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engineering principles and practices

1990 Directory Issue

Products

Design Guide

Company Index

1200+ RF Companies

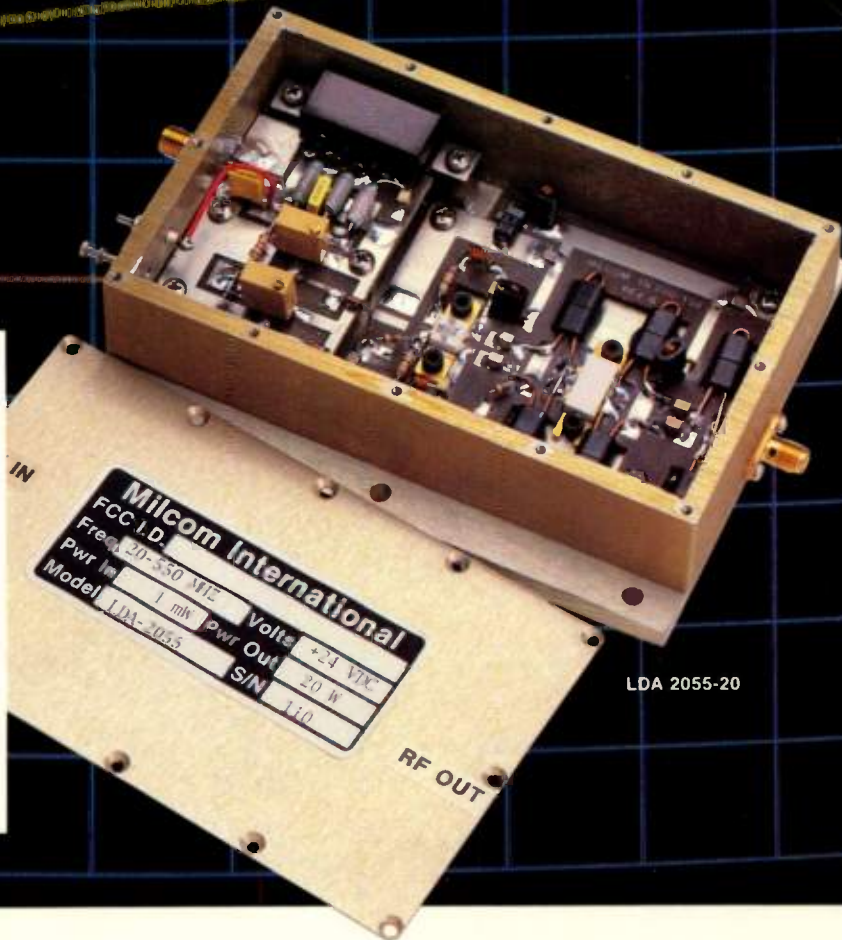
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LDA2040-20	225-400 MHZ	20 W	46DB	\$1,195	In Stk
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LDA5010-10	500-1000 MHZ	10 W	40DB	\$1,450	Stk-8wks
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Actual Frequency Response for LDA 2055-20 Amplifier:

