	DC-1GHz	0-102.5dB	1dB
<b>870</b>	75Ω	DC-1GHz	0-132dB	1dB
<b>4440</b>	50Ω	DC-1.5GHz	0-130dB	10dB
<b>4450</b>	50Ω	DC-1.5GHz	0-127dB	1dB
<b>1/4450</b>	50Ω	DC-1GHz	0-16.5dB	.1dB
<b>4467</b>	75Ω	DC-1GHz	0-31dB	1dB
<b>0/400</b>	50Ω	DC-500MHz	1-13dB	—
<b>0/410</b>	75Ω	DC-400-MHz	2-14dB	—

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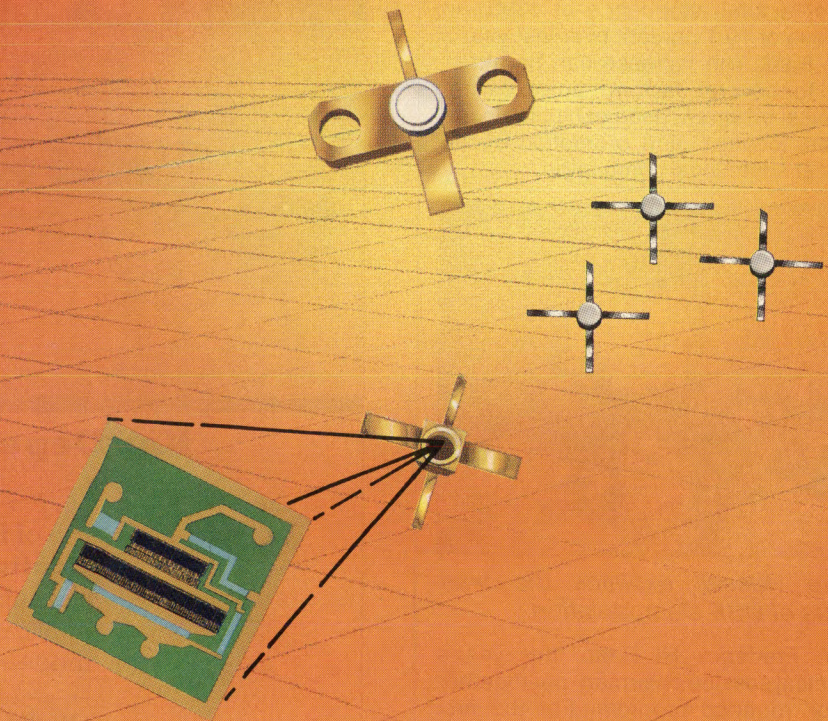
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Model #	Freq. GHz	Gain (dB)	Gain* (dB)	Pout (dBm)	N.F. (dB)	Vd/Id (v/ma)
---------	--------------	--------------	---------------	---------------	--------------	-----------------

AMP3020	0.5	8.0	8.5	29.5	8.0	17/300
	0.2	8.2	8.5	31.0	7.0	17/300

AMP0135	1.0	19.0	20.0	3.0	4.5	5/17
AMP0235	1.0	12.5	13.0	5.0	5.5	5/25
AMP0335	1.0	11.5	12.5	10.0	6.0	5/35
AMP0435	1.0	8.5	9.1	11.2	6.0	5.3/50
AMP0420	1.0	10.0	11.5	14.0	6.5	6.3/90
AMP0520	1.0	9.7	9.2	23.0	6.5	12/165
AMP0635	1.0	19.0	20.0	4.5	3.0	3.5/16
AMP0735	1.0	13.0	13.7	5.5	4.5	4/22
AMP0835	1.0	19.0	31.0	14.0	3.0	7.8/36
AMP0910	1.0	8.7	8.0	11.3	5.5	7.8/35
AMP1025♦	1.0	6.3	7.6	27.0	9.0	15/325
AMP1120	1.0	11.5	12.2	17.5	3.5	5.5/60

\*Gain @ 0.1 GHz

♦Measured in a 50Ω System

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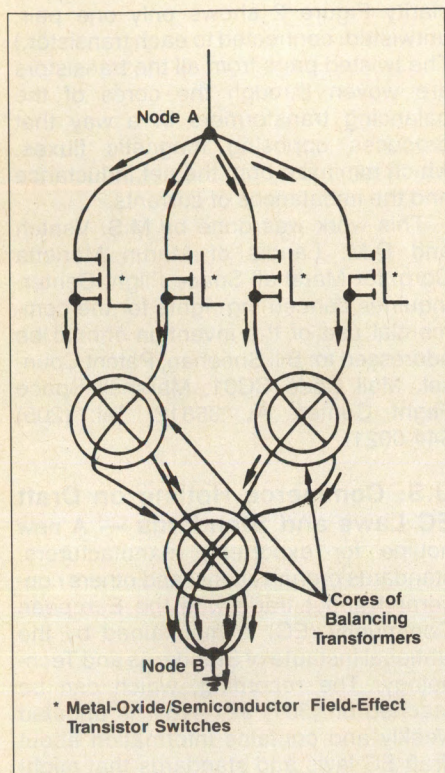


Figure 1. In the old wiring configuration, currents are balanced on single wires, with return current running elsewhere. This configuration has relatively high inductance.

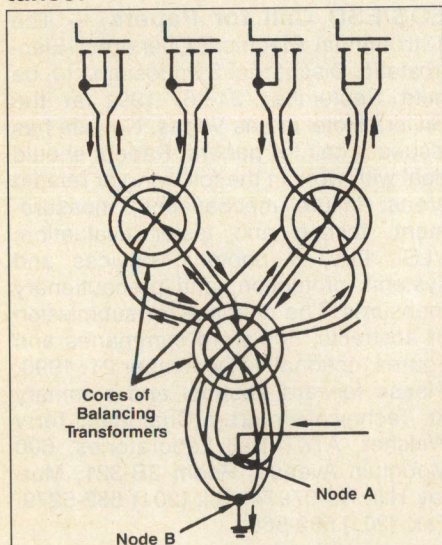


Figure 2. In the new wiring configuration, currents are balanced on twisted pairs of wires. Because the twisted pairs carry both the "hot-side" and return currents, this configuration has relatively low inductance.

**Low-Inductance Wiring for Parallel Switching Transistors** — This note originally appeared in the *NASA Tech Briefs* October 1990 issue. It describes transistors that share current equally, without sacrifice of switching speed.

A simple configuration for the wiring of multiple parallel-connected switching transistors minimizes the stray wiring

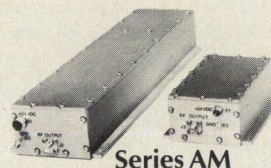
inductance while providing for the use of balancing transformers, which equalize the currents in these transistors. The balancing of currents is necessary to prevent overloads in individual transistors, and the minimization of inductance is essential for fast switching of high currents. High-current transistor switches that could benefit from the new configuration are found in controllers for

## CLASS A LINEAR POWER

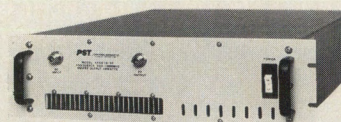
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Interleaving and multifilar windings can reduce the leakage inductances of transformers to acceptable levels. Previously, however, the use of balancing transformers entailed excessive distrib-

uted wiring inductances because the connections to the transistors were made with relatively-high-inductance single wires (see Figure 1).

The new configuration (see Figure 2) is based on the established technique of reducing inductance by laying wires in twisted pairs. The source and drain leads of each transistor are connected to six twisted pairs of wires. (To enhance

clarity Figure 2 shows only one pair, untwisted, connected to each transistor.) The twisted pairs from all the transistors are woven through the cores of the balancing transformers in a way that produces opposing magnetic fluxes, which minimize both the net inductance and the imbalances of currents.

This work was done by M.S. Veatch and D.M. Landis of Martin Marietta Corp. for Marshall Space Flight Center. Inquiries concerning rights for the commercial use of the invention should be addressed to: Bill Sheehan, Patent Counsel, Mail Code CC01, Marshall Space Flight Center, AL 35812. Tel: (205) 544-0021.

#### **U.S. Commerce Hotline on Draft EC Laws and Standards**

— A new hotline for exporters, manufacturers, standards organizations, and others concerned about trade with the European Community (EC) is maintained by the National Institute of Standards and Technology. The recording, which can be reached on (301) 921-4164, is updated weekly and contains information about draft EC laws and standards that might create technical trade barriers. Hotline topics are listed by subject area and product. Information is provided on deadlines for comments and a point of contact for obtaining a review copy of the text.

#### **EOS/ESD Call for Papers**

— The 13th annual Electrical Overstress/Electrostatic Discharge Symposium, to be held September 24-26, 1991 at the Riviera Hotel in Las Vegas, Nevada has issued a call for papers. Papers should deal with work in the following or related areas: failure mechanisms; measurement, testing and tester evaluation; VLSI, III-V, & photonic devices and systems protection; and precautionary measures. The deadline for submission of abstracts, 500 word summaries and figures (optional) is December 21, 1990. Please forward abstract and summary to: Technical Program Chairman, Terry Welsher, AT&T Bell Laboratories, 600 Mountain Avenue, Room 3B-321, Murray Hill, NJ 07974. Tel: (201) 582-5279. Fax: (201) 582-5661.

#### **Session Proposals Sought for Frequency Control Symposium**

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ory and design of piezoelectric resonators, resonator processing techniques, filters, SAW devices, quartz crystal oscillators, sensors and transducers, frequency and time coordination and distribution, and applications of frequency control. Two copies of a summary (500 words) together with the author's name, address and telephone number should be sent to: Dr. Thomas E. Parker, Raytheon Research Division, 131 Spring Street, Lexington, MA 02173. The deadline for submission of summaries is January 14, 1991. The symposium will be held May 29-31, 1991 at the Los Angeles Airport Marriott, Los Angeles, CA.

### Ball Corporation Demonstrates Satellite Communication System —

Ball Corporation's Airlink™ conformal, aeronautical phased-array antenna was recently demonstrated aboard a Japan Air Lines aircraft. The external portion of the antenna system consists of two conformal antenna arrays, each less than 0.3 inches thick, mounted on opposite sides of the aircraft fuselage. During the inaugural flight from Tokyo to San Francisco, all passengers were provided phone and fax services to anywhere in the world by using the system. Upon successful completion of the Phase II demonstration project, JAL expects to place this service into normal operation aboard this aircraft.

### Valpey-Opt Company Formed —

Valpey-Fisher Corporation and OPT Industries have joined together to form a company that will offer a wide range of quality crystal and L/C filters. Using CAD, analysis and production techniques, Valpey-Opt will apply standard and non-standard design to customers' requirements in the time and frequency domain. These designs include Butterworth, Chebyshev, Elliptic and linear phase shift filters as well as customized applications demanding phase and amplitude equalization, single-side band rejection or wide-band reject filters. Inquiries are invited. Phone Wim van den Akker, Chief Engineer, tel: (508) 435-6831, ext. 212.

### New Divisions Created at Crystal Technology —

Crystal Technology, Inc., has been divided into two new divisions, one focusing on crystal growing and wafer fabrication, the other on components utilizing these materials. The Materials Division will supply lithium niobate crystals as well as single crystal

material, such as non-linear optical crystals for laser frequency doubling. The Components Division will be responsible for the company's integrated optic, electro-optic, acousto-optic and SAW product lines.

### Phoenix Microwave Labs Formed

— Phoenix Microwave Labs is a new company formed to provide design,

development and manufacture of microwave energy sources. The company's initial product offerings include dielectric resonator oscillators, phase locked DROs, solid state power amplifiers, custom up and down converters and transmitter subsystems. The company is located at 360 Scarlet Blvd., Oldsmar, FL 34677. Tel: (813) 854-4822. Fax: (813) 854-2020.



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OUTPUT WAVEFORM	CMOS, TTL, SINE	y=C CMOS y=S SINE	SQUARE WAVE
OUTPUT LEVEL	Waveform Compatible	Waveform Compatible	TTL, HCMOS, LSTTL (V <sub>OH</sub> =V <sub>CC</sub> –0.5V) (V <sub>OL</sub> =0.5V max)
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AGING	±1 ppm/year	±1 ppm/year	±0.5 ppm/year
FREQUENCY ADJUSTMENT	External Trimming Option	±10 ppm (Internal trimmer) (Opt. Voltage Adjustment)	±5 ppm (By mechanical means)
CASE STYLE	Fig. 1	Fig. 2	Fig. 2

Fig. 1

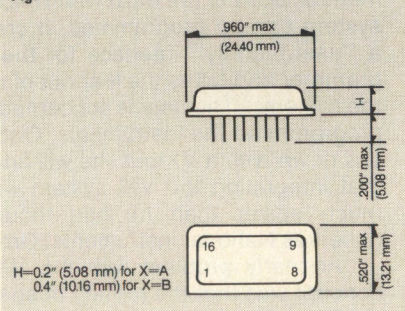
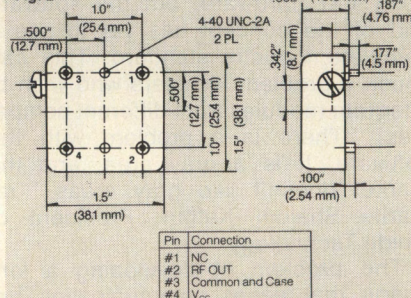


Fig. 2





# VXIbus: A Standard for the Modular Instruments Market

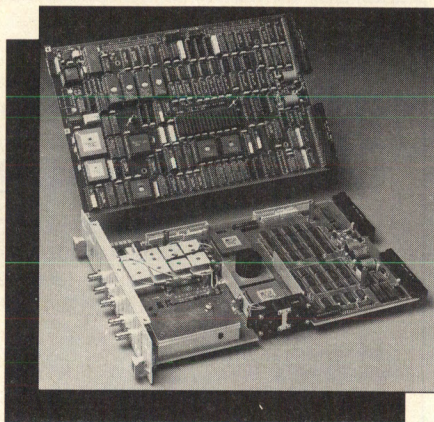
By Charles Howshar and Liane Pomfret,  
Assistant Editors

Of all the developments in modular RF instruments, probably none has had as much impact as VXIbus. With a VXI-based system, engineers can develop custom test systems for far less money than was feasible only a few years ago. That is not to say that there have not been problems along the way. Initial pricing was too high, integrated software was lacking, and hardware integration had few guidelines. Now companies are expanding their marketing efforts to include all aspects of the test and measurement market. As Malcolm Levy, Business Manager for Racal-Dana remarks, "We're making a big push to demonstrate VXI as an RF and microwave testing system."

There are many companies who are excited about the potential of VXIbus for modular instruments. And, although it is still a young market, some companies have put a great deal of their time and money into the VXIbus design. Levy indicates, "For us VXI is very large. It makes up fifty percent of our business today, and eighty percent of our engineering effort is wrapped up in it." This support seems to be well-placed as interest in VXI is growing. "The market last year was about \$20 million; this year it will be about \$50 million," agrees John Graff, VXI Product Manager for National Instruments.

Standards have been set for VXI hardware, and using this framework, companies have developed some very good VXI hardware, but the software end of the concept has yet to see a complete industry standard. This can cause a problem for users who wish to integrate modules from different companies. "There is a problem with the software. It is a gray area, and the market doesn't like gray areas," remarks Shalom Kattan, President of Guide Technology.

The problem of developing a language that eases communication between modular instruments regardless of the maker is slowly being solved. The Test and Measurement System Lan-



guage (TMSL) from Hewlett-Packard was developed as an integrated language for use in any VXI system. Many companies agree that, although a good language for test and measurement, TMSL, like other VXIbus languages, has its limitations. For one, it only supports proprietary instruments. So in order to employ the software, a user first has to have the hardware to run it on. From TMSL and other languages the Standard Commands for Programmable Instruments (SCPI) has been developed. The introduction of SCPI was a significant step. "With its acceptance, the bottleneck that developed because of the software problems was removed. It's been a godsend to VXI users," comments Roger Muller, Marketing Section Manager for Hewlett-Packard.

An added benefit from the development of SCPI is the ease with which the system can be programmed. It creates a "user-friendly" interface for the programmer and takes the fear out of using the system. "The user is concerned with programming the instruments. Once he sits down with a VXIbus, he will find out that integrating the VXI system will be much easier than he had thought," observes National Instruments' Graff.

An early problem with the VXIbus concept was that the military pushed for size and weight concerns and ignored price and practicality. When it became available to the public, they found it too

expensive. Since then, prices have become more realistic and industry has become an integral part of VXI development. "Military was the early backer but they have scaled back because of budget cuts. General industry has picked up the load," comments Graff. Now improvements in the VXIbus design are appearing on the market. One idea that has just been developed by Rapid Systems is called the PCXI system. The attraction of this design is that it is less expensive than a VXI design. Although PCXI is not applicable for the higher speeds of VXI system uses, Rapid Systems is developing a system with a burst mode of 33 MHz to fit these applications. "For people who are looking for a modular test system, PCXI does most of the operations of a VXI system at a fraction of the cost," remarks Matthew Arksey, Research and Development Engineer for Rapid Systems.

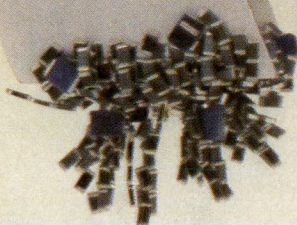
A critical problem for VXI designs, because of the close proximity of many RF circuits is electromagnetic interference (EMI). Most problems with EMI in the VXI design have been successfully dealt with, and improvements are coming onto the market all the time. Donn Mulder, Vice-President of Marketing and Sales for EIP Microwave observes, "Simple attention to grounding and shielding design causes EMI not to be a problem." An interesting observation can be made about EMC concerns for VXI systems: What EMI characteristics will a system have if it incorporates modules from different companies? Mulder notes that the problem is being addressed, "Racal-Dana has come up with a chassis that is more robust and forgiving. It provides a lot of internal shielding."

Test engineers have been waiting for the VXIbus design to become the design standard it was projected to be eighteen months ago. And from the interest it has generated, it seems to be moving into that position. The development of a standard language for VXI will make the test system a mainstay of the modular instruments market in the coming years.

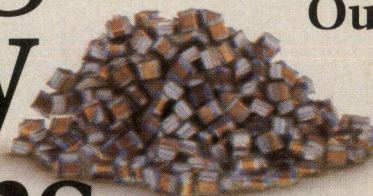


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## Circuits for Wide-band FM Demodulation

By Alvin Wong and Ali Fotowat  
Signetics Company

In applications such as telemetry, high-speed data transmission, and FM over fiber-optics, wide-band discriminators are highly desirable. In such cases, large frequency deviations are needed to use the benefits of wide-band FM. For low distortion demodulation of these signals, higher final IF frequencies such as 21.4 MHz or 70 MHz should be used. These frequencies are generally much harder to work with than lower IFs such as 10.7 MHz, or 455 kHz. The circuits described here use the property that the sinusoidal output of a bandwidth-limited IF limiter can be as effective as a square wave in producing low distortion, wide-band demodulated signals, when the sinusoid is effectively clipped by the demodulator mixer. The complete mathematics of the problem are also derived with normalized design curves. A Fortran 77 program is available for computing demodulator performance.

The circuits presented give stable IF amplifier gains at 21.4 MHz and 70 MHz, and wideband discriminators for 300 kHz baseband signals, with 600 kHz frequency deviation and distortion of the order of 1 percent.

### Theory

FM demodulation can normally be achieved by multiplying two symmetric square waves with a 90 degree phase difference at the carrier frequency and a linear phase relationship around it (see Figure 1). The required phase shift can be obtained from an LC resonator, or a delay line among others. In the case of an FM demodulator based on an LC resonator, the resulting phase difference is:

$$\phi = \tan^{-1} \left[ \frac{\frac{\omega_0}{Q\omega}}{1 - \left(\frac{\omega_0}{\omega}\right)^2} \right] \quad (1)$$

where  $Q = R(C_p + C_s) \omega_0$ , and  $\omega_0 = 1/\sqrt{L(C_p + C_s)}$ .  $C_p$  is the quad tank capacitor;  $C_s$  is the capacitor coupling to the tank, and  $R$  is the total effective loading resistance.

If a delay line is used, the resulting phase difference will be:

$$\phi = \frac{\pi}{2} \left( \frac{\omega}{\omega_0} \right) \quad (2)$$

where the time delay is  $\tau = \pi/2\omega_0$ .

In a phase detecting multiplier, the ideal case is that of two square waves. Therefore, the low pass filtered output's DC level will be linearly proportional to the phase difference. In reality however, the waveforms may not be perfect square waves. For a quad tank, the phase shifted signal will be sinusoidal due to the resonator. In cases where the limiter itself may not have enough bandwidth to pass the odd numbered harmonics of the IF, both signals will be sinusoidal no matter what type of phase shifting technique is used. The use of a multiplier like a Gilbert cell which requires a small signal (50-100 mV peak) to switch hard, will then be equivalent to multiplying clipped sinusoids in a perfect multiplier. The multiplier transfer function will then be:

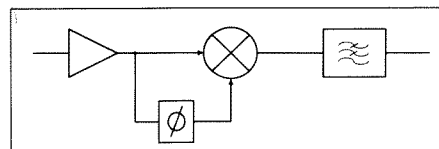


Figure 1. Basic FM demodulator.

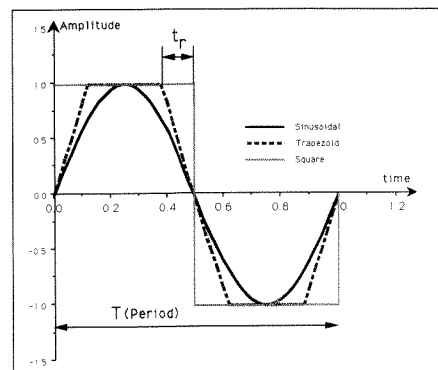


Figure 2. Different wave forms used in the analysis.

$$VDC_{out}(\phi) = \frac{1}{T} \int_0^T Z \left( \frac{2\pi}{T} t \right) Z \left( \frac{2\pi}{T} t - \phi \right) dt \quad (3)$$

By making various assumptions for the shape of the signal  $Z((2\pi/T)t)$ , see Figure 2, this integral can be evaluated as shown below. If a unity peak amplitude square wave is assumed, the output DC transfer function of the discriminator will be:

$$VDC_{out} = 1 - \frac{2\phi}{\pi} \quad (4)$$

Although a sinusoidal wave is the worst case in practice, it is mathematically easy to evaluate. For this example we have assumed an equivalent sinusoid peak amplitude of 1.

$$VDC_{out} = \frac{1}{2} (\cos \phi) \quad (5)$$

More realistic, however, is an effectively clipped sinusoid which is approximated by a trapezoidal wave-form of unity peak amplitude. Then,

$$VDC_{out}(\phi) = 1 - \frac{8r}{3} - \frac{\phi^2}{2\pi^2 r} + \frac{\phi^3}{24\pi^3 r^2} \quad (6)$$

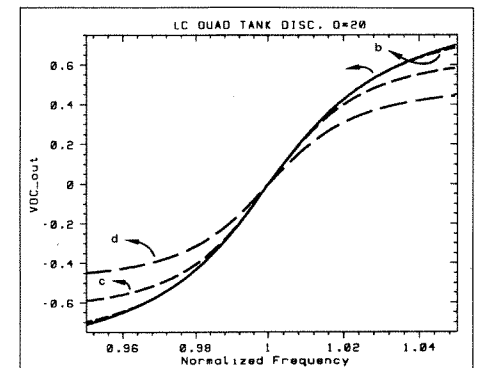
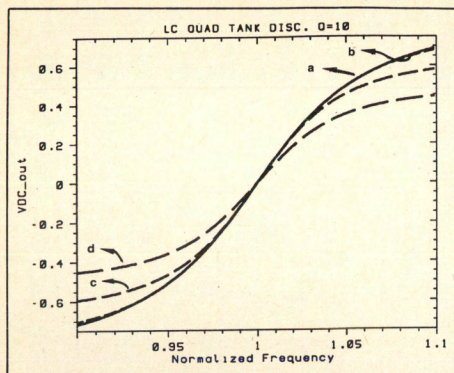


Figure 3. LC quad tank discriminator  $Q=20$ , a) square wave assumption b) trapezoidal wave assumption  $2t_r = 0.125T$  c) Trapezoidal wave assumption  $2t_r = 0.25T$  d) Sinusoidal wave assumption.





**Figure 4. LC quad tank discriminator  $O=10$ , a) square wave assumption b) trapezoidal wave assumption  $2t_r = 0.125T$  c) Trapezoidal wave assumption  $2t_r = 0.25T$  d) Sinusoidal wave assumption.**

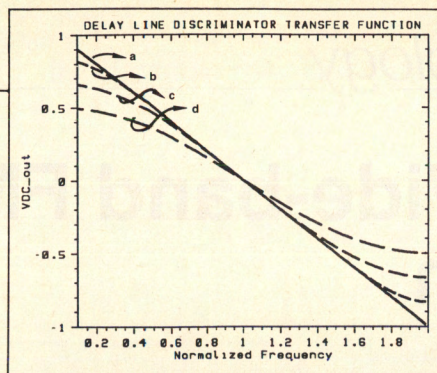
for  $0 < \phi < 4\pi r$

$$VDC_{out}(\phi) = 1 - \frac{2\phi}{\pi} \quad (7)$$

for  $4\pi r < \phi < \pi - 4\pi r$

$$VDC_{out}(\phi) = -VDC_{out}(\pi - \phi) \quad (8)$$

for  $\pi - 4\pi r < \phi < \pi$ , where  $r = t_r/T$  is the ratio of one half of the trapezoidal waveform's rise time over the period as shown in Figure 2, and  $r < 0.125$ .



**Figure 5. Delay line discriminator transfer function, a) square wave assumption b) trapezoidal wave assumption  $2t_r = 0.125T$  c) Trapezoidal wave assumption  $2t_r = 0.25T$  d) Sinusoidal wave assumption.**

By using the phase difference in the above formulas, the FM demodulator transfer function can be obtained as shown in Figures 3-5. Obviously, a delay line discriminator can result in much wider band discriminators as expected. The interesting point, however, is that a trapezoidal wave form with a total normalized rise or fall time of 25 percent will also result in a perfectly linear discriminator curve in the middle of the

transfer function. As the rise and fall times are reduced to 12.5 percent the resulting transfer function approaches that of a perfect square wave.

To get an estimate of distortion, a 3rd order polynomial can be fitted on the part of the transfer function in use. In that way, the 2nd and 3rd harmonic distortion can be estimated.

$$VDC_{out}(f) = a_0 + a_1(f - f_0) + a_2(f - f_0)^2 + a_3(f - f_0)^3 \quad (9)$$

$$f - f_0 = A \cos(\omega_m t) \quad (10)$$

where  $f_0$  is the carrier frequency and  $\omega_m$  is the baseband modulating frequency. The 2nd and 3rd harmonics will then be:

$$H_2 = 20 \log \left[ \frac{2a_2 A}{4a_1 + 3a_3 A^2} \right] \quad (11)$$

$$H_3 = 20 \log \left[ \frac{a_3 A^2}{4a_1 + 3a_3 A^2} \right] \quad (12)$$

### Circuits

Figures 6-9 show the circuits for the

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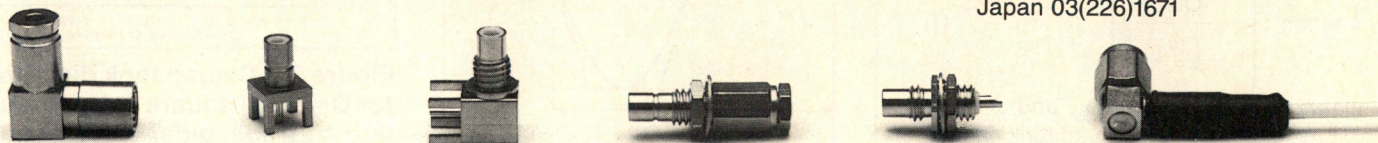
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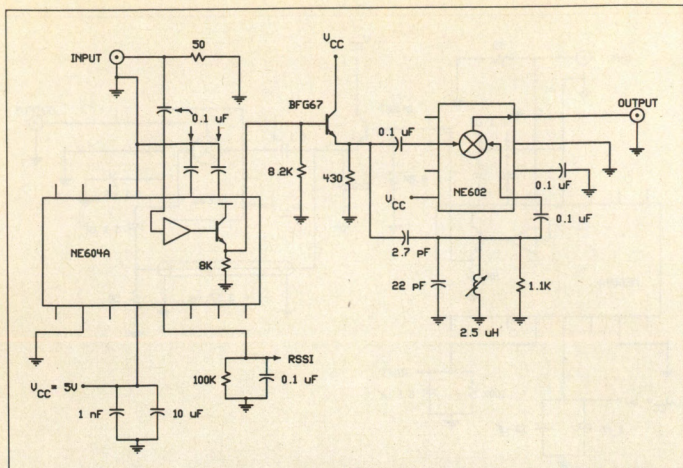


Figure 6a. 21.4 MHz LC discriminator.

quad tank and delay line discriminators at 21.4 MHz and 70 MHz. For the limiter the second IF amplifier of the NE604A chip is used. This part of the chip is a standard FM IF limiter with about 60 dB of low frequency gain, high input impedances and a standard self biasing scheme as shown in the data sheet. The operation of the device with 1500 ohms input impedance at 21.4 MHz, or 70 MHz will obviously result in severe bandwidth limitations as well as instability. Reduction of the limiter's input impedance by an external 50 ohm termination solves these problems. The limiter generates a square wave (at 21.4 MHz) with 200 mV RMS output from an emitter follower with an 8k resistor to ground as shown in the data sheets. For wideband demodulators at 21.4 and 70

MHz, low impedances should be driven at high speeds which this buffer can not do. As a result, a BFG67 Philips RF transistor is added as a second buffer. An NE602 is used as the demodulator multiplier. The high impedances of the NE602 as well as its high speed make it almost an ideal mixer for this circuit. The Gilbert cell mixer switches hard as long as its inputs are higher than 50-100 mV peak as expected. Therefore, one can assume the limiter signals to be similar to the clipped sinusoid shown in the theoretical analysis. For the delay line,

accurately cut lengths of RG-174 cables measured by both TDR and Network Analyzer methods were used. A 50 ohm termination at the end of the delay line is needed for proper operation. The use of 50 ohm delay lines required high currents in the emitter follower. Higher impedance delay lines can also be used to save power.

#### Test Results

Figure 6a shows the 21.4 MHz LC Circuit. A 2.7 pF coupling capacitor is used for coupling to a quad tank with a

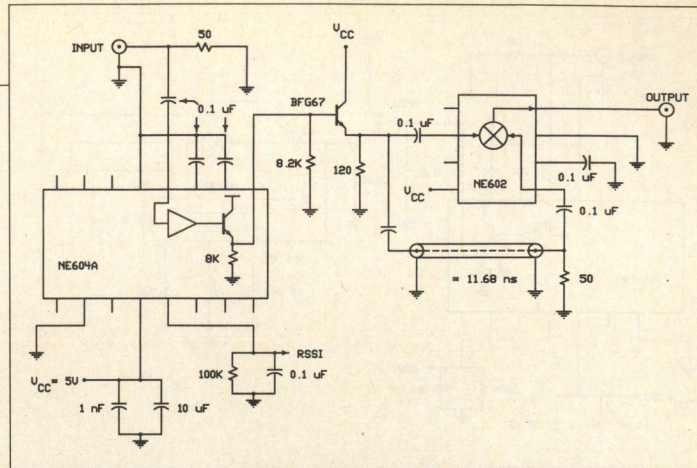


Figure 7a. 21.4 MHz delay line discriminator.

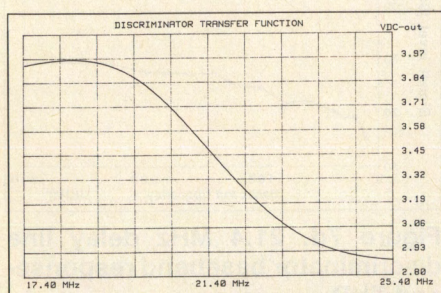


Figure 6b. Demodulator DC transfer function.

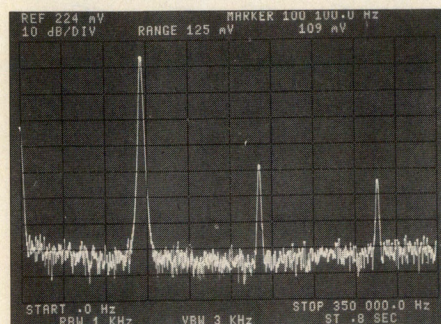
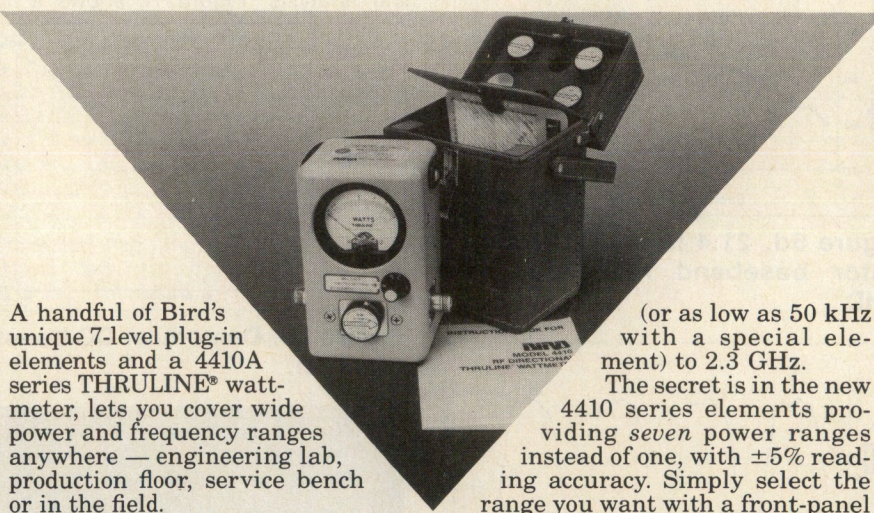


Figure 6c. Baseband output spectrum (600 kHz deviation).

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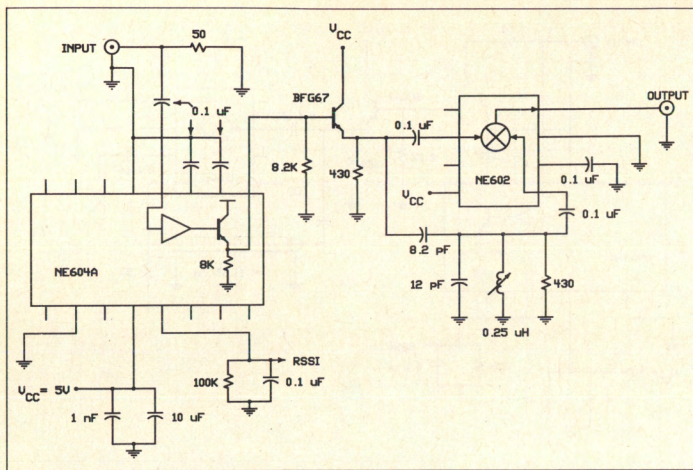


Figure 8a. 70 MHz LC discriminator.

loaded Q of 3.5. Figure 6b shows the discriminator's DC transfer function. The result agrees with analysis of the previous section. The 180 degree inversion of the curve is due to phase inversion of the Gilbert cell multiplier.

Figure 6c shows the baseband 100 kHz output of 109 mV with a 600 kHz frequency deviation and a THD (Total Harmonic Distortion) of 1 percent. Figure 6d shows the frequency response of the baseband and the THD plotted on the same curve. Note that the THD

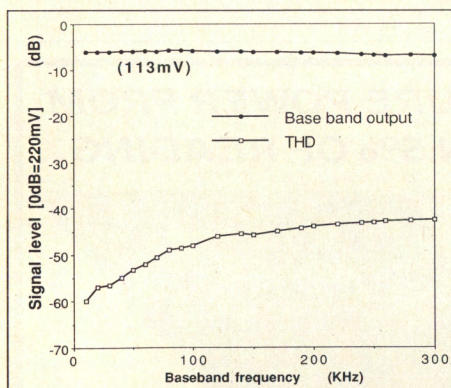


Figure 6d. 21.4 MHz LC discriminator baseband response and THD.

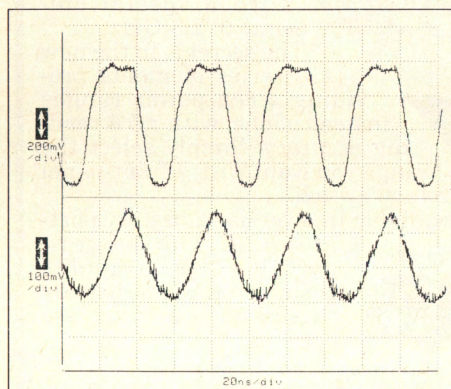


Figure 6e. Limited and phase shifted carrier in the time domain.

is a function of baseband frequency. With a maximum baseband frequency of 300 kHz, and 600 kHz frequency deviation, the total IF spectrum increases to 1.8 MHz (Carson's rule) resulting in less linearity. Figure 6e shows the 21.4 MHz limited carrier having 700 mV<sub>p-p</sub> amplitude and the phase shifted sinusoid (out of the quad tank; lower trace) with 250 mV<sub>p-p</sub>. When probing the quad tank for this measurement, care must be taken to avoid loading or de-tuning the resonator.

Figure 7a shows the 21.4 MHz delay line discriminator's performance. A 50 ohm termination is essential for proper operation of the delay line. Figure 7b shows the measured discriminator transfer function. The linearity range and the deviation from it agree with the theoretical analysis. Figure 7c shows a 13.3

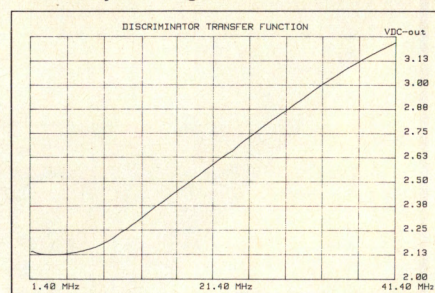


Figure 7b. Demodular DC transfer function.

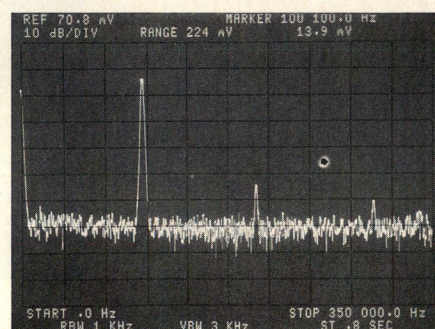


Figure 7c. Baseband output spectrum (600 kHz deviation).

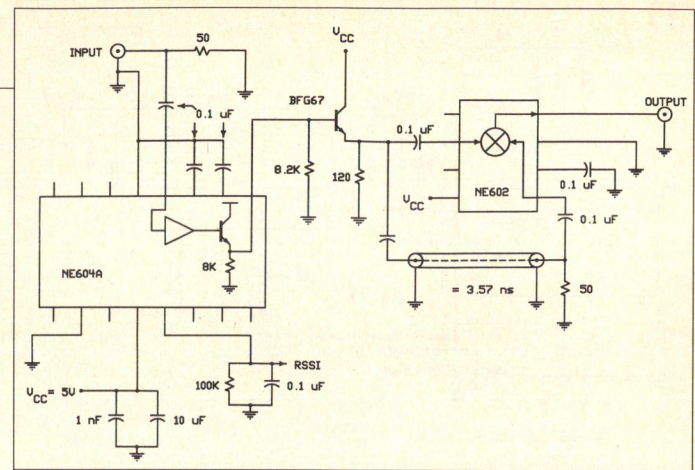


Figure 9a. 70 MHz delay line discriminator.

mV, 100 kHz output level with 1 percent distortion for 600 kHz deviation. Figure 7e shows the limiter output and the phase shifted signal in time domain. Both signals are 600 mV<sub>p-p</sub> in amplitude and close to a square wave with the 5th harmonic (106 MHz) attenuated due to bandwidth limitations.

Figure 8a shows the 70 MHz LC discriminator's performance. In the quad tank there is a trade-off between the coupling capacitor to the tank and the baseband output and its distortion. With

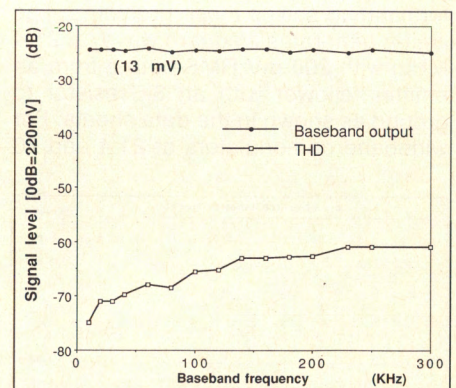


Figure 7d. 21.4 MHz delay line discriminator baseband response and THD.

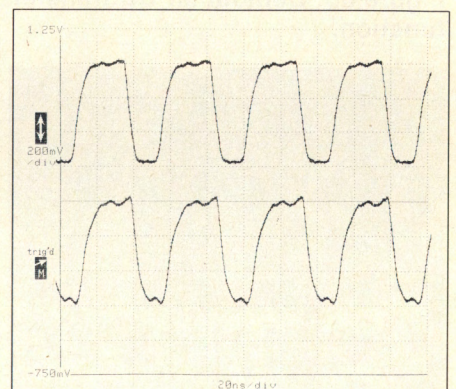
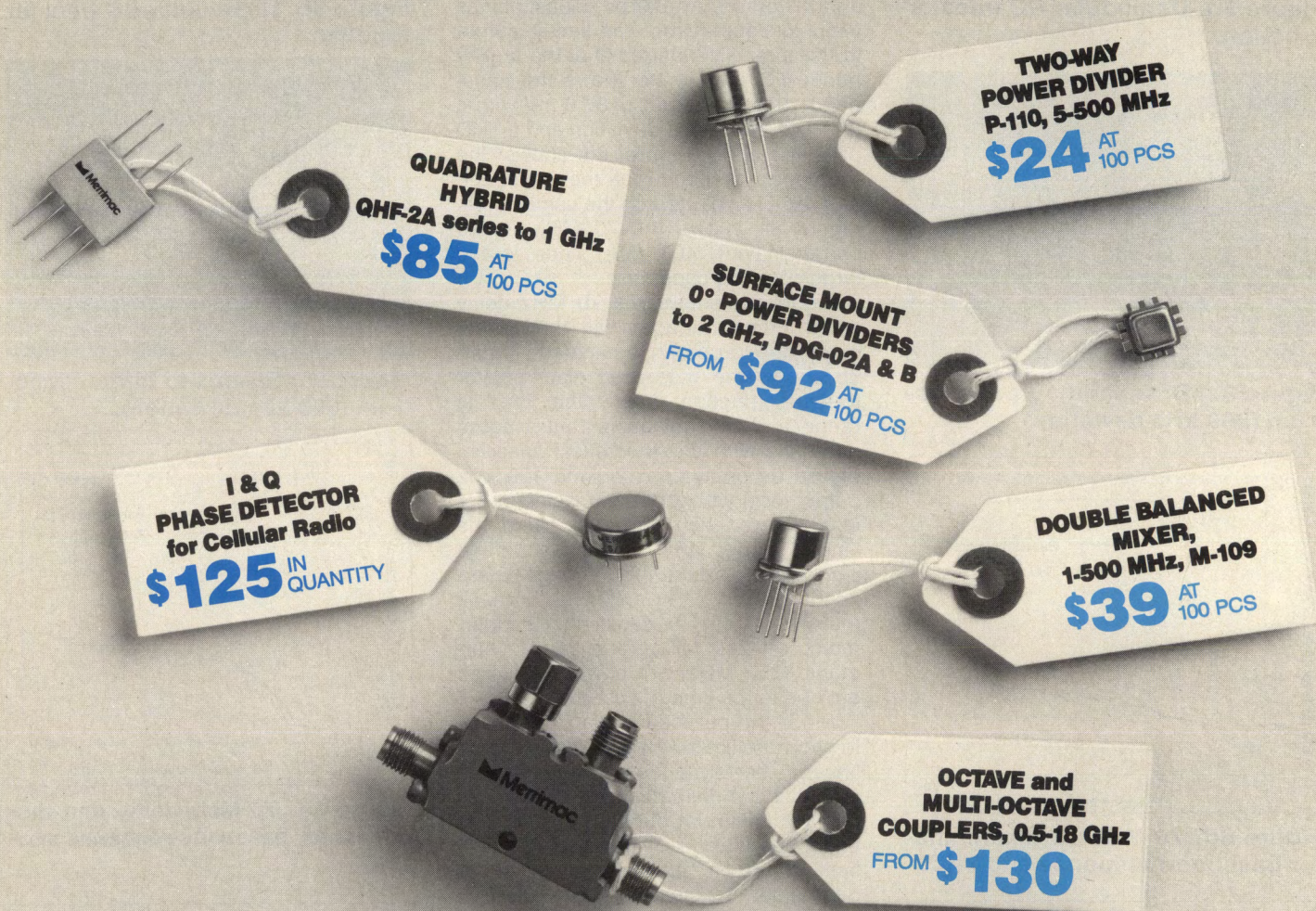


Figure 7e. Limited and phase shifted carrier in the time domain.



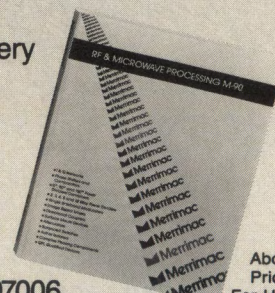
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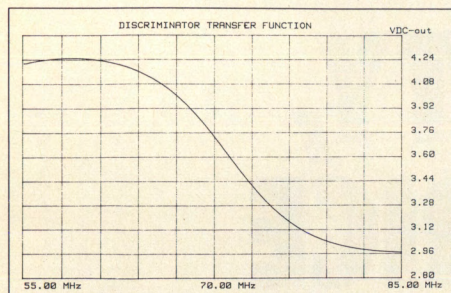


Figure 8b. Demodular DC transfer function.

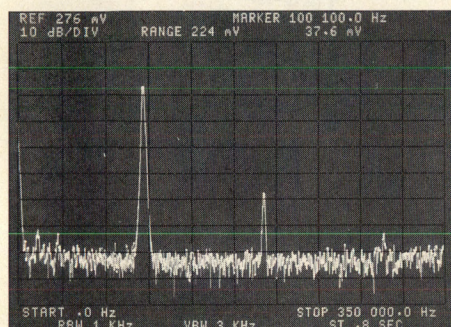


Figure 8c. Baseband output spectrum (600 kHz deviation).

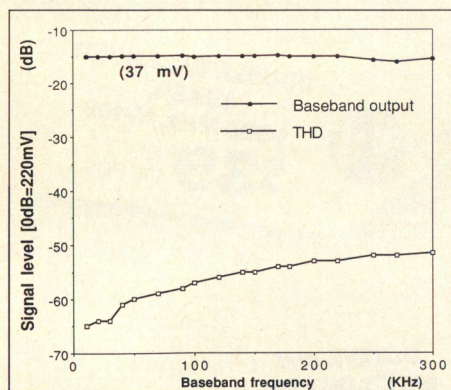


Figure 8d. 70 MHz LC discriminator baseband response and THD.

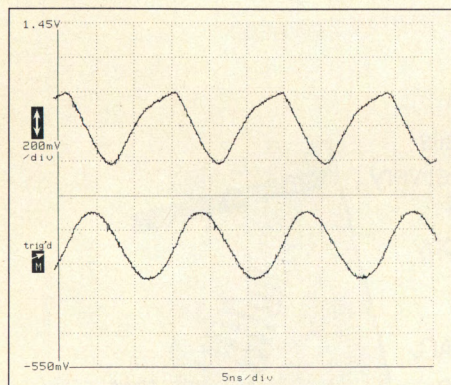


Figure 8e. Limited and phase shifted carrier in the time domain.

a small coupling capacitor a higher  $Q$  can be maintained. However, the output of the quad tank will be a low level sinusoid which will degrade the operation of the phase detector multiplier as shown in the theoretical analysis where a true sinusoidal input (effectively unclipped) was shown to degrade linearity very quickly. As a result the value of this capacitor should be optimized for the best combination of the baseband level and distortion. An 8.2 pF capacitor was used to couple from the limiter output to the tank. The loaded  $Q$  of the tank is about 4.5. Figure 8c shows the baseband 100 kHz output of 37.6 mV with a 600 kHz frequency deviation and a THD of 1 percent. Figure 8d shows the frequency response of the baseband and the THD plotted on the same curve. Figure 8e shows the 70 MHz limited carrier having 400 mV<sub>p-p</sub> due to the large coupling capacitor.

Figure 9a shows the 70 MHz delay line discriminator again with the 50 ohm delay line properly terminated. Figure 9b shows the measured discriminator transfer function where the linearity range and the deviation from it agree generally with the theoretical analysis. Figure 9c shows a 5.24 mV, 100 kHz output level with 2 percent distortion for 600 kHz deviation. Figure 9e shows the limiter output and the phase shifted signal in time domain. Both signals are about 380 mV<sub>p-p</sub> in amplitude but look more sinusoidal because the 3rd harmonic (at 210 MHz), and the 5th harmonic (350 MHz) are obviously filtered out due to the limiter's bandwidth.

Since the purpose of the circuits was to show the limiter and the discriminator linearity performance, no attempt was made to band-limit the input noise by using a minimum bandwidth (1.8 MHz) bandpass filter. As a result, only -3 dB limiting (the RF input level at which the demodulated baseband drops 3 dB from normal fully limited FM level) was measured. Full sensitivity tests will depend on the noise figure to be set by a low noise front-end.

As a final note, the RSSI output (Received Signal Strength Indicator) of the chip was also tested and found to exhibit a 60 dB range (logarithmic) at 21.4 MHz and a 50 dB range at 70 MHz (See Figure 10). Table 1 summarizes the performance of the four circuits.

#### Circuit Limitations

Figure 11 shows the test set-up used. To achieve wide frequency deviations an HP8640B RF at 550 MHz is mixed

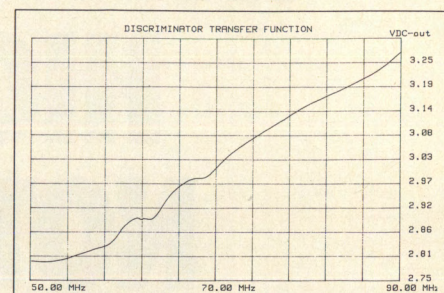


Figure 9b. Demodular DC transfer function.

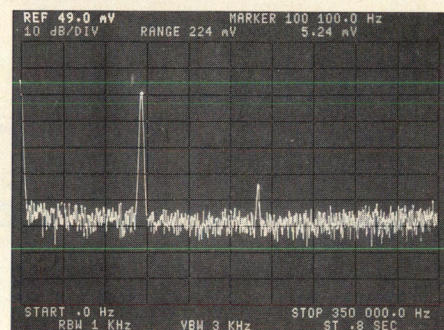


Figure 9c. Baseband output spectrum (600 kHz deviation).

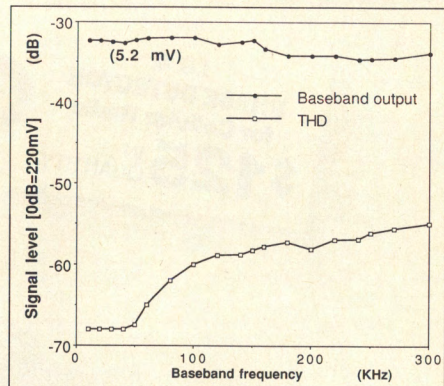


Figure 9d. 70 MHz delay line discriminator baseband response and THD.

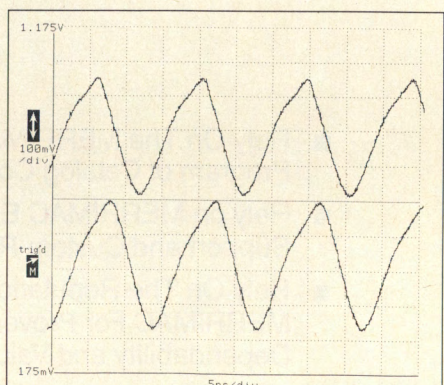
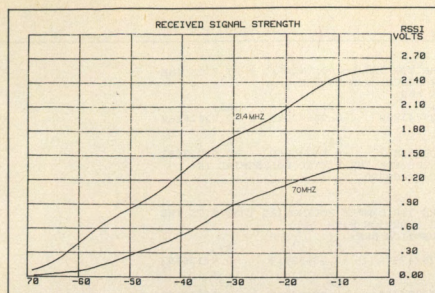


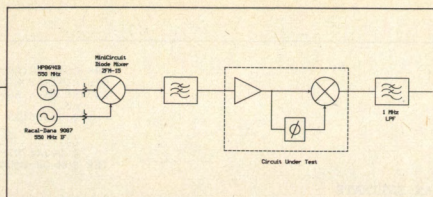
Figure 9e. Limited and phase shifted carrier in the time domain.





**Figure 10. RSSI response for 21.4 MHz and 70 MHz.**

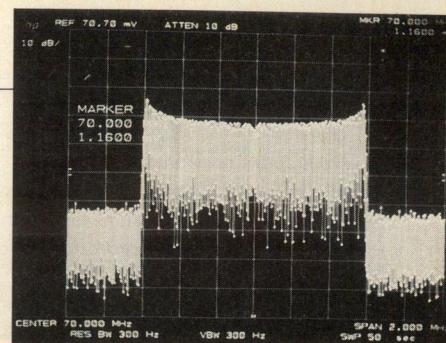
down to 21.4 MHz or 70 MHz by using a second generator. The output of the NE602 is DC coupled to a 1 MHz low pass filter. The distortion measurements are limited by the HP8640B 1 percent distortion specification. The instrument



**Figure 11. Test set up.**

also limits the modulating signal to 300 kHz. The reduction in the baseband level at 300 kHz is due to the generator and not the output bandwidth of the NE602 which is flat up to 15 MHz with a 1500 ohm output impedance.

Figures 6e-9e showed that the reduced input impedance IF limiter has at least 70 MHz bandwidth. Lack of small signal bandwidth in the limiter will cause



**Figure 12. 70 MHz IF response at -3 dB limiting.**

the distortion to be signal level dependent because of extra phase delay introduced when the IF limiter runs out of gain. Figure 12 shows the IF spectrum at 70 MHz with a 1 kHz tone and 600 kHz frequency deviation. This measurement has been done at -3 dB limiting. The symmetry of the signal while the limiter has run out of gain indicates a flat response at 70 MHz. In our circuits the distortion is not signal level dependent and the noise eventually captures the IF as expected.

As in any high speed IF stage, extreme care must be taken to avoid regenerative instabilities. These are radiated limiter output signals that are picked up by the limiter's input imped-

Discriminator Type	21.4 MHz IF		70 MHz IF	
	LC	Delay	LC	Delay
Baseband output level (100 KHz)	109 mV	13.3 mV	37.6 mV	5.24 mV
Distortion for 100 KHz baseband and 600 KHz deviation	1%	1%	1%	2%
-3 dB limiting	0.2 mV	0.15 mV	0.17 mV	0.18 mV
AM rejection for 1 KHz tone (30% AM, -40 dBm RF)	-25 dB	-18 dB	-17 dB	-15 dB
RSSI linear dynamic range	60 dB	60 dB	50 dB	50 dB

**Table 1. Performance Summary**

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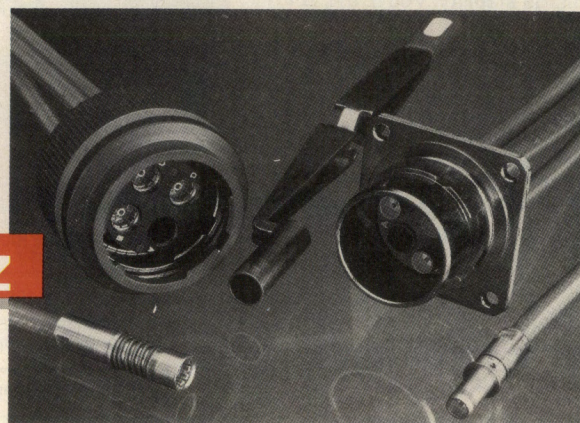
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### Example 3 (Telemetry application)

```
# INPUT PARAMETERS WITH A DECIMAL POINT
# WHAT IS YOUR IF FREQUENCY? (in MHZ)
21.4
# WHAT IS THE IF BANDWIDTH OVER WHICH YOU WANT
# TO HAVE A LINEAR DISCRIMINATOR? (in MHZ)
2.0
# FOR WHAT FREQUENCY DEVIATION DO YOU WANT TO SEE AN ESTIMATE
# OF DISTORTION? (in MHZ)
SET EQUAL TO 1/2 OF IF BANDWIDTH FOR WORST CASE RESULTS.
1.0
# HOW MANY POINTS SHOULD BE USED FOR THE ANALYSIS?
(MORE POINTS RESULT IN MORE ACCURACY, 100 Maximum)
20.
#WHAT IS THE LOADED Q FOR THE LC QUAD TANK?
5.0
```

\*\*\*\*\*

#N	FREQ	FREQ-DEV	PHASE	SQRE WVE	TRPZD WVE1	TRPZD WVE2	SIN WVE
1	20.400	-1.000	115.5824	-0.2842492	-0.2842492	-0.2765936	-0.2159046
2	20.500	-0.900	113.2577	-0.2584183	-0.2584183	-0.2526659	-0.1974333
3	20.600	-0.800	110.8614	-0.2317928	-0.2317928	-0.2276416	-0.1780539
4	20.700	-0.700	108.3988	-0.2044313	-0.2044313	-0.2015835	-0.1578148
5	20.800	-0.600	105.8766	-0.1764070	-0.1764070	-0.1745770	-0.1367834
6	20.900	-0.500	103.3027	-0.1478075	-0.1478075	-0.1467311	-0.1150476
7	21.000	-0.400	100.6858	-0.1187307	-0.1187307	-0.1181727	-0.0927112
8	21.100	-0.300	98.0360	-0.0892892	-0.0892892	-0.0890520	-0.0698979
9	21.200	-0.200	95.3642	-0.0596026	-0.0596026	-0.0595320	-0.0467435
10	21.300	-0.100	92.6816	-0.0297959	-0.0297959	-0.0297870	-0.0233931
11	21.400	0.000	89.9999	0.0000008	0.0000008	0.0000008	0.0000006
12	21.500	0.100	87.3307	0.0296593	0.0296593	0.0296506	0.0232859
13	21.600	0.200	84.6852	0.0590532	0.0590532	0.0589845	0.0463138
14	21.700	0.300	82.0742	0.0880639	0.0880639	0.0878363	0.0689448
15	21.800	0.400	79.5075	0.1165836	0.1165836	0.1160554	0.0910536
16	21.900	0.500	76.9939	0.1445124	0.1445124	0.1435064	0.1125275
17	22.000	0.600	74.5410	0.1717663	0.1717663	0.1700771	0.1332740
18	22.100	0.700	72.1553	0.1982740	0.1982740	0.1956758	0.1532187
19	22.200	0.800	69.8418	0.2239797	0.2239797	0.2202342	0.1723064
20	22.300	0.900	67.6045	0.2488384	0.2488384	0.2437023	0.1904985
21	22.400	1.000	65.4461	0.2728209	0.2728209	0.2660521	0.2077744

THE COEFFICIENTS OF THE 7TH DEGREE POLYNOMIAL THAT FITS THE DISCRIMINATOR TRANSFER FUNCTION WITH THE SQUARE WAVE ASSUMPTION ARE:

A0	A1	A2	A3
-0.00000103	0.29747990	-0.00692919	-0.02126262

A4, A5, A6, A7  
0.00141792 0.00255403 -0.00020260 -0.00023652

THE SUM-OF-SQUARED-ERRORS FOR THE FITTED POLYNOMIAL IS:  
0.00000000

THE COEFFICIENTS OF THE 7TH DEGREE POLYNOMIAL THAT FITS THE DISCRIMINATOR TRANSFER FUNCTION WITH THE TRAPEZOIDAL WAVE ASSUMPTION ARE:

A0	A1	A2	A3
-0.00000103	0.29747990	-0.00692919	-0.02126262

A4, A5, A6, A7  
0.00141792 0.00255403 -0.00020260 -0.00023652

THE SUM-OF-SQUARED-ERRORS FOR THE FITTED POLYNOMIAL IS:  
0.00000000

THE COEFFICIENTS OF THE 7TH DEGREE POLYNOMIAL THAT FITS THE DISCRIMINATOR TRANSFER FUNCTION WITH THE TRAPEZOIDAL WAVE ASSUMPTION ARE:

A0	A1	A2	A3
-0.00000141	0.29749090	-0.00691769	-0.03012002

A4, A5, A6, A7  
0.00197687 0.00455276 -0.00032921 -0.00060205

THE SUM-OF-SQUARED-ERRORS FOR THE FITTED POLYNOMIAL IS:  
0.00000000

THE COEFFICIENTS OF THE 7TH DEGREE POLYNOMIAL THAT FITS THE DISCRIMINATOR TRANSFER FUNCTION WITH THE SINUSOIDAL WAVE ASSUMPTION ARE:

A0	A1	A2	A3
-0.00000128	0.23363860	-0.00543066	-0.02518363

A4, A5, A6, A7  
0.00165373 0.00383577 -0.00028750 -0.00045154

THE SUM-OF-SQUARED-ERRORS FOR THE FITTED POLYNOMIAL IS:  
0.00000000

THE 2ND HARMONIC DISTORTION, (in dB)

SQRE WVE	TRPZD WVE1	TRPZD WVE2	SIN WVE
-38.198	-38.198	-38.005	-37.962

THE 3RD HARMONIC DISTORTION, (in dB)

SQRE WVE	TRPZD WVE1	TRPZD WVE2	SIN WVE
-34.479	-34.479	-31.248	-30.658

Portran STOP

Figure 13. Computer program example: Telemetry application with a 21.4 MHz LC quad discriminator.

ance. The reduction of the input impedance reduces this effect considerably. If more IF gain is needed, physical isolation (shielding) of the IF stages is

essential.

Finally, if wider band operation is needed, the Q of the tank can still be reduced. As long as the noise of the

system is dominated by the receiver front end and not by the phase detector, the audio output, no matter how small, can be post amplified without loss of the dynamic range.

## TCXO's in DIP packages

ALL Logic Families  
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DIP .5"x.8"x.375"

### SPECIFICATIONS

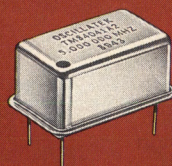
**Voltage stability:**  $V_{cc} \pm 5\%$ ,  $\Delta F$   
 $< 1\text{PPM}$

$V_{cc} \pm 10\%$ ,  $\Delta F$   
 $< 12\text{PPM}$

**Aging:** 5 PPM first year; 2PPM per year  
 thereafter

**Frequency Adj.:** Control voltage in  
 range of 9 to 5V (variable capacitor  
 optional)

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TEMPERATURE RANGE/ STABILITY	FREQUENCIES AVAILABLE
0° - +50°C, $\pm 2\text{PPM}$	TTL = 1KHz to 32MHz
0° - +50°C, $\pm 5\text{PPM}$	*CMOS = 1KHz to 10MHz
0° - +70°C, $\pm 5\text{PPM}$	ECL = 4MHz to 32MHz
-30° - +70°C, $\pm 10\text{PPM}$	SINE = 4MHz to
-30° - +85°C, $\pm 15\text{PPM}$	

### Computer Program

A program written in standard Fortran 77 has been written to accomplish the above-described performance calculations. Figure 13 is an example using 2 MHz bandwidth and 21.4 MHz IF, as might be found in a telemetry application. Only the LC quad configuration case is shown here, although the program generates performance data for the delay line discriminator, as well. The program is available on disk from the RF Design Software Service. See page 71 for ordering information.

RF

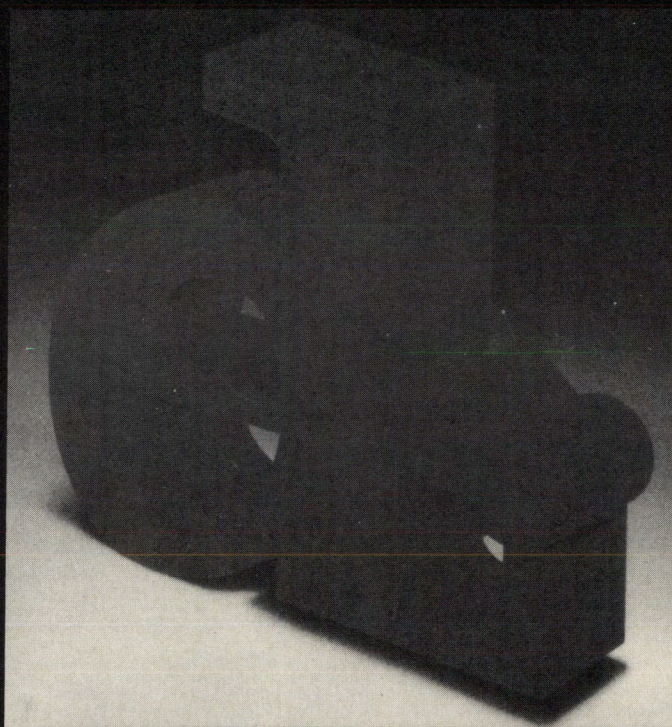
### About the Authors

Alvin Wong is an RF Applications Engineer at Signetics. He holds a BSEE from San Jose State University. Ali Fotowat is an RF Design Engineer at Signetics. He holds a BS degree from CalTech, and an MS and PhD degrees from Stanford University. They may be reached at: Signetics Company, 811 E. Arques Ave., PO Box 3409, M/S 60, Sunnyvale, CA 94088-3409. Tel: (408) 991-2000.



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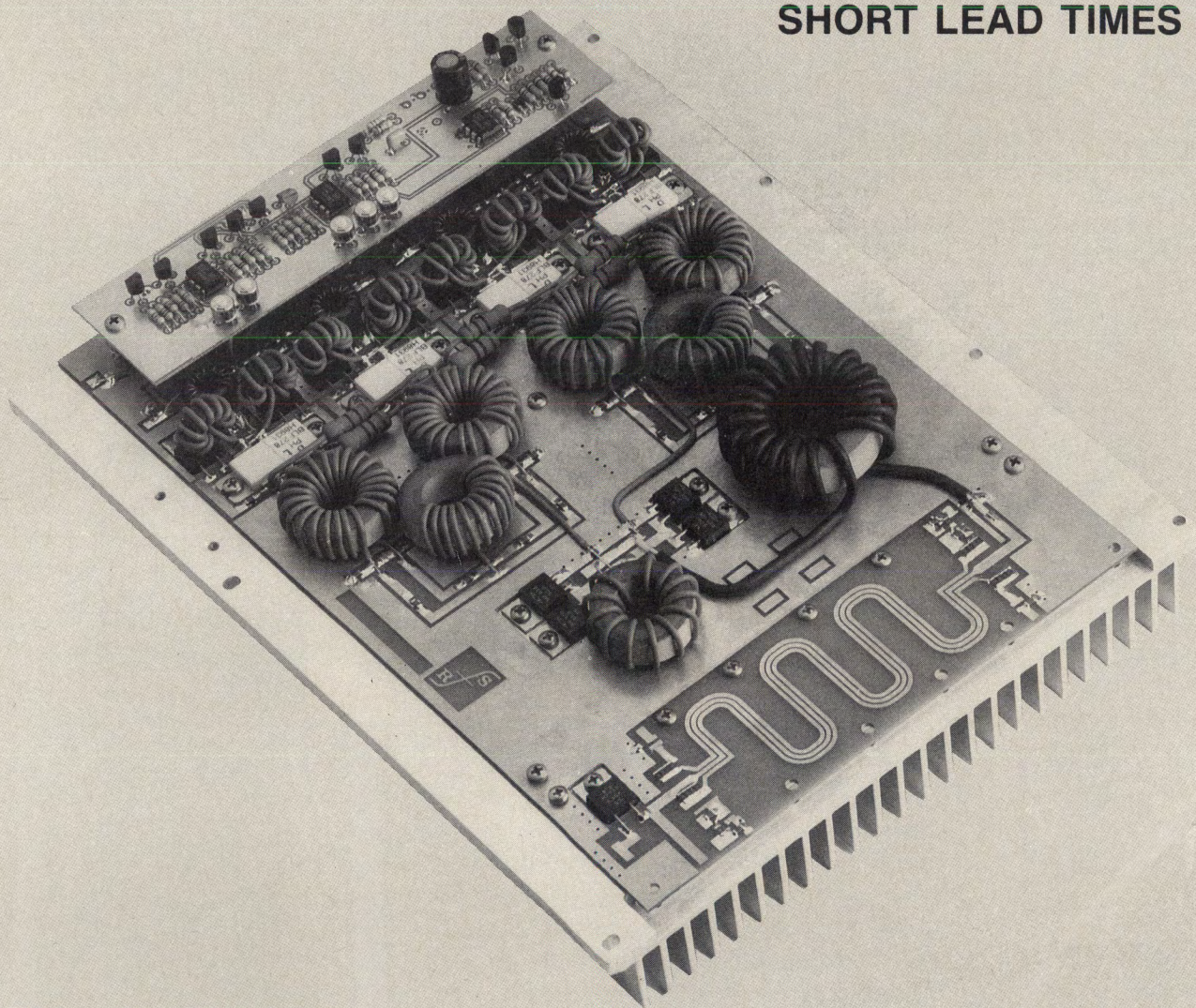
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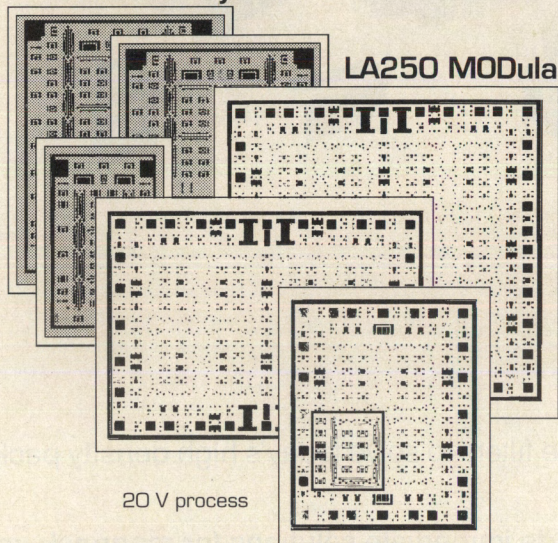
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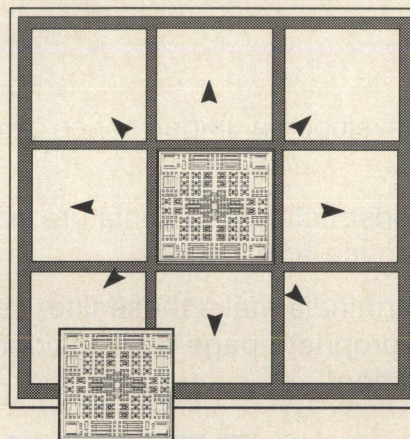
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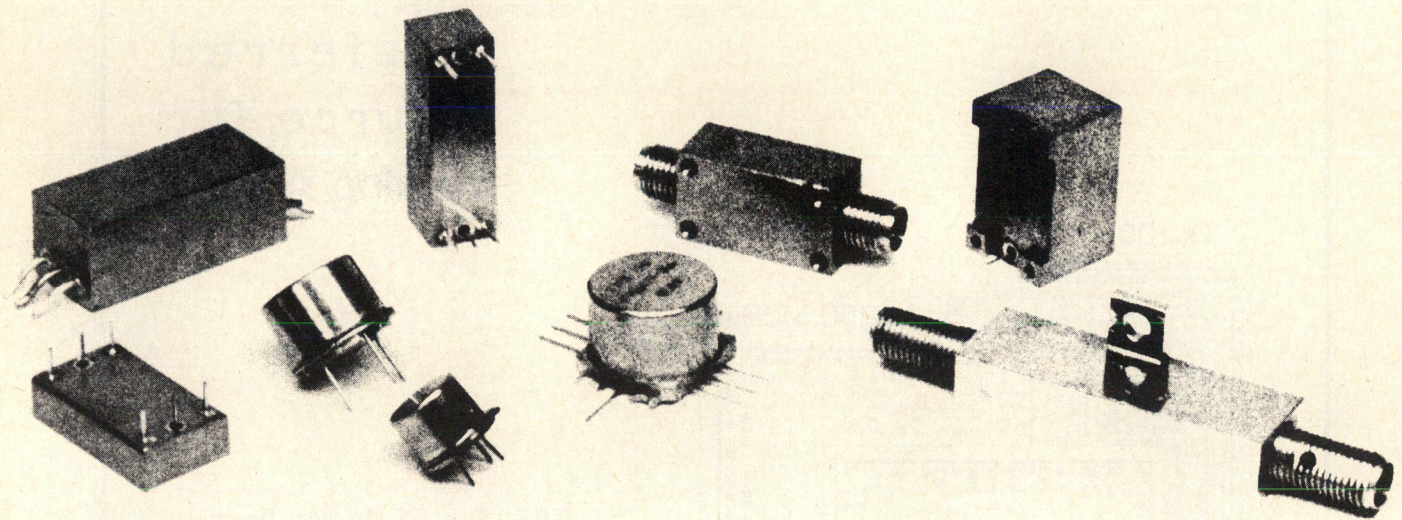
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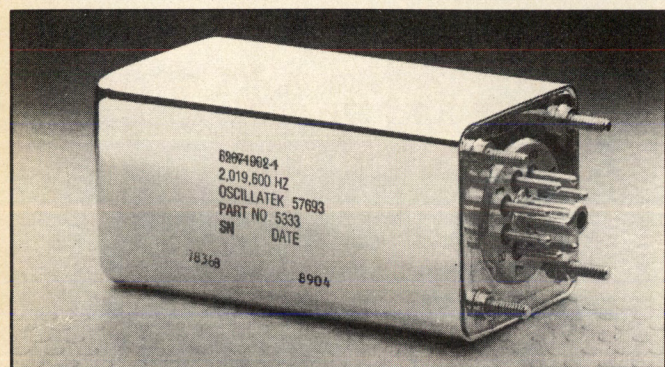
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**Output Options (Standard):**  
TTL, CMOS, HCMOS, or SINE:  
+7dBm, 50  $\Omega$   
**Harmonics:** -20dBc

**Supply (Standard):**  
**Voltage:** 24VDC  $\pm$  5%  
**Stability:**  $1 \times 10^{-9}$ /percent

**Power:**  
**Turn-on:** 8 W maximum  
**Stabilized @ +25°C:** 3 W maximum

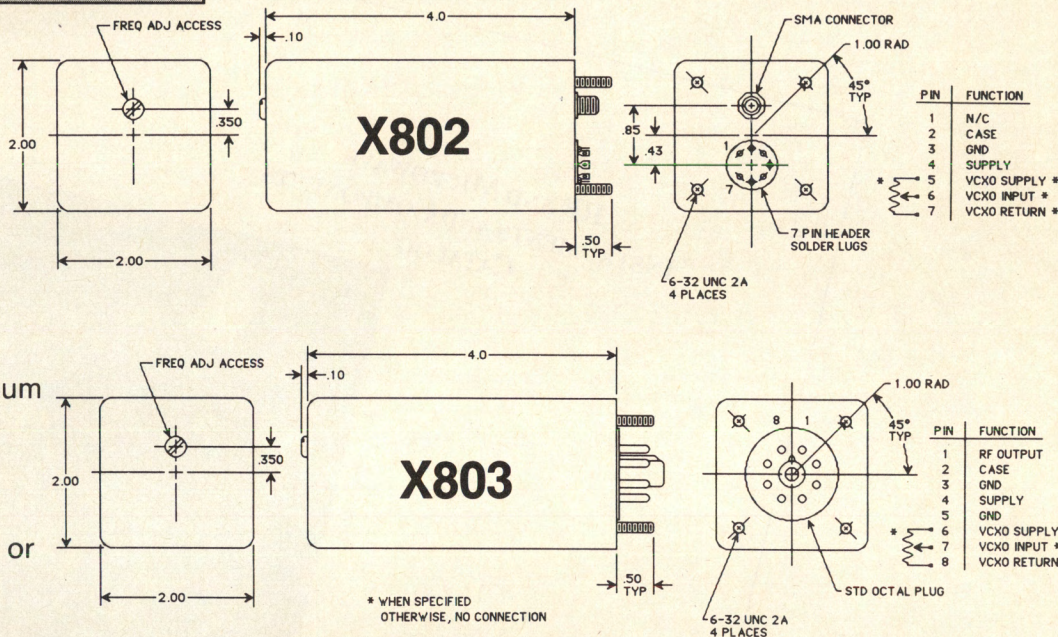
**Frequency Adjust:**  
**Mechanical:** For 10 years aging  
**Electrical (with option "E"):**  
 $3 \times 10^{-7}$  minimum for 0 to +5V or  
external 20K potentiometer

**Phase Noise (Sine, 10MHz):**

Offset	SSB Level
100 Hz	-130dBc/Hz
1 KHz	-140dBc/Hz
10 KHz	-145dBc/Hz
50 KHz	-150dBc/Hz

**Supply Options:**  
Any specified voltage in the range of 12VDC to 28VDC.  
Multiple supplies for oven, RF and Logic.

**Output Options (Additional):**  
**Sine Output:** Levels from -20dBm to +13dBm  
**ECL:** Available with addition of -5.2V supply



## ORDERING METHOD

OUTPUT LOGIC	MODEL NUMBER	AGING PER DAY	TEMP. RANGE	FREQUENCY STABILITY OPTION		VOLTAGE CONTROL	FREQUENCY	SUPPLY VOLTAGE
T = TTL	X802	1 = $1 \times 10^{-8}$	A = 0°C to +50°C	39 = $\pm 3 \times 10^{-9}$	28 = $\pm 2 \times 10^{-8}$ 18 = $\pm 1 \times 10^{-8}$	E IF VOLTAGE CONTROL OPTION IS DESIRED	TTL = .01Hz to 30MHz	SPECIFIED SUPPLY VOLTAGE: 12VDC TO 28VDC
S = SINE	X803	2 = $5 \times 10^{-9}$	B = 0°C to +70°C	59 = $\pm 5 \times 10^{-9}$	58 = $\pm 5 \times 10^{-8}$ 18 = $\pm 1 \times 10^{-8}$		SINE = 100KHz to 30MHz	
C = CMOS		3 = $3 \times 10^{-9}$	C = -20°C to +70°C	18 = $\pm 1 \times 10^{-8}$	58 = $\pm 5 \times 10^{-8}$ 28 = $\pm 2 \times 10^{-8}$		CMOS = .01Hz to 15MHz	
HC = HCMOS		4 = $1 \times 10^{-9}$	D = -40°C to +70°C	28 = $\pm 2 \times 10^{-8}$	17 = $\pm 1 \times 10^{-7}$ 58 = $\pm 5 \times 10^{-8}$		HCMOS = .01Hz to 30MHz	
			E = -55°C to +70°C	58 = $\pm 5 \times 10^{-8}$	27 = $\pm 2 \times 10^{-7}$ 17 = $\pm 1 \times 10^{-7}$			

EXAMPLE

S	X802	2	B	18	-	10 MHz	,	24V
---	------	---	---	----	---	--------	---	-----

**NOTE:** SX8022B18-10MHz,24V is a catalog number which defines an ovenized crystal oscillator in the X802 package with a 10MHz Sine output, aging of  $5 \times 10^{-9}$ /day, stability of  $\pm 1 \times 10^{-8}$  over the temperature range of 0°C to +70°C, operating on a +24VDC supply.

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## PTS 040

Range: 0.1-40 MHz  
Resolution: 0.1Hz-100KHz (opt)  
Switching: 1-20 $\mu$ s

Output: +3 to +13dBm; 50ohm  
Spurious Outputs: -75dB  
Phase Noise: -75dBc (0.5Hz-15KHz)

Freq. St'd: OCXO, TCXO, Ext.  
Interface: BCD par. or GPIB  
Price: \$4,800.00\*

## PTS 120

Range: 90-120 MHz  
Resolution: 0.1Hz-100KHz (opt)  
Switching: 1-20 $\mu$ s

Output: +3 to +10dBm; 50ohm  
Spurious Outputs: -75dBc  
Phase Noise: -75dBc (0.5Hz-15KHz)

Freq. St'd: OCXO, TCXO, Ext.  
Interface: BCD par. or GPIB  
Price: \$4,800.00\*

## PTS 160

Range: 0.1-160 MHz  
Resolution: 0.1Hz-100KHz (opt)  
Switching: 1-20 $\mu$ s

Output: +3 to +13dBm; 50ohm  
Spurious Outputs: -75dB  
Phase Noise: -63dBc (0.5Hz-15KHz)

Freq. St'd: OCXO, TCXO, Ext.  
Interface: BCD par. or GPIB  
Price: \$5,850.00\*

## PTS 250

Range: 1-250 MHz  
Resolution: 0.1Hz-100KHz (opt)  
Switching: 1-20 $\mu$ s

Output: +3 to +13dBm; 50ohm  
Spurious Outputs: -70dB  
Phase Noise: -63dBc (0.5Hz-15KHz)

Freq. St'd: OCXO, TCXO, Ext.  
Interface: BCD par. or GPIB  
Price: \$6,700.00\*

## PTS 300

Range: 0.1-300 MHz  
Resolution: 1Hz  
Switching: 1-20 $\mu$ s  
Phase Continuous: 1Hz-100KHz steps

Output: +3 to +13dBm; 50ohm  
Spurious Outputs: Type 1  
-70/65 (typ/spec)  
Phase Noise: -68dBc (0.5Hz-15KHz)

Type 2  
-65/60dB  
-63dBc

Freq. St'd: OCXO, TCXO, Ext.  
Interface: BCD par. or GPIB  
Price: Type 1 \$5,800.00\* Type 2 \$5,300.00\*

## PTS 500

Range: 1-500 MHz  
Resolution: 0.1Hz-100KHz (opt)  
Switching: 1-20 $\mu$ s

Output: +3 to +13dBm; 50ohm  
Spurious Outputs: -70dB  
Phase Noise: -63dBc (0.5Hz-15KHz)

Freq. St'd: OCXO, TCXO, Ext.  
Interface: BCD par. or GPIB  
Price: \$7,850.00\*

## PTS x10

Range: 10 MHz band, selected decade 0.1-100 MHz  
Resolution: 1Hz  
Switching: 1-5 $\mu$ s  
Phase Continuous: 2 MHz band, even or odd steps

Output: +3 to +13dBm; 50ohm  
Spurious Outputs: -65/-60dB (typ/spec)  
Phase Noise: -70dBc (0.5Hz-15KHz)

Freq. St'd: OCXO, TCXO, Ext.  
Interface: BCD par. or GPIB  
Price: \$2,450.00\*

## PTS 620

Range: 1-620 MHz  
Resolution: 0.1Hz-100KHz (opt)  
Switching: 1-20 $\mu$ s

Output: +3 to +13dBm; 50ohm  
Spurious Outputs: -70dB  
Phase Noise: -63dBc (0.5Hz-15KHz)

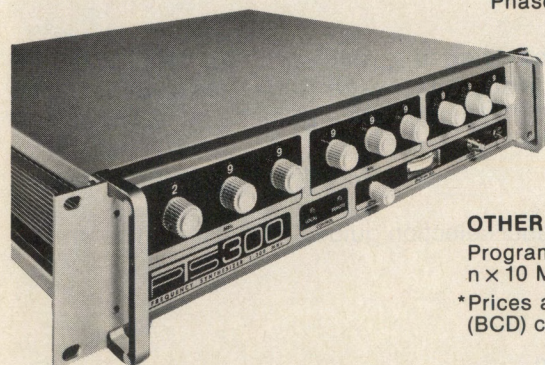
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Interface: BCD par. or GPIB  
Price: \$8,650.00\*

## PTS 1000

Range: 0.1-1000 MHz  
Resolution: 0.1Hz-100KHz (opt)  
Switching: 5-10 $\mu$ s

Output: +3 to +13dBm; 50ohm  
Spurious Outputs: -70dB (0.1-500 MHz),  
-65dB (500-1000 MHz)  
Phase Noise: -60dBc (0.5Hz-15KHz)

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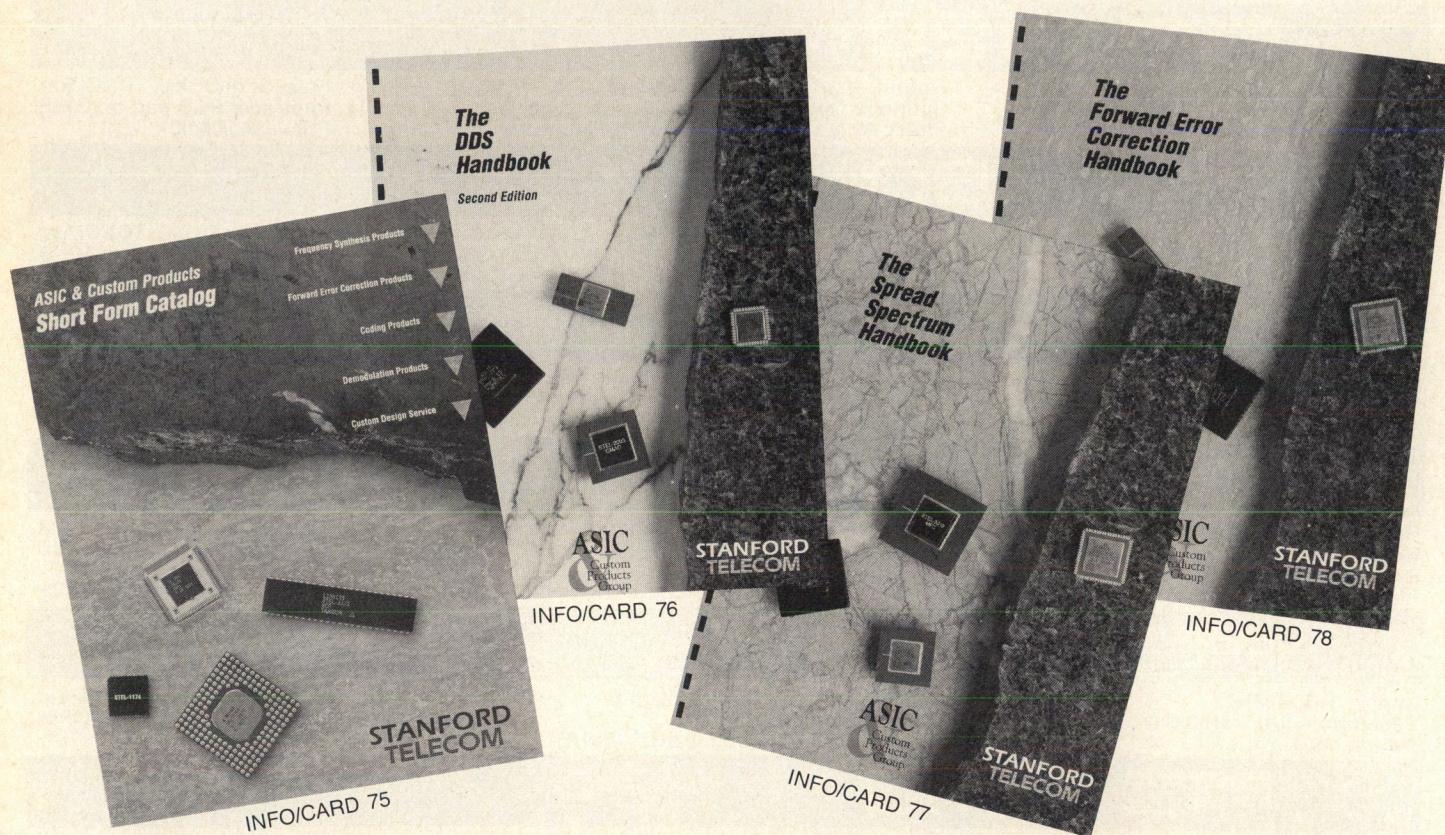
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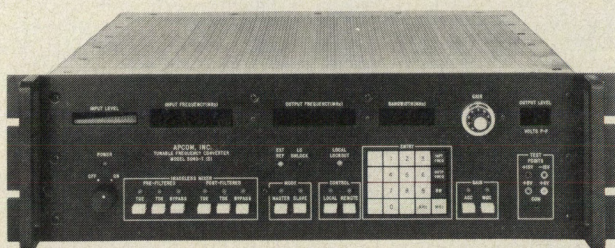
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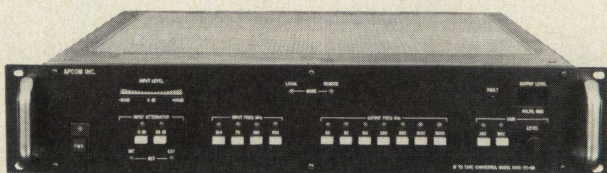


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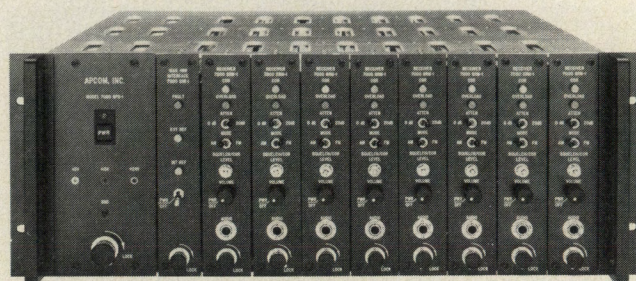
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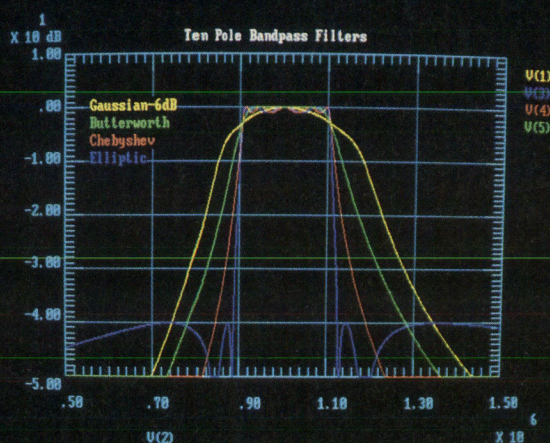


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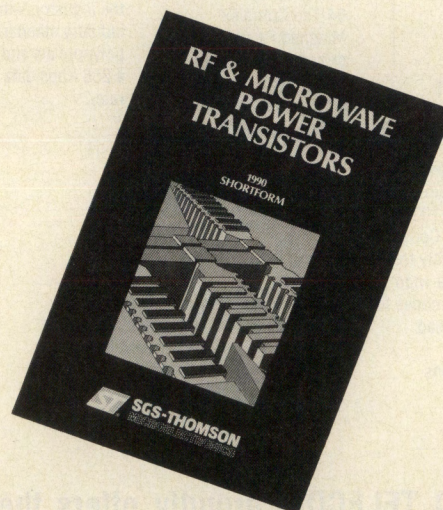
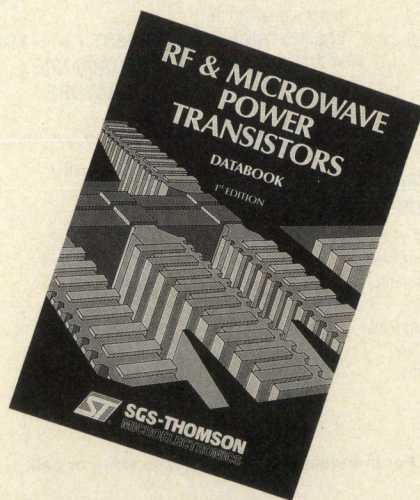
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## Using Current Feedback Amplifiers

By Al Little  
Signal Processing Products  
Harris Semiconductor

*Current feedback amplifiers are becoming increasingly popular for high frequency and RF designs. They are used like ordinary op amps, yet they offer higher speed, nearly constant bandwidth at increased gains, and a number of other benefits. This article describes these benefits as well as the special features and applications of the Harris HA-5004, a 100 MHz current feedback amplifier.*

Although conventional op amps have been greatly improved over the past decades, very few are good enough for accurately processing analog signals approaching 100 MHz. At these frequencies, the open loop gain (and AC performance) of op amps usually drops so low that they are unusable. Even those that do work at high frequency often need to be tweaked with external feedback networks to achieve the best performance. For applications that need more than unity gain, even the best high speed op amps may still fall short due to the well known gain-bandwidth trade-off.

To combat this shortcoming, current feedback amplifiers have now become widely used. Due to their unique circuit topology, current feedback amplifiers side-step the "gain-bandwidth" tradeoff of conventional op amps altogether, delivering nearly equal bandwidth over a wide range of gains. This feature makes them ideal for high frequency applications like video drivers, pulse amplifiers, radar and IF signal processing.

Figures 1 and 2 illustrate this basic difference between conventional op amps and current feedback amplifiers. Figure 1 shows the frequency response of a 100 MHz op amp at several closed loop gains. As shown, the unity gain bandwidth actually exceeds 100 MHz, but some undesired gain peaking also occurs. At higher closed loop gains, the bandwidth is proportionally lower, such that at a gain of 10 (20 dB) the amplifier delivers less than 10 MHz of signal bandwidth.

By contrast, Figure 2 shows the same response for the Harris HA-5004, a high

performance monolithic current feedback amplifier. At unity gain, the amplifier also provides a bandwidth of 100 MHz, but without significant gain peaking. At a gain of 10, the HA-5004 rolls off at only 80 MHz, providing 8 times the bandwidth of the comparable op amp.

Extended bandwidth is by no means the only advantage of current feedback amplifiers. Unlike op amps, current feedback amplifiers do not exhibit slew rate limiting. This remarkable property is illustrated in Figure 3, in which the large signal response of the HA-5004 looks virtually identical to the small signal response. With no slew rate limiting, the output rise time and fall time remain constant, independent of signal amplitude. This linear characteristic translates to extremely low distortion, which makes the amplifier ideal for high fidelity video and RF signal processing applications.

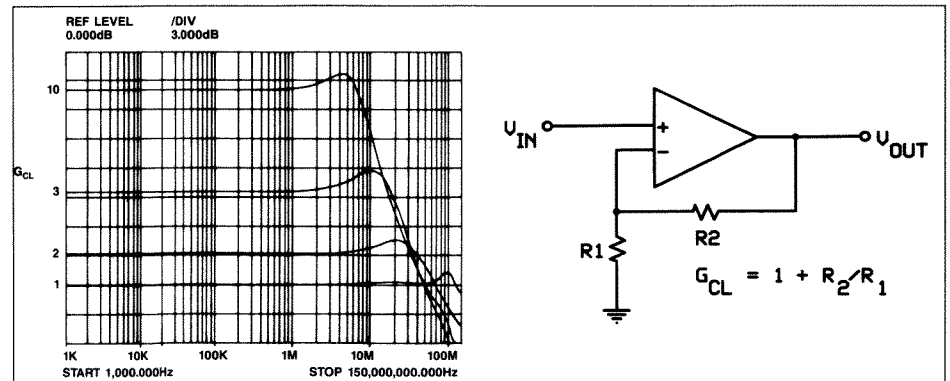
Current feedback amplifiers achieve their superior performance by using a slightly different principle than traditional op amps. Figures 4a and 4b illustrate these differences. Inside the current feedback amplifier, there is a unity gain buffer from the non-inverting (+) terminal to the inverting (-) terminal. The inverting terminal is, by definition, a low impedance point at all times. Error currents are sensed at the inverting input and amplified such that a small change in input current produces a large change in output voltage. The ratio of

output voltage delta due to input current delta is the transimpedance,  $Z$ , of the device. Like voltage gain in a conventional op amp, transimpedance is a function of frequency.

Steady state currents at the inverting input are very small because the transimpedance is large (typically 100 V/mA in the HA-5004). The voltage across the input terminals is nearly zero (typically 1 mV) due to the small offset voltage of the buffer amplifier. The ideal properties of zero input current and zero offset voltage are also true for current feedback amplifiers, and likewise simplify circuit design and analysis. The response to reactive feedback elements, however, is entirely different than to op amps due to this difference in structure, so great care must be used. Fortunately, current feedback amplifiers often eliminate the need for reactive feedback (like feed-forward designs to compensate op amps) in the first place.

Equations 1a through 4a show the relationships for traditional operational amplifiers. In equation 2a, as long as the open loop gain  $A(s)$  is much greater than the ideal closed loop gain  $G$ , the transfer function closely approximates the ideal. As frequency increases negatively or as  $G$  is increased negatively the overall bandwidth is limited by the bandwidth of  $A(s)$ . These equations demonstrate why voltage feedback amplifiers must always trade bandwidth for gain.

In the case of current feedback,



**Figure 1. Frequency response and schematic of a 100 MHz op amp at several closed loop gains.**



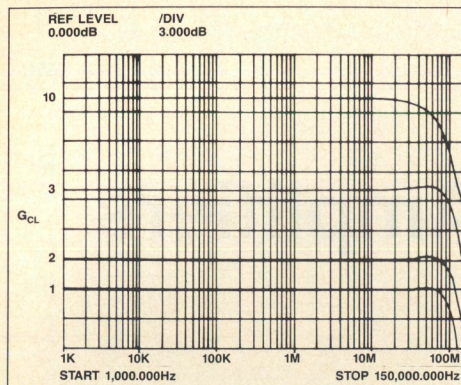


Figure 2. Frequency response and schematic of a 100 MHz current feedback amplifier at several closed loop gains.

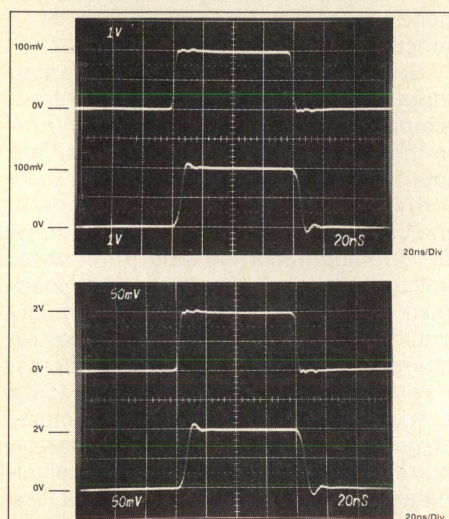


Figure 3. Comparison of slew rate limiting of the large signal response and the small signal response for the Harris HA-5004 current feedback amplifier.

however, while the transfer function 2b is similar, the closed loop bandwidth is no longer a function of gain  $G$ . Instead, bandwidth is limited only by the ratio of  $R_2$  to  $Z(s)$ , independent of the ideal closed loop gain  $G$ . Although  $Z(s)$  decreases with frequency like  $A(s)$  for op amps (equations and graphs 4a and 4b), it does not factor into the closed loop bandwidth the same way. Thus, the closed loop bandwidth for current feedback amplifiers is independently set by  $R_2$ . In actual use, bandwidth is slightly reduced at higher gains due to the non-zero output impedance at the inverting terminal. This bandwidth reduction can easily be recovered, however, by simply lowering the value of  $R_2$ . It should also be noted that  $R_2$  is always required, even for unity gain configurations.

Current feedback amplifiers are ideal for a wide range of high frequency applications. Figure 5a shows how the HA-5004 can be used as a basic video buffer. Setting the amplifier for a gain

of +2 provides unity gain with matching 50 Ohm output and load resistances. Resistors  $R_{c+}$  and  $R_{c-}$  may be used, if needed, to limit the output current for fault conditions.

Since the HA-5004 typically requires only 12 mA of quiescent current to operate, power dissipation is rarely a problem. But for particularly demanding applications at high temperature and output current, the HA-5004 has the capability to automatically sense overheating and shut down its output stage. Special circuitry triggers when the junction temperature exceeds approximately 180 degrees Celsius. If this occurs, an open collector output signal (TOL) can be used to notify microprocessor circuitry that a thermal overload condition exists. Once disabled, and the chip temperature drops, the HA-5004 will automatically reactivate itself and resume normal operation. If desired, a TTL input (TOI) can override or prevent the disable operation; in this case, thermal overload will not be indicated by the TOL output.

By using the output enable (OE) function manually, the HA-5004 can be configured in parallel to form a multiplexed video amplifier as shown in figure 5b. The TTL-compatible Enable control turns off the amplifier so that outputs can be tied together in common. Only one pair of output protection resistors is needed in this case since only one output is active at any time. When switching takes place from one channel to the next, a small overlap may occur in which one amplifier becomes active before the last has reached a high impedance state. To prevent this condition, the enable command to each amplifier should be skewed by a few microseconds.

Another common application for the HA-5004 is to buffer the input of a flash A/D converter as shown in Figure 6. This function can be deceptively demanding on a buffer amplifier because of the high transient currents associated with CMOS

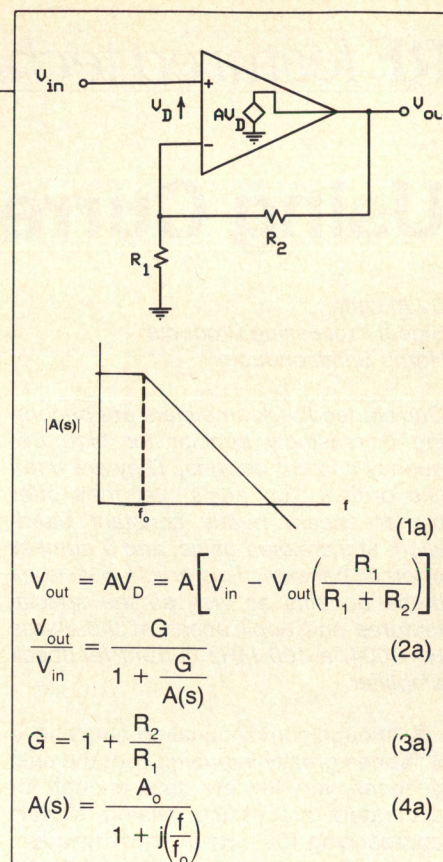


Figure 4a. Voltage feedback amplifier schematic.

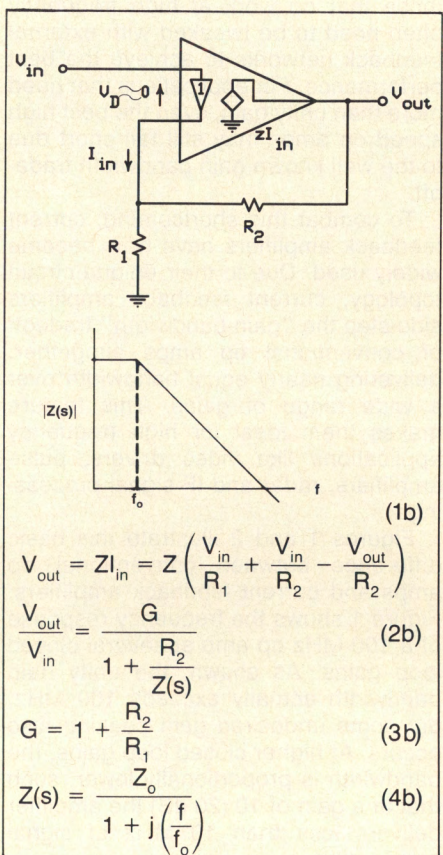
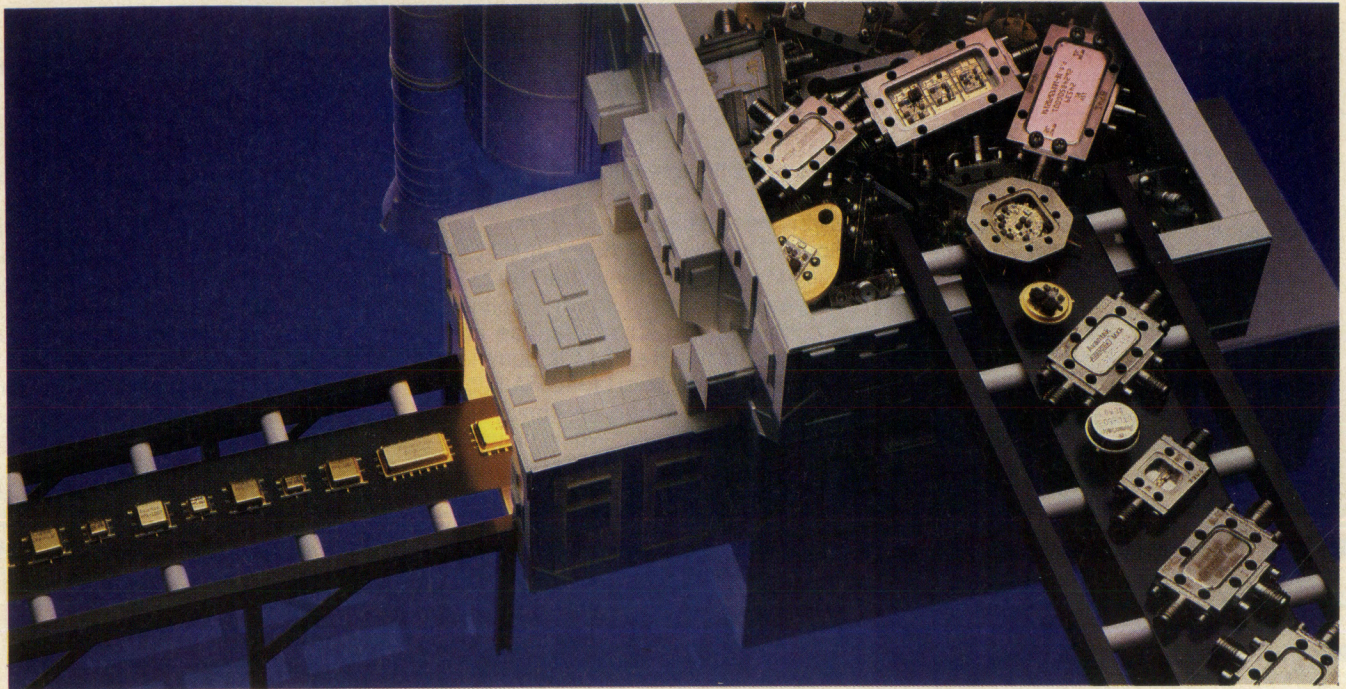


Figure 4b. Current feedback amplifier schematic.



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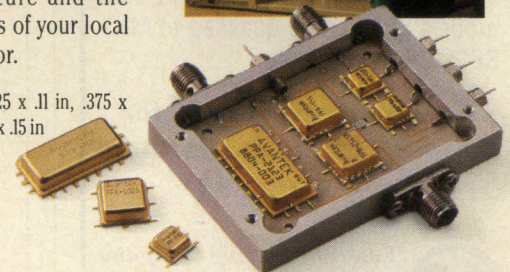
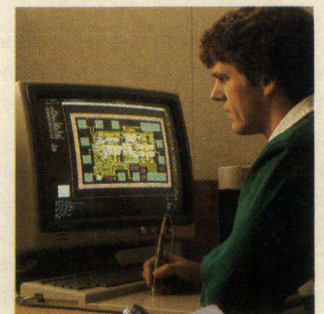
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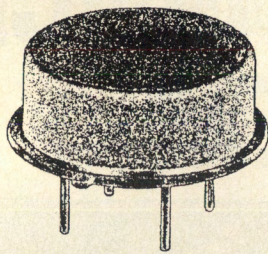
†Three sizes: .25 x .25 x .11 in, .375 x .375 x .15 in and .4 x .8 x .15 in

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flash converters. The buffer must quickly charge the input capacitance at the beginning of each conversion cycle, settling to within 1/2 LSB before the sample time is complete. In the case of the HI-5700 8-bit flash converter operating at 20 MHz with a 50 percent duty cycle clock, full settling must be complete within 25 ns. The 100 mA output current capability of the HA-5004 makes it ideal for fast settling, high current applications like this.

Figure 7 shows how the HA-5004 can be used in conjunction with the HA-2547 high speed multiplier to form a voltage controlled amplifier or mixer. The current output of the HA-2547 is summed at the inverting terminal of the HA-5004, providing current-to-voltage conversion the same way an op amp would be used. Using a 250 Ohm feedback resistor, the full 100 MHz bandwidth of the current feedback amplifier preserves the bandwidth capability of the HA-2547. Alternately,

the internal feedback of the 2547 may be used instead, with some sacrifice to the bandwidth.

With few exceptions, the rules for properly applying current feedback amplifiers are the same as those for high speed op amps:

- Keep all component leads as short as possible, particularly at the inverting input. Always minimize the stray capacitance at this node.
- Separate signal grounds from power grounds and connect them together at only one common (star) ground point.
- Use properly terminated coaxial cable at the input and output if they are located some distance from the amplifier.
- For best performance, use a ground plane for PC mounted devices.
- Make input and feedback resistors as small as possible consistent with the specified feedback resistance, output

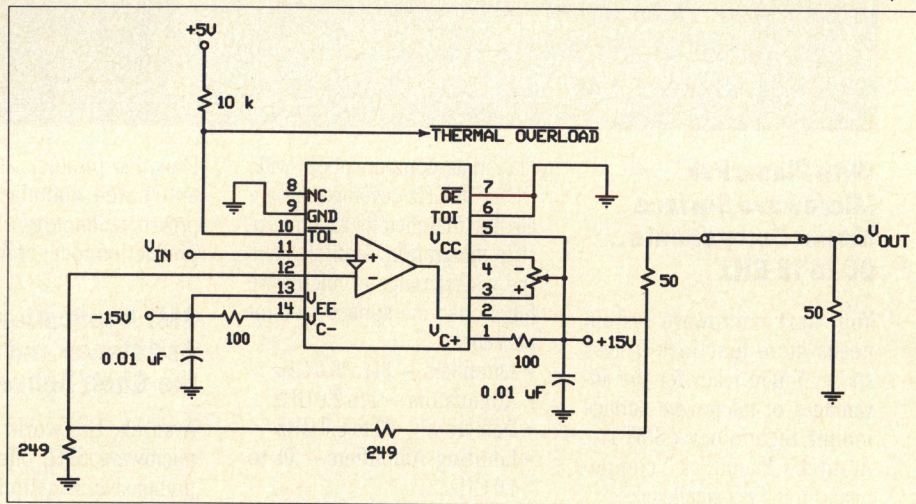


Figure 5a. Current feedback amplifier used as a video buffer.

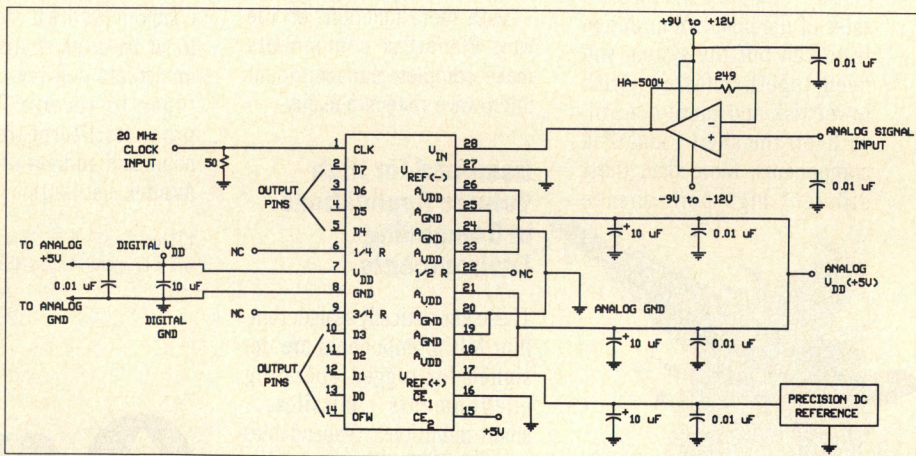
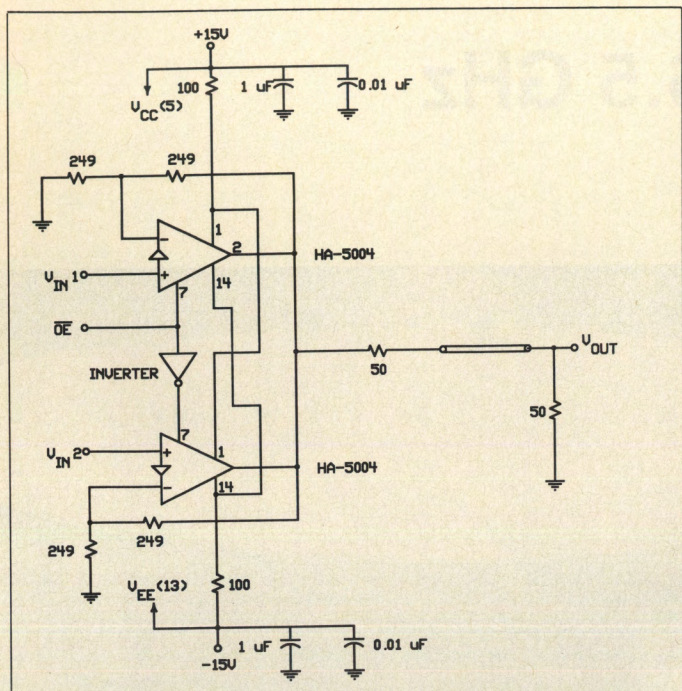


Figure 6. Current feedback amplifier used as a buffer for a flash A/D converter.





**Figure 5b.** Current feedback amplifier used as a multiplexed video amplifier.

drive capability, and circuit requirements. In current feedback amplifiers, bandwidth can usually be increased somewhat by using lower value feedback resistors. Values too small, however, may cause stability problems.

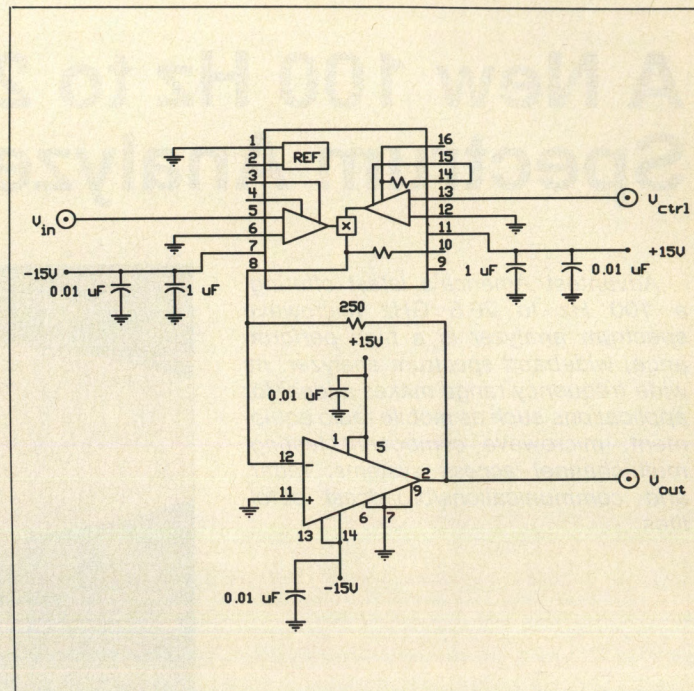
- Use good power supply bypass capacitors and connect them right at the power supply pins. A tantalum capacitor in parallel with a ceramic capacitor gives good bypass performance at both low and high frequencies.

In summary, current feedback amplifiers now give designers new freedom in high frequency designs. Compared to op amps, they provide wider bandwidth at higher gains, lower distortion, better large signal performance and faster settling time with relatively low power. External compensation, often required with high speed op amps, and other kinds of external optimizing adjustments are eliminated as well. The next generation of current feedback amplifiers are certain to provide even higher levels of performance for analog signal processing applications approaching one gigahertz.


**RF**

#### About the Author

Al Little is an Applications Engineering Manager for the Signal Processing Products Division of Harris Semiconductor. He may be reached at (407) 724-3842.



**Figure 7.** Current feedback amplifier used as a voltage controlled amplifier or mixer.



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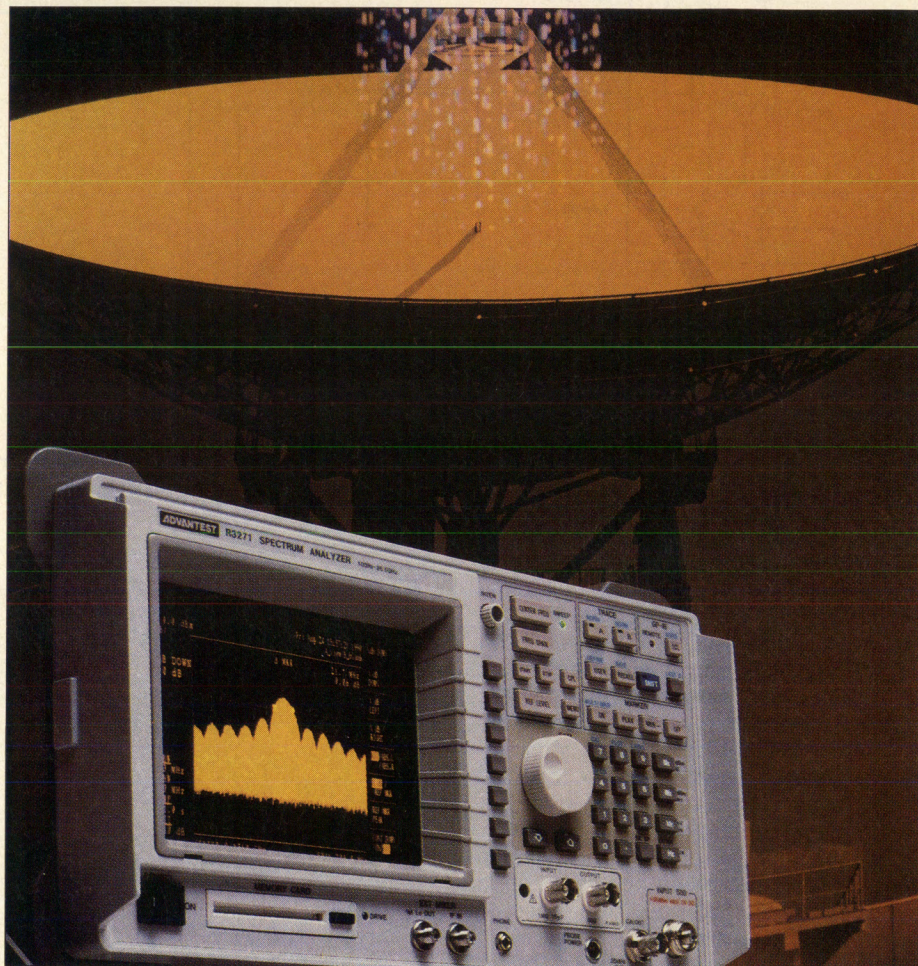
The R3271 Microwave Spectrum Analyzer makes use of advanced high frequency measurement techniques to cover its frequency range. The frequency range of the analyzer is internally preselected, providing more than 100 dB of true dynamic range and a displayable dynamic range of 100 dB. A synthesized local oscillator provides excellent frequency accuracy. In addition, its counter mode provides accurate frequency measurements of broadcast signals up to 26.5 GHz. Its high input sensitivity, coupled with its built-in counter, make it ideal for detecting individual signals in a complex waveform at extremely low-level signals. Precise measurements of modulated and complex signals are also possible. Key specifications are listed in Table 1.

Advantest's R3271 has a standard GPIB which provides for full remote control making it an excellent choice for use in automated measurement systems. It has a controller option for built-in computing as well as a memory card.

The R3271 measures 7 inches high by 14 inches wide by 18 inches deep and weighs less than 50 lbs. These dimensions and relatively light weight make it an ideal instrument for field applications.

The Advantest R3271 is priced at \$32,000 and will be available for delivery by March 1, 1991.

For more information on the R3271 spectrum analyzer from Advantest America, circle Info/Card #170.



**Advantest's R3271 Microwave Spectrum Analyzer.**

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**Table 1. Primary specifications for the R3271 Microwave Spectrum Analyzer.**

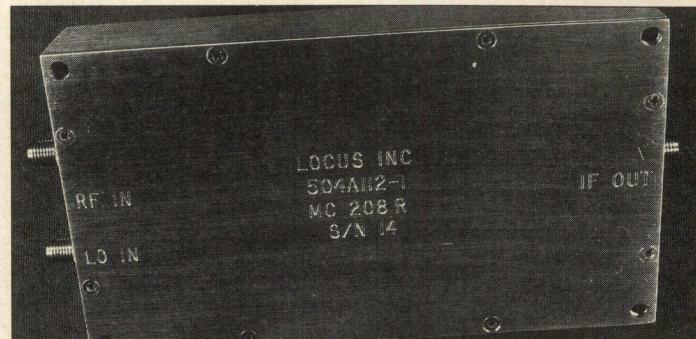


## Image Reject Mixers

Locus, Incorporated announces their new lines of image reject mixers. The lines of mixers include image reject down-converters, single-sideband up-converters, and I-Q modulators/detectors. There are twenty-three image reject down-converters with image rejection ranging from 25 to 55 dB, RF-IF conversion gain from -15 to +35 dB typically, and LO drive level ranging from 10 to 17 dBm. The single-sideband up-

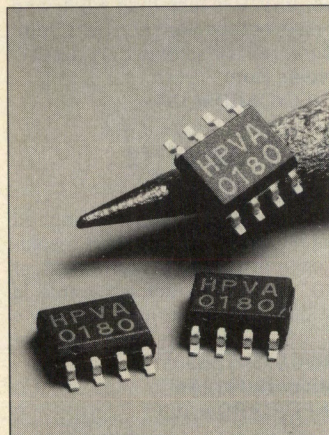
converters have image rejection ranging from 15 to 60 dB minimum, RF-IF conversion gain ranging from -15 to +54 dB typically, and carrier suppression ranging from 15 to 45 dB minimum. The I-Q modulators/detectors have phase balance ranging from  $\pm 0.1$  to  $\pm 5$  degrees, RF-IF conversion gain of -14 to +3.5 dB, and LO drive level of 10 to 13 dBm.

**Locus, Inc.**  
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## Monolithic Variable Gain Amplifier

The HPVA-0180 silicon-monolithic, variable gain amplifier from Hewlett-Packard features a 3 dB

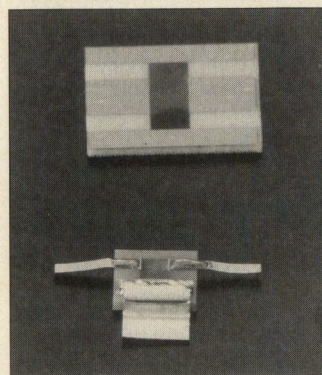


bandwidth of 2.5 GHz, provides 20 dB gain with 20 dB gain control over its DC to 2.5 GHz bandwidth, and dissipates only 250 mW from a single 6 volt power supply. It is packaged in a plastic, surface-mount SO-8 package and is suitable for wide or narrow bandwidth applications from DC to 2.5 GHz. The HPVA-0180 is designed for VHF/UHF receivers, RF data links, and broadband local area networks. Pricing for the HPVA-0180 is \$14 each in quantities of 1-99, \$13.25 each in quantities of 100-499, and \$12.25 each in quantities of 500-999. The unit is available from HP sales offices or distributors of HP component products.

**Hewlett-Packard Company**  
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## Surface Mount RF Attenuators

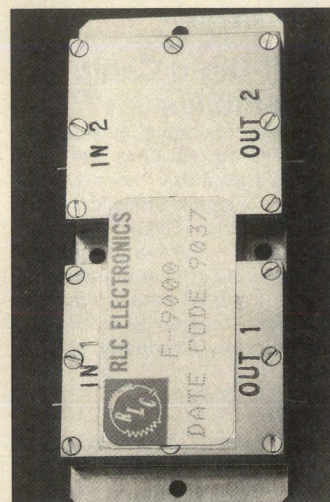
Hokuriku USA, Ltd. has released the PAD/PADU series of attenuators for both small signal and high power applications. The series is based on thin film ceramic substrates and utilizes a coaxial design. PAD types are available for applications from DC through 5 GHz, with PADU types rated at up to 100 watts for transmitting applications up to 4 GHz. Both types offer low-profile surface mounting, and impedance is 50 ohms. The PAD series can be specified for any attenuation between 1 and 32 dB, and the PADU series is available in 3, 6, 10, and 20 dB versions. The PAD series is sized at 8 mm x 18 mm, and the PADU series measures 21 mm x 11.6mm. Applications include communications equipment, CATV equipment, video systems, test instruments, and RF/microwave accessories.



These attenuators are available in OEM quantities with prices starting at less than \$8 each.  
**Hokuriku USA, Ltd.**  
**INFO/CARD #209**

## Matched Filter Sets

RLC Electronics has introduced a new line of matched filter sets. The filters are available with

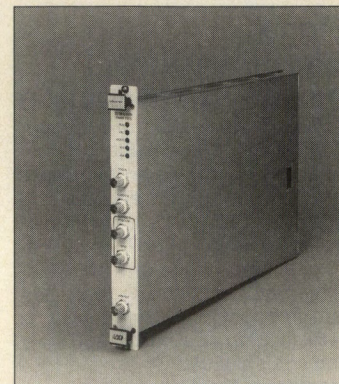


phase, amplitude, and delay matching. Typical requirements include phase matching at  $\pm 5$  degrees, group delay at  $\pm 2$  ns, or amplitude matching with a  $\pm 0.2$  dB variation over a portion of the filter's passband. The filter pictured is a surface mount device with two matched filters on a common carrier plate. The device is phase, delay, and amplitude matched over center frequency  $\pm 20$  MHz. Full Mil-Spec environmental requirements such as vibration, shock, and temperature are applicable. Pricing for the matched filter sets start at \$500 per set and are they available for delivery in small quantities in 2 to 4 weeks.

**RLC Electronics, Inc.**  
**INFO/CARD #208**

## VXI Arbitrary Waveform Generator

Wavetek has introduced Model 1375, a VXI arbitrary waveform generator (arb) compatible with the Standard Commands for Programmable Instruments (SCPI) remote programming format. The 20 MHz arb is a single slot C size VXI module that has 12 bit vertical resolution. It is equipped with 32K of volatile memory expandable to 128K. The module's automatic waveform scaling feature provides the ability to expand or contract a waveform into a larger or smaller memory space than originally occupied. A user can divide memory space into blocks of 8K and perform phase continuous switching between memory blocks upon receipt of a valid trigger signal either over the VXI backplane or externally. The price for Model 1375 is \$3995, and the expanded memory option costs \$1595.



**Wavetek Corporation**  
**INFO/CARD #207**



## UHF Backpack Booster Amplifier

Model M200U-BP contains a 200 Watt CW from 200-400 MHz (broadband) into 50 ohms, when driven by a 3-5 Watt exciter. It is linear and can be used for voice communications and CW. RF power is monitored from an LED bargraph meter, and automatic antenna T/R switching is also built-in. 50, 100, 150, and 200 Watt power output levels can be selected and power consumption is 28 volts and 18 amperes at 200 Watts output power. The unit weighs 19 lbs.

**Kalmus Engineering International, Ltd.**  
INFO/CARD #206

## Spectrum and Network Analyzer

The Spectrum & Network Analyzer FSBS from Rohde & Schwarz features a user-selectable frequency offset of its tracking generator and sensitivity of more



than -145 dBm over the range from 100 Hz to 5 GHz. It is designed for scalar frequency-conversion network analysis and spectrum analysis.

**Rohde & Schwarz**  
INFO/CARD #205

## Signal Microprocessor Development Station

The Signal MicroProcessor (SMP) Development Station contains the hardware and software needed to program and operate a 128 tap, 100 MHz programmable transversal filter for applications such as waveform equalization, matched filtering, and pattern recognition. The device is self-clocking at a rate of 360 MHz and costs \$8500.

**Electronic Decisions Incorporated**  
INFO/CARD #204

## Universal Counter

The CDC250 dual-channel counter will count signal frequency of sine, square, and triangle waves from 5 Hz to 175 MHz at input levels from 20 mV to 24 V peak. The instrument also provides period measurements, frequency ratio, time interval, and total measurement functions. It is priced at \$595.

**Tektronix**  
INFO/CARD #203

## Portable 9600 BPS RF Modem

UDS has announced the DR 96, a 9600 BPS radio frequency modem that uses the 470 MHz frequency band and has a sensitivity of 0.35  $\mu$ V. The DR 96 can transmit in both synchronous and asynchronous modes, and it has 10 ms RTS and CTS signaling time. Initial pricing is set at \$1,295.

**UDS**  
INFO/CARD #202

## Semi-Rigid Cable Assemblies

Semi-rigid cable assemblies from DC to 18 GHz are now available from the Phoenix Company of Chicago. Statistical Process Controls (SPC) are used throughout the assembly process to control all critical dimensions. SPC and VSWR data is available with each shipment at no charge.

**The Phoenix Company of Chicago**  
INFO/CARD #201

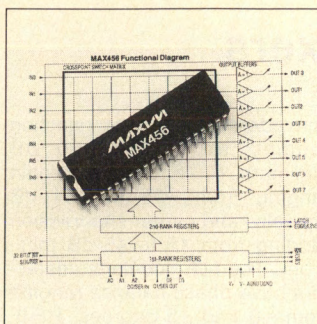
## Toroid Transformer Tester

Atlantic Magnetics has announced a toroid transformer tester for in-process testing and inspection of transformers and inductors wound on toroid cores. Model 3500A handles toroid transformers with inside diameters from 0.060" to 8.0". It completely tests transformers of 10 to 2000 turns in less than 15 seconds with an accuracy of  $\pm 0.2$  percent or better. Model 3500A is priced at \$1995.

**Atlantic Magnetics, Inc.**  
INFO/CARD #200

## 8 x 8 Video Cross-point Switch with Buffers

The MAX456 uses a digitally controlled 8 x 8 switch matrix to



connect eight high speed signals to any or all of the eight output channels. It has eight switch-matrix outputs that connect to eight 35 MHz, 250 V/ $\mu$ s video buffer amplifiers that can be disabled under digital control. Channel control logic is also included on chip. Single channel crosstalk is -70 dB at 5 MHz, and prices start at \$19.97 for quantities of 1000 and up.

**Maxim Integrated Products**  
INFO/CARD #199

## SPDT Switch

Model 62P004, an SPDT switch, offers 80 dB isolation from 410 MHz to 2300 MHz. It is TTL compatible and is available with SMA connectors. Video leakage is less than 10 mV in a 100 MHz bandwidth, insertion loss is 3.5 dB maximum, and VSWR is 2:1 maximum. It is priced at \$397 in quantities of 10-24.

**ECM Devices, Inc.**  
INFO/CARD #198

## High Speed 250 MHz Counter

The KL-5402D covers the 10 Hz to 250 MHz frequency range and measures 1ppm in 0.02 seconds. Counting accuracy is independent of input test frequency, and the unit is RS-232C compatible. Resolution is 1 ppm at 0.02 seconds gate time and improves to 0.0001 ppm at 200 seconds gate time.

**Kolinker Industrial Equipment, Ltd.**  
INFO/CARD #197

## RF Amplifier

Amplifonix Model TM6181 RF amplifier features 2.5 dB max. noise figure and 9 dB gain max. over the 10 to 400 MHz frequency range. Reverse isolation is -10 dB max. and VSWR is 2.0:1. Screening to the tables of Mil-Std-883 is available.

**Amplifonix**  
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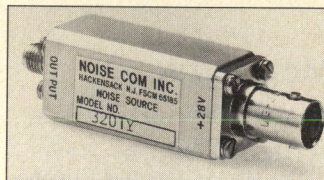
## 5 GHz SPDT Switch

Mini-Circuits' YSW-2-50DR SPDT switch has a built in driver and operates over the DC to 5 GHz range with 3 ns switching speed. It has greater than 40 dB isolation in the off state and less than 1 dB insertion loss in the on state. It is priced at \$19.95 in quantities of 1-9.

**Mini-Circuits**  
INFO/CARD #195

## Broadband Noise Source

The NC 3201Y is a broadband coaxial noise source that delivers white Gaussian noise from 10



kHz to 1100 MHz. Noise output rise and fall times are less than 1  $\mu$ s, VSWR is less than 1.2:1, and noise output variation with temperature is less than 0.01 dB/C.

**Noise Com**  
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## 8 Bit 100 MHz Flash ADC

The MN5901 is a high speed monolithic ADC with 8 bit resolution and a conversion speed of 100 MHz. Differential linearity is  $\pm 0.95$  LSB maximum and SNR is 38 dB minimum. Prices are from \$75 in quantities of 100.

**Micro Networks**  
INFO/CARD #193

## Alternating Voltage Measuring System

The model 4920 Alternating Voltage Measurement Standard from Datron Instruments measures signals from 1 Hz to 1.25 MHz with uncertainties of  $\pm 28$  ppm in stand-alone mode. It measures from 100 mV to 1000 V RMS and is priced at \$9,995.

**Datron Instruments Ltd., A Division of Wavetek Corporation**  
INFO/CARD #192

## Self-Biased Gain Block Amplifier

A GaAs monolithic gain block amplifier for use in the 2 to 10 GHz frequency range has been developed by Texas Instruments. It uses on-chip DC blocking which



allows it to be directly cascaded. Output power of the TGA8810 at 1 dB compression is typically 14 dBm and gain is 15.0 dB nominal at 5 V and 80 mA.

**Texas Instruments**  
INFO/CARD #191

## 2-Way Power Divider

The model PD032-2 is a 2-way power divider spanning from 300 MHz to 2000 MHz. Insertion loss is 0.8 dB, isolation is 18 dB, VSWR input is 1.35:1, and VSWR output is 1.3:1. Amplitude balance for the PD032-2 is 0.2 dB and phase balance is 3 degrees.

**Microwave Research and Development, Inc.**  
INFO/CARD #190

## DSP Array Processing Chip Set

Array Microsystems, Inc. has released a DSP chip set in a 144-pin grid array. The set includes the A66111, digital array signal processor and the A66211, a programmable array controller. The A66111BCG and

A66211BCG each sell for \$495.  
**Array Microsystems, Inc.**  
INFO/CARD #189

## 1000 MHz Frequency Synthesizer

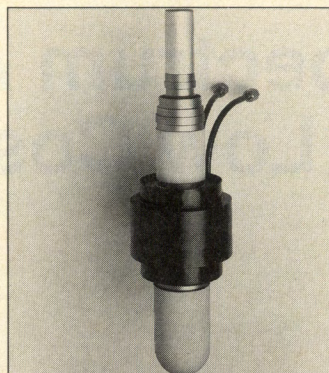
The PTS 1000 frequency synthesizer covers the 0.1 to 1000 MHz frequency range with resolution from 0.1 Hz to 100 kHz. Harmonics are at -30 dB at full output, phase noise is -60 dBc, and noise floor is at -130 dBc/Hz. The PTS 1000 is priced at \$11,500.

**Programmed Test Sources, Inc.**  
INFO/CARD #188

## High Power Magnetron

Burle Industries has introduced the C94604E 60 kW CW magnetron. The C94604E generates 60 kW of power at 915 MHz with an anode input of 16.1 kV and 4.4 amperes for a tube efficiency of 85 percent. Tube warranty of the C94604E is within two years of the tube's shipping date.

**Burle Industries, Inc.**  
INFO/CARD #187



## Ferrite-Free Divider

Sage Laboratories has released an 8-way reactive power divider which covers the range of 250 to 300 MHz. Typical performance over the band features insertion loss of less than 0.2 dB, input return loss of greater than 20 dB, amplitude unbalance of less than 0.1 dB and phase unbalance of less than 1.2 degrees.

**Sage Laboratories, Inc.**  
INFO/CARD #186

## Programmable Pulse Generators

The PSPL Model 10,000A features 40 volts amplitude and 400 ps risetime along with a GPIB, IEEE-488 interface. The amplitude can be adjusted down to 26 mV in 1/8 dB steps, and the pulse duration is adjustable from 1 ns to 100 ns in 25 ps steps. Negative polarity pulses to -40 V are also included.

**Picosecond Pulse Labs, Inc.**  
INFO/CARD #185

## Digitally Refreshed Spectrum Display

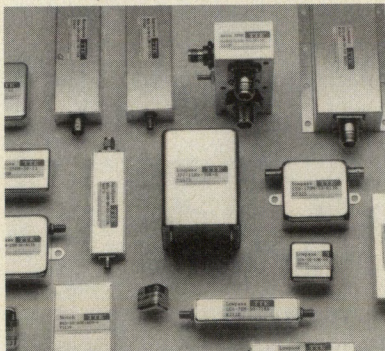
M/A-COM Government Products has released Model DRD-3572, a Digitally Refreshed Spectrum Display. The display has a wideband sweep of 40 MHz, and IF inputs at 70 MHz or 160 MHz are available. Other features include 70 dB dynamic range and two amplitude-calibrated adjustable markers.

**M/A-COM Government Products, Inc.**  
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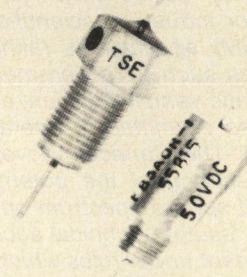
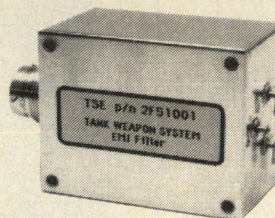
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# Spread Spectrum ASIC Eases Design of Low Cost Part 15 Systems

By Raymond W. Simpson  
O'Neill Communications, Inc.

Part 15 of the Federal Communications Commission's Rules governs the operation of RF communications devices without an individual license. Previously, Part 15 devices were limited to extremely low transmitted field strengths (low effective power), which reduced the ability to provide efficient, reliable communications in a mass marketed, no-license device. The power limits were necessary to assure that mass marketed, license-less devices would not cause harmful interference to vital communications services. Recognizing that spread spectrum techniques reduce the potential for harmful interference, in 1985 the FCC allowed the use of power output levels up to 1 watt in three frequency bands which are primarily used for industrial, scientific and medical (ISM) applications (although other services such as government systems, automatic vehicle location, and amateur are also permitted to operate in these bands). The Canadian government has also recognized the potential of unlicensed spread spectrum operation and is now issuing technical acceptances of equipment under rules which are similar (but not identical) to the FCC rules (1).

The FCC Rules (2) permit operation in the bands 902-928 MHz, 2.4 - 2.4835 GHz, and 5.725-5.850 GHz using direct sequence and frequency hopping spread spectrum (explained below). Other spreading schemes are not currently permitted under this section of the rules, although hybrid modes have just been added (3). There is no limit on the antenna gain which may be used (until 1994) (4), up to 1 watt power output may be used, and, except for the so called "forbidden bands" (5) the out-of-band emission suppression requirements are not stringent. Operation under Part 15

is on a "sufferance" basis, i.e. the Part 15 system must not cause harmful interference to licensed services and must accept any interference caused by other services. Thus, the only protection a Part 15 device can have from interference is in the cleverness of its design.

O'Neill Communications, Inc. has developed a wireless data communications product, the LAWN<sup>R</sup> for intercon-

necting personal computers and peripherals and a spread spectrum RF modem, shown in Figure 1. They both use an application specific integrated circuit, the OCI-100 which is being marketed separately for use in low cost Part 15 spread spectrum systems.

## Frequency Hopping and Direct Sequence

Frequency hopping (Figure 2a) is the

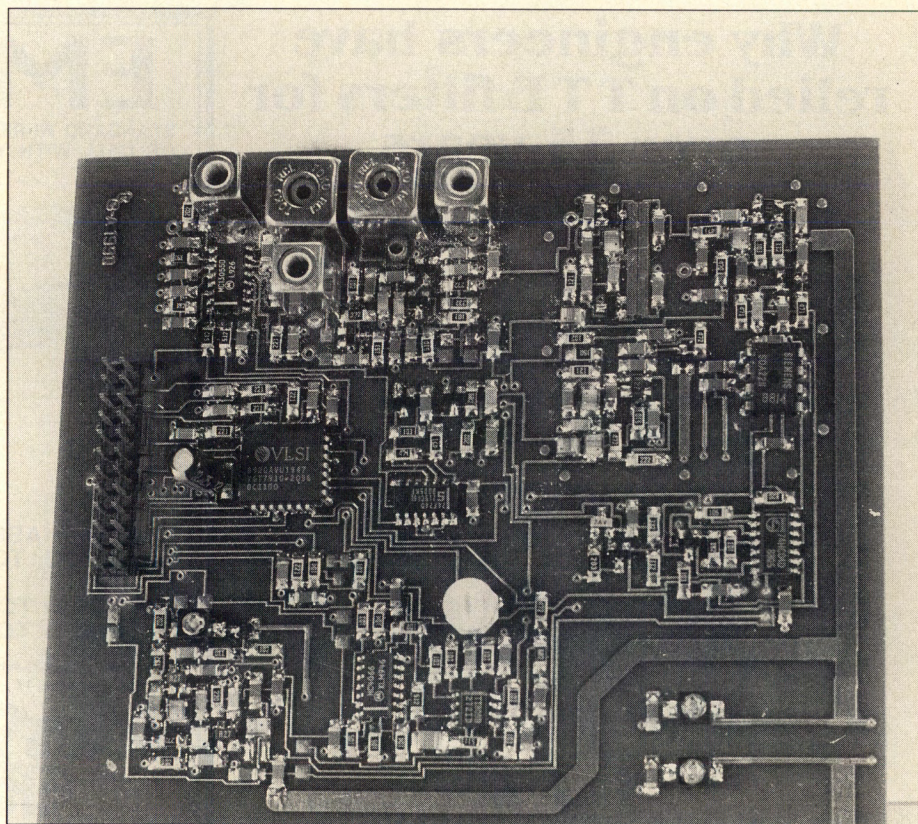


Figure 1. Wireless data communication product using a spread spectrum ASIC for interconnecting PCs, peripherals, and a spread spectrum modem.



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SAS-200/512	200 - 1800 MHz	Log Periodic	SAS-200/560	per MIL-STD-461	Loop - Emission
SAS-200/518	1000 - 18000 MHz	Log Periodic	SAS-200/561	per MIL-STD-461	Loop - Radiating
SAS-200/530	150 - 550 MHz	Broadband Dipole	BCP-200/510	20 Hz - 1 MHz	LF Current Probe
SAS-200/540	20 - 300 MHz	Biconical	BCP-200/511	100 KHz-100 MHz	HF/VHF Crnt. Probe
SAS-200/541	20 - 300 MHz	Bicon'l, Collapsible			

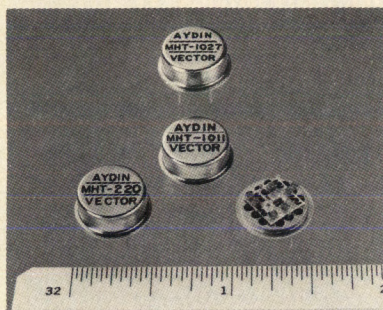
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easier to understand of the two FCC accepted spreading methods. A frequency hopping radio is just a conventional radio communication system which frequently (at least every 400 ms under the rules) changes its operating frequency, usually in a pseudorandom manner. The details of finding other units, maintaining synchronization in a multi-unit system, and making fast enough synthesizers, make the design of working, low cost systems a challenging undertaking. As the OCI-100 ASIC is not optimized for frequency hopping systems, they will not be discussed further here.

Direct sequence spread spectrum (DSSS) relies on combining a high rate (usually) binary sequence with the signal to be transmitted in such a way that the high rate sequence dominates the modulation bandwidth and directly determines the spread bandwidth (see Figure 2b). The high rate sequence is called a chipping sequence and the high rate bits are called "chips" to distinguish them from the information bits. This can be easily accomplished by adding the high rate sequence modulo-2 to the data to be transmitted, i.e. exclusive OR the two streams, as shown in Figure 3. An alternate way of looking at the process is to consider the chip sequence to be a series of plus or minus one elements, and view the modulo-2 addition as a multiplication (view the XOR as a digital balanced modulator). The two streams don't have to be synchronous, but simplifications in clock recovery can be achieved at the receiver if an integer number of chips are sent for each data bit, as in the OCI-100.

Most of the receiver signal processing in the direct sequence method can be done in the digital domain with relatively simple circuits, thus it is usually less costly to implement simple DSSS than

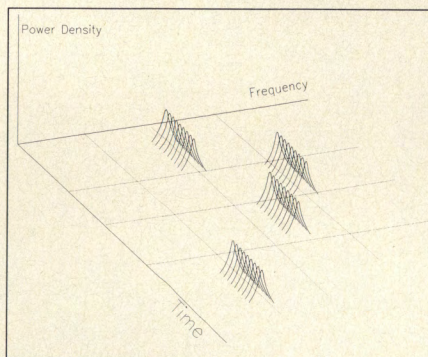
frequency hopping SS.

To receive a DSSS signal coded as above, it is possible to simply multiply the received chips with a locally generated, synchronized copy of the chip sequence. The multiplication can be done at baseband, IF or RF. The output of the multiplication is then low pass filtered (bandpass filtered if RF or IF processing is used) to provide the despread, filtered data. If a baseband implementation is used, the multiplication can be accomplished with an XOR gate.

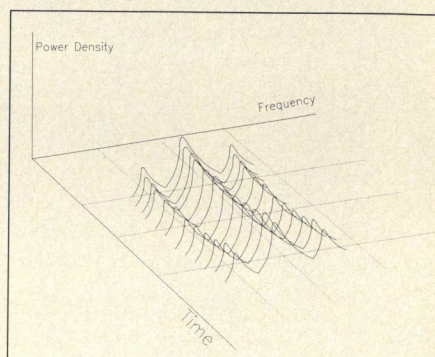
Synchronizing the local copy of the chip sequence can be accomplished by a variety of techniques (6) such as the delay locked loop. Delay locked loops require an acquisition time to attain synchronization, which can be a problem in asynchronous communication applications.

Another method of despreading a DSSS signal is to use a matched filter, especially a digital matched filter (7). The digital matched filter is just a shift register with taps at each stage. The output of the filter is the weighted sum of the outputs of the taps. Usually, binary weights suffice. The digital matched filter may be viewed as a digital approximation to an analog matched filter using a 1 bit analog to digital converter (i.e. a comparator).

The comparator should make the decision on each chip at the time at which the signal to noise at its input is maximum. This requires either recovery of the chip clock or operation at an oversampled rate. The latter approach was selected in the design of the OCI-100 as it allows a simpler implementation in CMOS and faster sequence acquisition. The received chips are sampled at 4 times the chip rate, and the correlation peak on each bit (16 chips, or 64 samples) is detected and used for

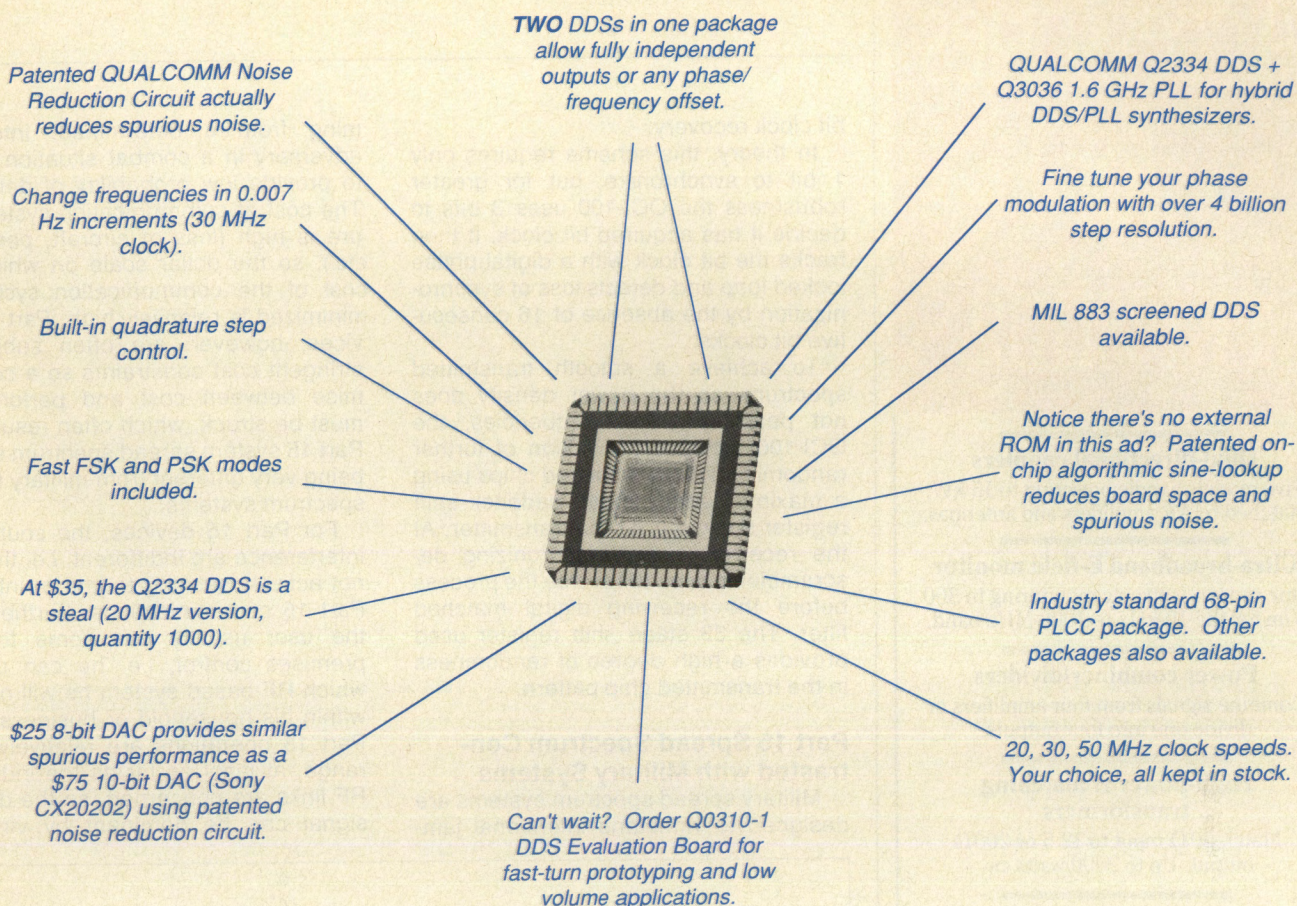


**Figure 2a. Graphical representation of frequency hopping spread spectrum.**



**Figure 2b. Graphical representation of direct sequence spread spectrum.**





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bit clock recovery.

In theory, this scheme requires only 1 bit to synchronize, but for greater robustness the OCI-100 uses 3 bits to decide it has acquired bit clock. It then tracks the bit clock with a digital phase locked loop and detects loss of synchronization by the absence of 16 consecutive bit clocks.

To achieve a smooth transmitted spectrum, so the power density does not peak at a few frequencies, the OCI-100 includes the option of further randomizing the transmitted chips using a maximal length linear feedback shift register scrambler at the transmitter. At the receiver, a self-synchronizing descrambler is used to reverse the process before the receiving digital matched filter. The 33 stage shift register used provides a high degree of randomness in the transmitted chip pattern.

## Part 15 Spread Spectrum Contrasted with Military Systems

Military spread spectrum systems are designed to withstand intentional jam-

ming from a determined, intelligent adversary in a combat situation and/or to provide low probability of detection. The cost of communication system failure is high (loss of aircraft, personnel etc.), so the dollar scale on which the cost of the communication system is minimized is relatively high. Part 15 devices, however, are often subject to stringent cost constraints so a compromise between cost and performance must be struck, which often results in a Part 15 system spread spectrum system being very different from military spread spectrum systems.

For Part 15 devices, the sources of interference are indifferent, i.e. they are not actively seeking to jam or intercept Part 15 communications. Furthermore, the user usually has some form of premises control, i.e. he can choose which RF based system he will operate within his household or business. Most Part 15 operations are relatively short range, at least as far as the individual RF links are concerned, so the desired signal can be dominant by virtue of

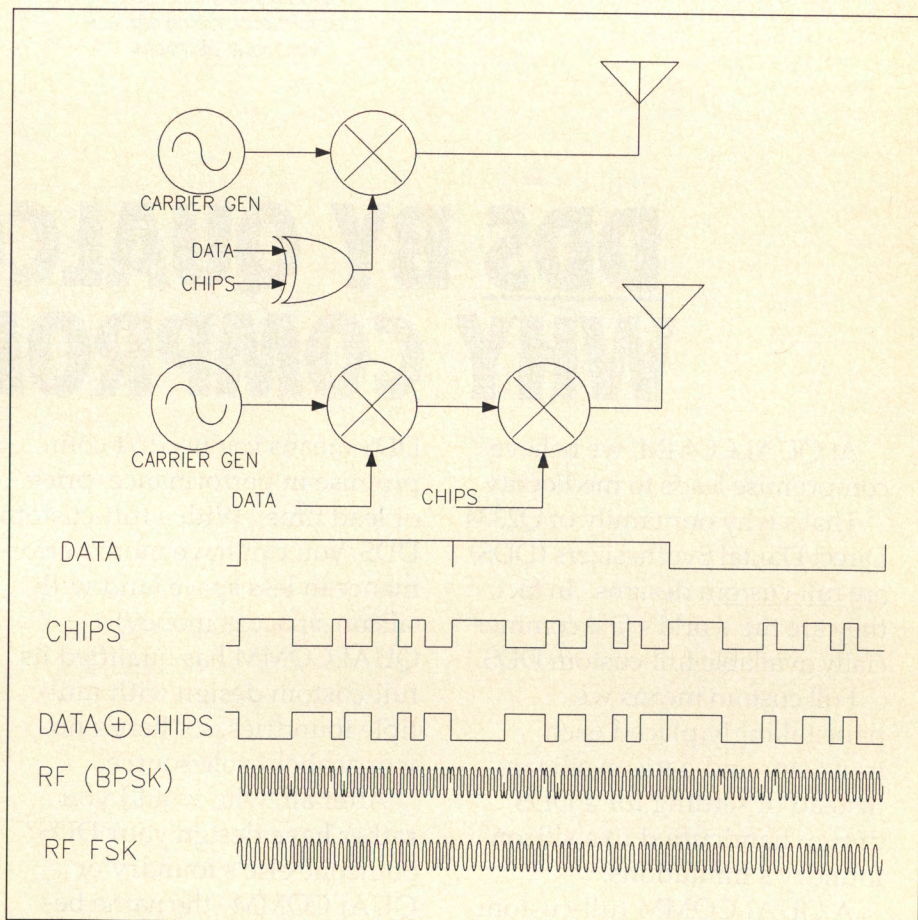


Figure 3. Distinguishing high rate sequence from information bits by using an exclusive OR.



A major consideration for Part 15 spread spectrum systems is the multipath environment, especially within buildings (8, 9). Within buildings, multipath propagation causes a pattern of nulls to be generated when the signal components arriving by the various paths arrive out of phase and nearly cancel. The spacing between nulls is related to the wavelength; in the 915 MHz band the nulls tend to be spaced within 6 to 12 inches of one another. A narrowband system has a non-negligible probability of being in a null for a given path. The wider bandwidth of a spread spectrum system effectively diffuses the nulls, reducing their depth. At OCI, we have seen that even relatively simple spread spectrum techniques can provide significant improvements in the reliability of indoor communication relative to narrowband techniques.

Unfortunately, there has been a tendency, in the less technical press, toward simplifications of the type "spread spectrum is interference free". In fact, given a time and bandwidth allocation, there are about

degrees of freedom in which to disperse a spread spectrum signal (6). Thus there is a limit to the number of fully separable signals which can occupy a given time bandwidth product, and beyond this there will be interference, even in an ideal implementation. The spreading process at the transmitter spreads the transmitted power, so the power density (W/Hz) is reduced. Thus there is less power in any single narrowband channel, but there is some transmitted power in more narrowband channels.

process or “despreads” the signal. The despreading process actually spreads signals which are not correlated to the spreading function, so that narrowband interferers are spread. The despreader is followed (at least equivalently) by a narrowband filter, which passes the despread signal but rejects the major portion of the uncorrelated interfering signal (which was spread rather than despread). The gain in signal to interference ratio under these conditions is often called the processing gain,  $G_p$ :

where  $B_{RF}$  is the spread RF bandwidth and  $B_I$  is the information bandwidth. In real systems, the inequality always holds. For “successful” communication to occur, the received desired signal power,  $P_d$  in dB, must obey

where  $P_i$  is the interfering power and  $J$  is the jamming margin, i. e. the required final signal to interference power required by the demodulator to operate at a "successful" level of performance.

have at most 30 dB of processing gain. Such a system would occupy the entire 902-928 MHz band; if it caused or received harmful interference to or from a licensed service, there would be little the operator of the Part 15 system could do to resolve the problem, except to cease operating. It may not be in the best interests of the system designer, therefore, to rely only on processing gain to prevent interference. If the example system were instead designed with 6 dB lower processing gain, it could offer the user a choice of 4 sub-bands in which to operate, providing an additional dimension of flexibility for solving interference problems.

A low cost spread spectrum system needs a low cost radio as well as low cost processing. Most of the elements of low cost 915 MHz radios are readily available, if a suitable radio architecture is used. A modulation method and frequency control scheme which help control cost should be selected.

FSK helps in low cost realizations.





because complete IF/demodulator systems are available in low cost single chip form, such as the Motorola MC13055 and MC3356 (10). At 16 chips per bit, the FSK approach tends to produce a more uniform spectrum than BPSK. Sidelobe control can be easily obtained by low pass filtering the chip stream between the OCI-100 output and the modulator. This results in a compact, smooth

spectrum with most of the transmitted energy in the main lobe. An FSK transmitter can be as simple as a VCO, as shown in Figure 4.

The performance of FSK using non-coherent detection and post-detection despreading is compared with non-coherent detection following ideal despreading, both using 16 chips per bit, in Figures 5 and 6. Figure 5 shows the

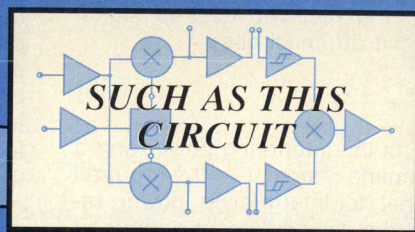
bit error rate as a function of carrier to noise ratio (in terms of bit energy  $E_b$  and the noise power spectral density,  $N_0$ ) with a constant signal to jammer ratio of 3 dB. Figure 6 compares the probability of successful packet completion, assuming 256 byte packets with no forward error correction. The figures show that in most practical cases (i.e. where a probability of packet success must be better than 90 percent) the difference in jammer rejection is less than 3 dB. It must be considered that the performance of a real system using pre-detection despreading will be degraded from the performance of the ideal despreader by the error (phase noise) in its synchronization of its local chip sequence with the chip sequence of the received signal.

The penalty for using post-detection despreading with FSK in additive white Gaussian noise is about 3 dB relative to ideal despreading. This is also an acceptable tradeoff for most Part 15 applications.

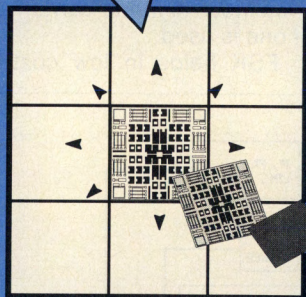
Frequency control at low cost is usually a problem if phase locked loop (PLL) ICs must be added to the design. Most of the standard PLL ICs are designed for the land mobile radio market, and are optimized for closer channel spacings than are needed for spread spectrum systems. Achieving the close channel spacing requires the use of a low reference frequency, with consequent longer loop lock up time. The OCI-100 contains two internal phase locked loops which, with external prescalers, can be used to control the receiver local oscillator and transmitter frequencies.

The receiver PLL uses a single modulus divide by 64 prescaler, such as the

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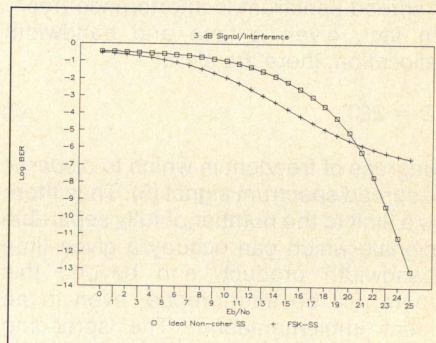


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**Figure 5. Bit error rate vs. SNR for comparing FSK (using non-coherent detection and post-detection despreading) and non-coherent detection following ideal despreading.**



Siemens SDA 2211 or similar. The single modulus prescaler saves cost relative to a dual modulus prescaler. The subsequent dividers and a tri-state frequency/phase detector are contained in the OCI-100. (There is a provision for using external phase detectors with the internal counters of the OCI-100 if desired.) The tri-state phase detector may be used with either passive or active loop filter, depending on the system requirements.

It is usually desirable for the transmitter to have a minimum turn on delay, because the turn on time represents an overhead during which there is no data being transmitted. In carrier sense multiple access (CSMA) packet radio systems, long turn on time increases the probability of collisions. To reduce the turn on time as well as cost, a mixing type of transmitter control loop is used in the OCI-100 design. The transmitter VCO is mixed with the receiver local oscillator (if the two circuits are on one board, it's almost inevitable that this mixing will occur, so you might as well make use of it as fight it). Mixing translates the transmitter frequency down to the radio's intermediate frequency without frequency division.

Depending on the chosen intermediate frequency, the difference frequency may be used directly or divided externally by 2 or 4 before being applied to the OCI-100. The result is a PLL with a much higher sampling frequency, therefore much quicker frequency acquisition than would result if the transmitter VCO frequency was divided down to the reference frequency. Yet, since the transmitter is referenced to the receiver local oscillator, it can have the same tuning resolution as the receiver. This approach does require the receiver local oscillator to be stable enough not to lose lock while the transmitter is keying up, but this has not proved to be much of a problem in practice.

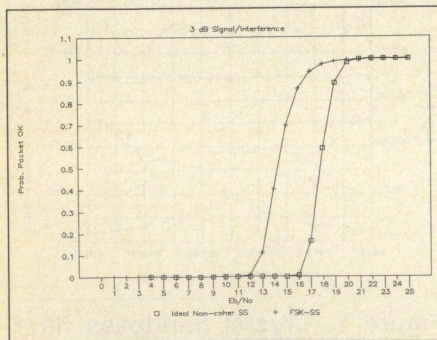


Figure 6. Probability of success.

Operation of the OCI-100 at 50 kb/s with a peak frequency deviation of 200 kHz will result in a 6 dB bandwidth of 640 kHz, comfortably above the FCC minimum of 500 kHz. Operation at rates up to 125 kb/s is supported by the OCI-100.

### Conclusion

Direct sequence spread spectrum communications devices can be simple enough to be manufactured at low cost, if small reductions in performance relative to "classical spread spectrum architectures" are acceptable. The radio section needn't be very different from the usual FM radio and the spreading/despreading functions can be handled by a low cost integrated circuit. The resulting package can be low in cost, small and operate at low power consumption.

RF

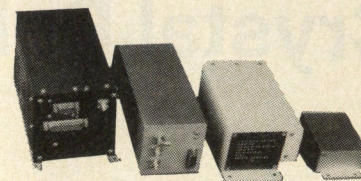
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### About the Author

Raymond W. Simpson is the Vice President, Systems and Technology at O'Neill Communications, Inc. He may be reached at 100 Thanet Circle, Suite 202, Princeton, NJ 08540. Tel: (609) 924-1095. Fax: (609) 924-5319.

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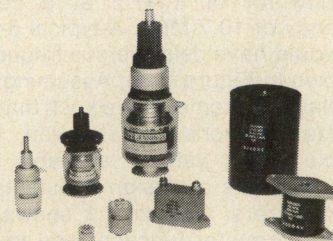
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# Crystal Delay Equalizers

By William B. Lurie  
Consultant

In a previous article (August, 1989), a method of providing envelope delay equalization for crystal bandpass filters was discussed, ending with the design of a single equalizer section in semi-lattice form, containing two crystals. More often than not, the amount of equalization needed necessitates the use of more than one "section" of equalizer. In general, these are, over the narrow band of interest, constant-impedance devices, and they can be cascaded with little interaction.

In practice, as was seen in Figure 7 of my previous article (1), there is a need to provide additional inductors, beyond what the basic filter and equalizer designs actually require. Inductors, in general, are the least desirable components, as compared with crystals and capacitors, because of their size, dissipation factor, and their temperature characteristics. A method has been devised, therefore, for combining a number of equalizers into a single section, putting multiple crystals in parallel in each of the two branches of the half-lattice, thereby minimizing the number of added inductors.

As an example, consider a filter with a bandwidth of 8 kHz at a center frequency of 10.7 MHz. A typical 6-pole filter would have delay versus frequency as shown in Figure 1 (A). Assuming that the delay requirements are such that two equalizer sections are required, it is possible using various available programs to arrive at design constants for the equalizers. Equation 1 shows the equalizer transfer function, and in Figure 1, curves B, C, and D show the delays of equalizer sections alone and together. Figure 1 also shows the composite delay of the filter with both equalizers added (curves E and F, where the scale for curve F is the axis to the right, expanded 5 times compared with E). It is readily seen that the delay variation without equalizers is 105 microse-  
conds across the 3 dB passband, but only 18 microse-  
conds with the equalizers added.

What is sought next is a single all-pass lattice which has the same transfer function as:

$$s_{12} = \frac{(p - p_1)(p - p_2)(p - p_3)(p - p_4)}{(p + p_1)(p + p_2)(p + p_3)(p + p_4)} \quad (1)$$

where

$$\begin{aligned} p_1 &= -2253 + j10698420 \\ p_2 &= -2253 - j10698420 \\ p_3 &= -2252 + j10701580 \\ p_4 &= -2252 - j10701580 \end{aligned}$$

all in Hertz

Each single equalizer section has two frequencies at which the transfer phase is a multiple of 180 degrees (F1 and F2 in Figure 2a), and the combination of two such equalizers, in cascade or combined, must have four such frequencies (F3 through F6 in Figure 2b). The location of these frequencies is not immediately obvious, but a computer program has been written to accomplish this.

A straightforward analysis of the phase of the transfer functions of the two single equalizers, added together, gives the phase versus frequency of the combined equalizer. It is a simple task to examine the phase curve or table, visually or by computer, to find the frequencies at which the phase becomes all multiples of 90 degrees. In the example, these were found to be as shown in the following table.

PHASE	FREQUENCY(Hz)
90	10694194
180	10697250
270	10698698
360	10700001
450	10701305
540	10702753
630	10705810

Now, by a process of mathematical induction, the schematic of the combined half-lattice is as shown in Figure 3, with all element values listed. Note

that the shunt capacitance in the "A" branch has been arbitrarily set at 4 pF, and in the "B" branch, -4 pF. These values, knowing the pole and zero locations of the branch reactances, uniquely determine all the element values, simply by using a partial-fraction expansion (program available, of course). The scaling of impedance level to match the filter itself is trivial.

As discussed for the single-section equalizer, the negative capacitance is swamped and made positive by the "borrowing" of capacitance from sources external to the lattice (1). The process of combining equalization properties of several sections into one section does serve to lower the number of inductors and capacitors as well, but it is not entirely without penalty. Although quartz crystals have stability and Q far better than coils, capacitors, or combinations of both, they suffer nevertheless from certain practical limitations. Their motional inductance or capacitance can only be obtained in a rather narrow range of values. The setting tolerance, at 10.7 MHz, is of the order of  $\pm 100$  Hz, and the motional parameters, in practical manufacturing terms, can only be held to about 2 percent. None of these factors is any more serious for the combined equalizer than for the individual sections, but it is more difficult, with fewer adjustments possible, to tune up or align a more complex section than two simpler ones. It is beyond the scope

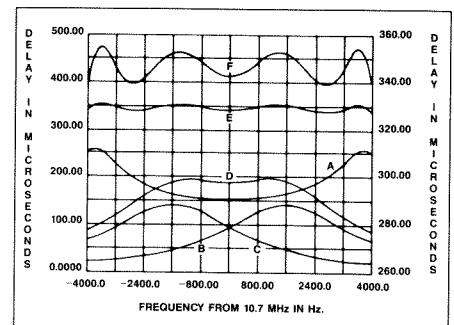


Figure 1. Crystal bandpass filter with and without equalizers.



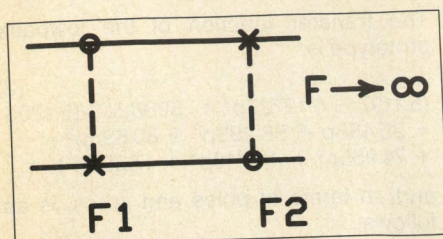


Figure 2a. Branch reactances of single delay equalizer.

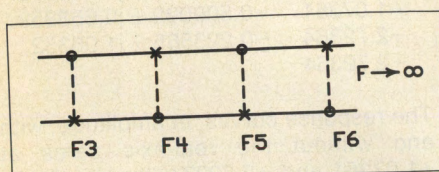
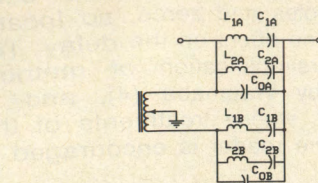


Figure 2b. Branch reactances of combined equalizers.

of this article to go into more details, but it is surely possible to program a computer to perform a tolerance study and predict yields and performance, using statistical data and assumptions regarding component variations. In addition, there are secondary methods of changing the equivalent crystal frequency and motional capacitance using additional tuning capacitors external to the crystals.

This series of articles has, so far, discussed some of the aspects of the creation of band-pass devices with envelope delay made more constant by the addition of one or more constant-impedance all-pass networks in cascade. In many respects, this technique has distinct drawbacks, as pointed out by Watanabe (4). It does seem less than optimum to design for amplitude first, create a certain amount of delay non-uniformity, and then correct the delay back toward uniformity at substantial cost. It should be pointed out, however,



Crystal 1A	0.0402249055 H	5.50616365 fF
Crystal 2A	0.179071127 H	1.23521117 fF
C <sub>0A</sub>	4 pf	
Crystal 1B	0.178985359 H	1.23640542 fF
Crystal 2B	0.0401877843 H	5.49929652 fF
C <sub>0B</sub>	-4 pf	

Figure 3. Combined equalizers with element values.

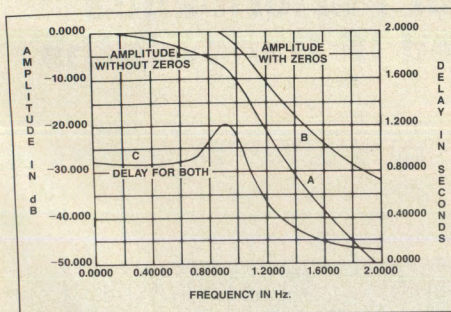


Figure 4. Rhodes N=6 low pass filter.

that the state of the art, as it progressed, allowed no other solution. More modern mathematical and network theory methods have presented several viable alternatives.

In modern network methods, the transfer function of the system (or filter) is created, such that the system's needs are met, and then a network is synthesized by one or another technique, to have that transfer function. In terms of classical filter-plus-equalizer methodology, one might say that the transfer function is created by pay-

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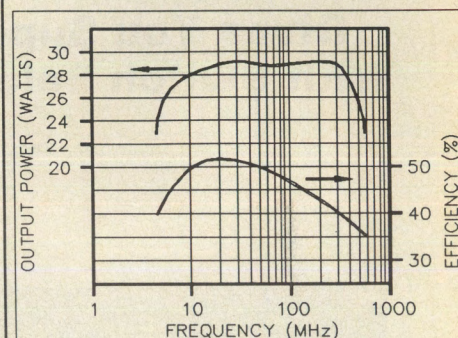
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ing attention only to amplitude, and then an all-pass transfer function is added to correct the phase (or delay), and then the two parts are synthesized and combined.

One alternative approach would be to design a transfer function loosely based on consideration only of the delay, which implies that the passband amplitude corners might be severely rounded, rolling off too soon. Then the

amplitude could be corrected using corrector poles and zeros, so located as to have no effect on the delay. This is an oversimplification of methods proposed by Watanabe (4), and by Rhodes (3), whose treatments of this approach the reader is encouraged to study.

As an illustration of this technique, consider a bandpass filter described by Herzig and Swanson (2).

The transfer function of the lowpass prototype is:

$$(5.1197 - 5.12723p^2 + .59882p^4)/(5.1203 + 25.455p + 58.203p^2 + 80.695p^3 + 74.955p^4 + 44.743p^5 + 17.202p^6) \quad (2)$$

and, in terms of poles and zeros, is as follows:

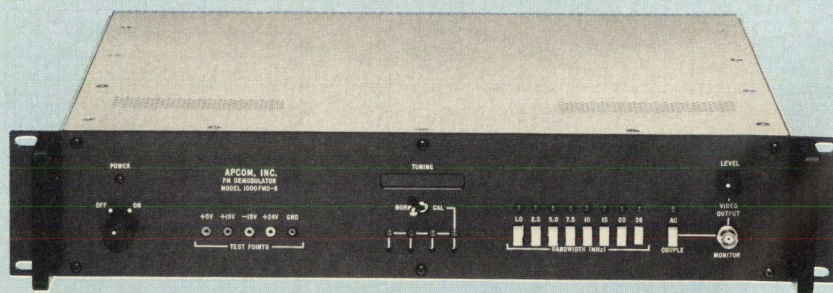
#### ZEROS

+1.07351  
-1.07351  
+2.72364  
-2.72364

#### POLES

-0.567877 ± j0.202613  
-0.509092 ± j0.658859  
-0.223565 ± j1.06352

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The response curves, in amplitude, with and without the real-axis zeros at ±1.07351 and ±2.72374 (the amplitude correctors) are curves A and B of Figure 4. Correspondingly, the delay curve of Figure 4 is applicable to both cases, since the two pairs of zeros, being symmetrical about the jω axis, have no effect on the delay. The bandpass filter was designed from this low pass prototype, based on a polynomial derived from Rhodes' articles, and an entire class of lowpass filters can be (and has been) created from these, as will be discussed in a separate article. In many respects, this class of filters is quite interesting, since, for any degree, a parameter can be selected by the designer, affecting the rate of cutoff in the stopband with relatively minor effect on the passband delay, and even yielding frequencies of infinite attenuation similar to the Elliptic filters.

RF

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#### About the Author

William Lurie has worked as a mathematician, physicist and electronics engineer in a variety of fields, including magnetic compasses, X-ray tubes and measuring equipment, airborne Doppler radar, and filters. He holds Life Senior Member status in the IEEE. He works as an independent consultant and can be reached at 8503 Heather Place, Boynton Beach, FL 33437. Tel.: (407) 369-3218.



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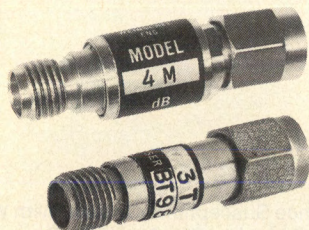
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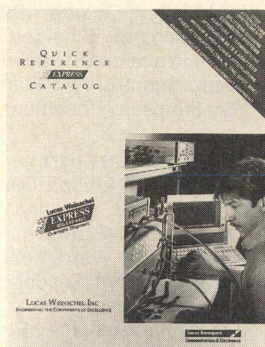


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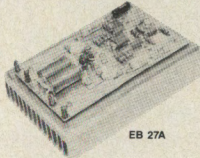
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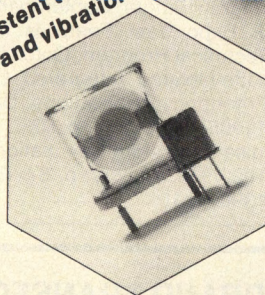
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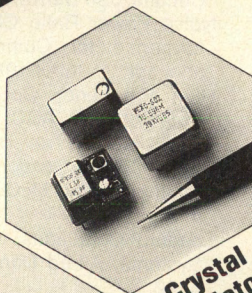
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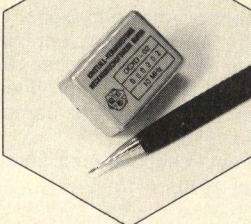
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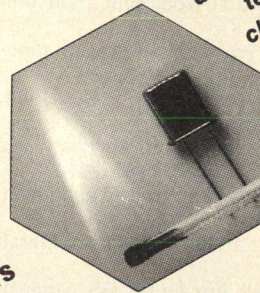
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ASP is a functional transistor amplifier simulation program developed by SW.I.F.T. Enterprises that includes design routines, performance data, matching circuits, auto 'Q', and a NF optimizing utility. ASP Version 3.0 operates on greater than DOS 2.11 and requires an EGA or better monitor. A math coprocessor is recommended and the price is \$75.

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## Schematic Entry System

Vutrax-II GES is an ECA-2/LCA-1 compatible schematic entry system available through Tatum Labs. It can be used as a stand-alone drawing package or as the drawing interface to circuit analysis. Vutrax-II GES runs on IBM or compatible computers with Hercules, CGA, EGA, or VGA monitors. It requires a hard disk with 4M available. The price is \$495 and it will interface with dot matrix and laser printers.

**Tatum Labs, Inc.**  
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## Optimizing Circuit Simulator

Meta-Software has released an enhanced

version of its optimizing analog circuit simulator, HSPICE™. Version H9007 features an enhanced transmission line model with an improved field solver for board/hybrid and LSI applications, a lossy first order skin effect model, and S-parameter calculations. It is available on Sun, MIPS, DEC, and Apollo workstations.

**Meta-Software, Inc.**  
INFO/CARD #216

## DSP Code Generation Software Upgrade

Burr-Brown has released an upgraded version of DSPlay XL™. The software allows users to quickly develop DSP code for AT&T's WER DSP32 and DSP32C processors through the creation of simple block diagrams. The upgrade, 3.16, expands the available DSP and related functions to over 100 from the program's original 60. DSPlay XL is priced at \$1495.

**Burr-Brown Corporation**  
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## Waveform Creation and Modification Interface

The SM 5000 WaveCAD software from John Fluke Mfg. provides a mouse-driven interface for the creation or modification of

waveforms. It allows users to sketch a waveform freehand or point-to-point. SM 5000 WaveCAD software is priced at \$1095 and is available on a four week delivery schedule.

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## IC Design Libraries

EEsof's new SMART (Simulatable Microwave ARTwork) libraries are used to translate a design developed from schematic circuit simulation into final physical layout. Each SMART library includes schematic symbols, simulation models, and layout artwork representations most often used in MMIC design. The libraries are priced at \$5,000 each.

**EEsof, Inc.**  
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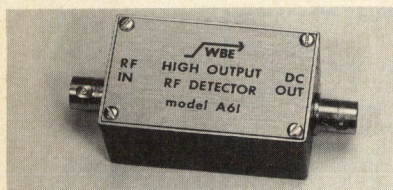
## Microstrip Analysis Software

Sonnet Software introduces version 2.1 of Em, an electromagnetic analysis package. This software evaluates any discontinuity, validates large portions of entire designs, and enhances existing design software. Em offers a matrix solution speed typically six times faster than version 2.0.

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"Circuits for Wideband FM Demodulation," by Fotowat and Wong of Signetics. Program computes distortion of discriminator circuit. Includes HELP file and six examples. (FORTRAN source code listing and compiled, executable version)

### Disk RFD-1190: November 1990

"Broadband Impedance Matching by Polynomial Synthesis" by David Lang. Generates a polynomial equalizer function directly from S-parameters, and synthesizes matching networks. (BASIC, source listing and compiled versions)

### Disk RFD-1090: October 1990

"Microstrip CAD Program," by Thomas Cefalo of MITRE Corp. Computes microstrip impedance, delay, inductance, capacitance, and other factors. (BASIC, compiled)

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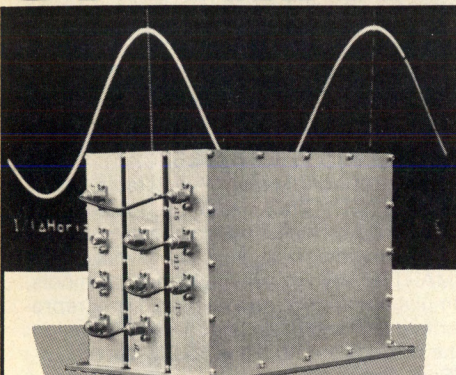
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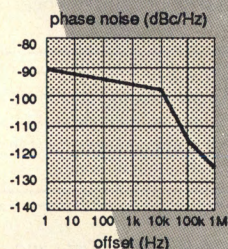


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### GaAs FETs and MMICs

Harris Microwave has released a data book titled GaAs FETs, MMICs & Foundry Service. The data book gives specifications and qualifications of Harris Microwave's GaAs FETs and MMICs and also includes application notes on these items.

**Harris Microwave Corporation**  
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### DSP Products

Burr-Brown has announced the availability of a brochure that describes the company's complete line of digital signal processing products. Described are software, DSP processors for both PC and VME platforms, and analog I/O boards and systems.

**Burr-Brown Corporation**  
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### Quartz Products Catalog

Tele Quarz has introduced a short form catalog covering their quartz crystals, oscillators, filters, and discriminators. The catalog contains crystal units between 1 and 300 MHz. Fundamental crystals up to 60 MHz and 3rd through 9th overtone crystals up to 300 MHz are described.

**Tele Quarz Group**  
INFO/CARD #227

### Mixers and Frequency Doublers Application Note

An application note which discusses mixers and frequency doublers has been published by FEI Microwave. The note includes a section on mixer terminology and mixer applications and a section on frequency doublers and their applications. Also included are product summaries of FEI Microwave's mixers and frequency doublers.

**FEI Microwave, Inc.**  
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### IF/RF Solid State Devices

Daico Industries has announced its 1991 catalog, which features Daico's complete line of GaAs, PIN diode, Schottky diode, and relay devices. Key features, typical performance, operating characteristics, and block diagrams are included.

**Daico Industries, Inc.**  
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### Optics Control Techniques

A booklet providing information on five optics control techniques has been released by Anritsu. The booklet describes high-speed modulation of lightwaves, control of optical waves, optical effects, development of a hyper-coherent optical sweep generator,

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phase-conjugate optics, and semiconductor quantum effect devices.

**Anritsu America, Inc.**  
INFO/CARD #225

### Quartz Crystal Oscillators

Piezo Crystal has announced their new product brochure covering quartz crystal oscillators for commercial, military, and GPS industries. It includes technical information on part numbers 2900082 and 2890080, Piezo's newest crystal oscillator products.

**Piezo Crystal Company**  
INFO/CARD #224

### Filter Catalog

Catalog Fastrap/90 from Microwave Filters Company defines and explains positive and negative trapping in cable systems, suggests strategic rules and illustrates a wide range of channel notch filters and tiering filters.

**Microwave Filter Company, Inc.**  
INFO/CARD #223

### Surface Mounted Components Catalog

Murata Erie North America has released a catalog that includes specifications and application information on surface mount capacitors, potentiometers, inductors, EMI/RFI fil-

ters, and ceramic filters and resonators.

**Murata Erie North America**  
INFO/CARD #222

### Phase Locked Sources

This catalog covers the XDMP series phase locked sources which utilize an internal crystal oscillator in the 100 MHz region which is locked to an external 5 or 10 MHz reference. These sources are for applications where high spectral purity and extremely low FM noise characteristics must be determined by a low noise crystal oscillator.

**Communication Techniques, Inc.**  
INFO/CARD #221

### Materials Measurement Package

A four page data sheet from Wiltron describes their materials measurement package which measures permittivity and permeability of a solid, powder, or liquid material in coaxial or waveguide holders at microwave frequencies. Also presented is an overview of the measurement method, sample preparation, and transmission line fixturing.

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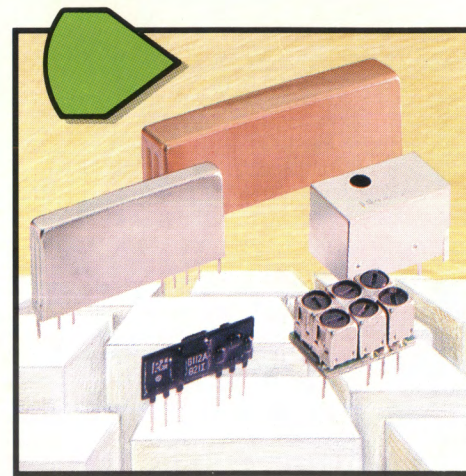
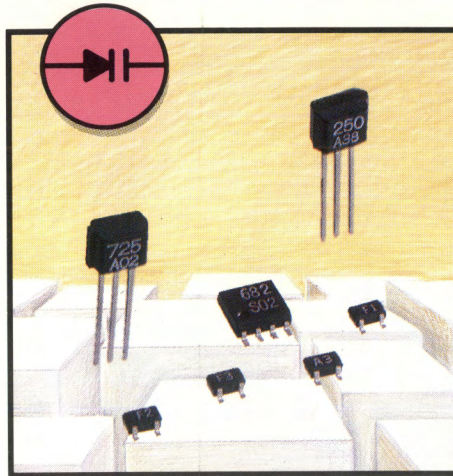
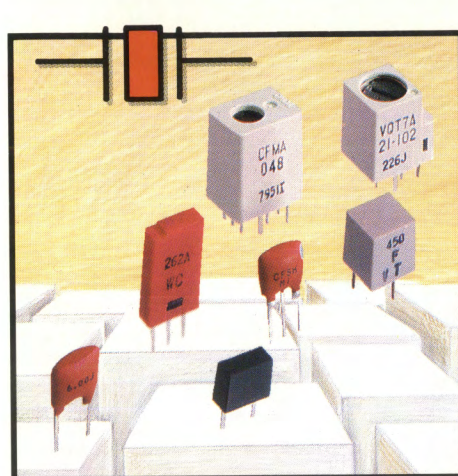
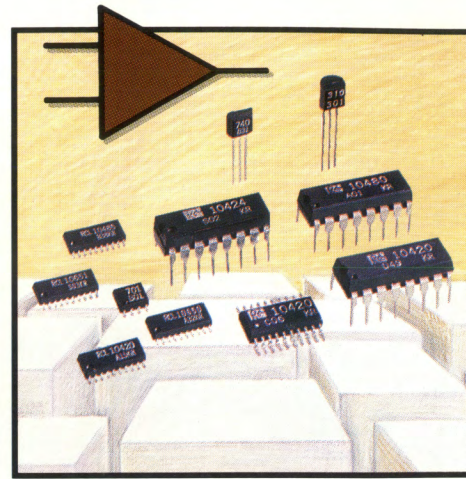
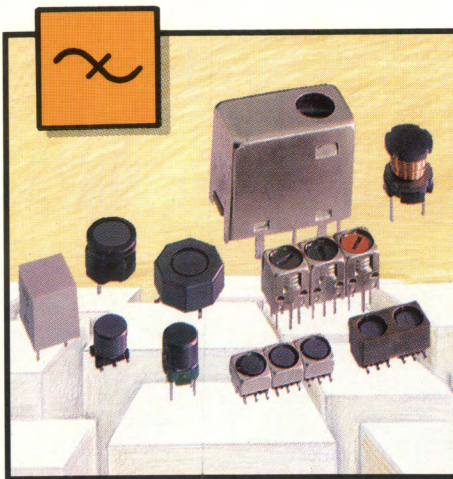
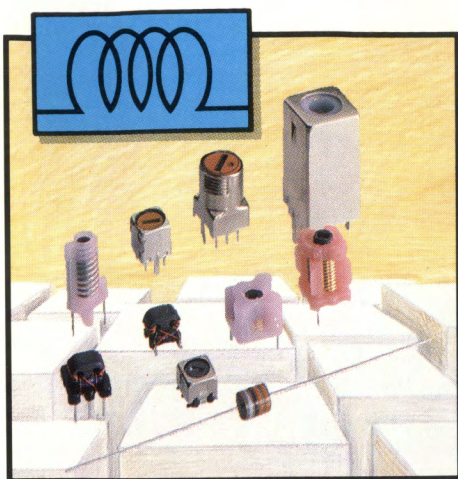
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INFO/CARD 63





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For I-F amplification and detection, Toko offers a variety of ICs to simplify design and reduce product size. Other available

types include: companders for noise reduction, equalizers, analog switches and voltage regulators.



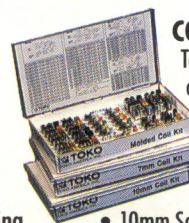
For I-F or oscillator circuits, Toko ceramic filters and oscillating elements are an alternative to more costly technology. These High Q devices are also extremely compact and stable.



With over 36 stock types in a wide range of capacitance values, Toko has the varactor diodes you need, in SMT or SIP packages. An exclusive manufacturing process eliminates tracking errors in matched sets.



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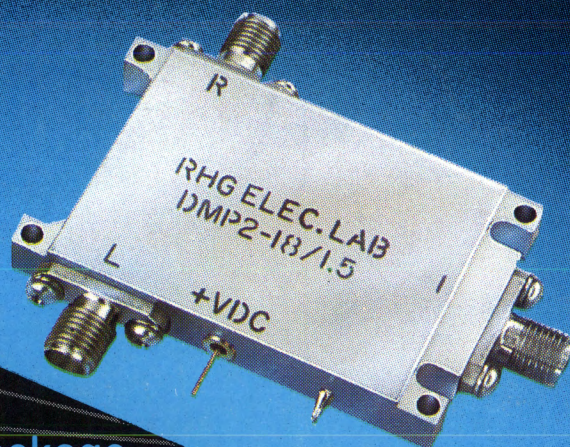
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- Hermetic drop-in package
- 2-18GHz RF/LO, 1.5GHz IF
- Low noise, high  $\frac{1}{2}$  IF rejection

## OPERATING SPECIFICATIONS, MODEL DMP2-18:

### INPUTS:

Frequency, RF/LO (GHz)	2-18
VSWR, RF/LO	2.5:1/2.0:1
Peak power (1 $\mu$ sec pulse) (W)	150
Power, LO (dBm)	+13
Power, DC (Volts/mA)	+12/75

### OUTPUTS:

Frequency, IF (GHz)	1.2-1.8
VSWR, IF	2.5:1
Intermodulation (dBm):	
2nd/3rd order	+20/+30
Power, IF @ 1 dB comp. (dBm)	+12

### TRANSFER CHARACTERISTICS:

Conversion gain, min. (dB)	15.5
Isolation, min. (dB)	25
Noise figure (dB)	10
Single tone intermodulation with -10 dBm RF input (dBc):	
1LO-2RF	50
1LO-3RF	60
2LO-1RF	25
2LO-2RF	50
2LO-3RF	60
3LO-2RF	65
3LO-3RF	70

**ENVIRONMENTAL:** Operating temperature -55°C to +85°C; hermetically sealed package;  
meets applicable requirements of MIL-E-5400.

**RHG ELECTRONICS LABORATORY, INC.**

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