SOLID STATE LINEAR POWER AMPLIFIER

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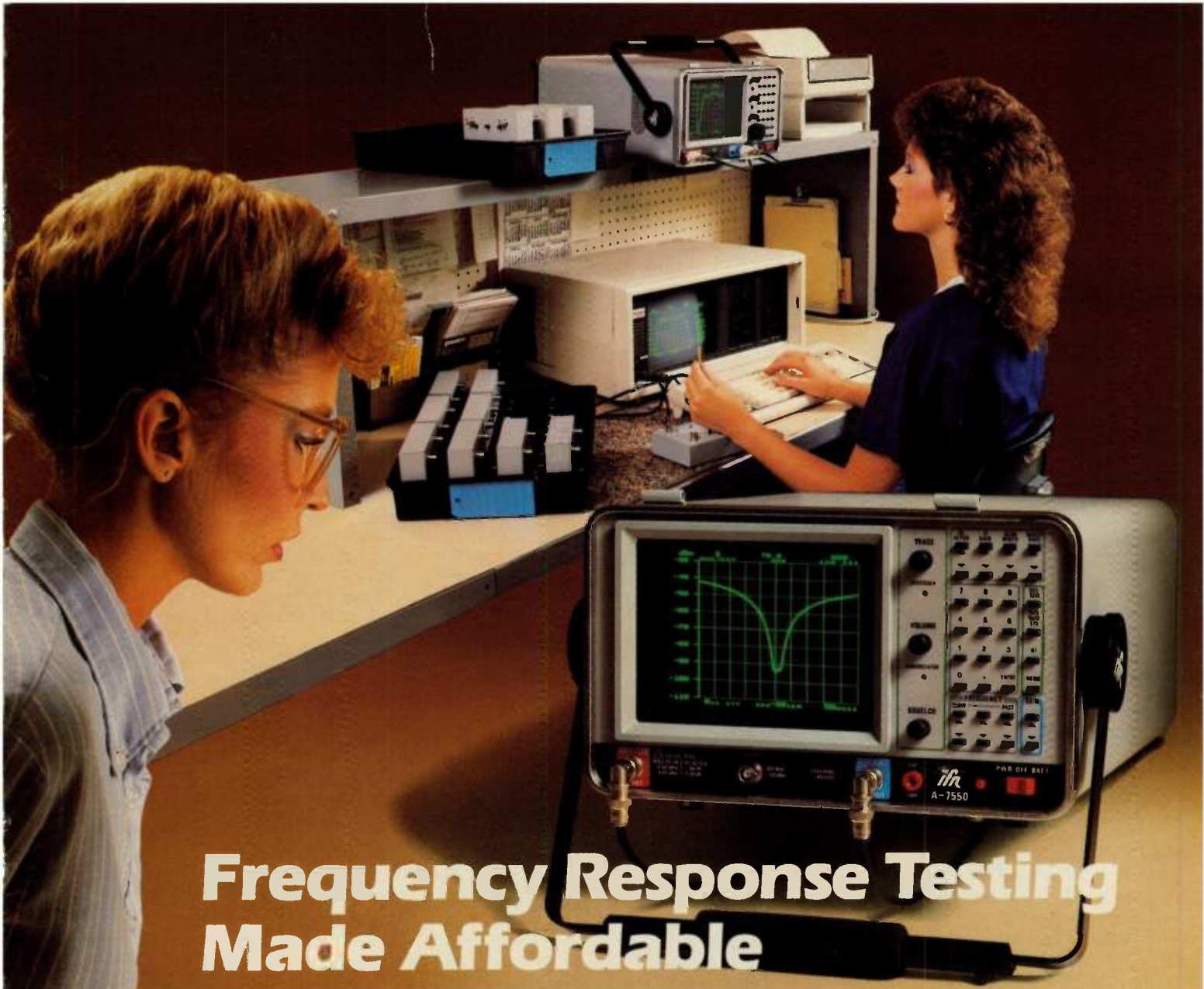
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Components

Capacitors	
Chip	.8
Other Fixed	.8
Variable	.9
Vacuum	.9
Crystals	
Quartz	.9
Ceramic Piezo	10
Diodes	
Varactor	10
PIN	10
Schottky	11
Noise	11
Inductors	
Fixed	12
Variable	12
Integrated Circuits	
RF Linear	12
High Speed Digital/RF	12
GaAs	17
Resistors	
Chip	17
Transformers	18
Transistors	
Small Signal Bipolar	18
Small Signal FET	18
RF Power FET	18
RF Power Bipolar	18
GaAs FET	18
Vacuum Tubes	
Transmitting	20
Receiving	20

Modules and Hybrids

Amplifiers	
Hybrid	20
Log	20
Limiting	22
Wideband OP Amp	22
Video	23
Detectors	23
Filters	
Crystal	24
SAW	26
LC	26
Active	28
Frequency Multipliers	28
Limiters	28
Mixers	
Diode Double Balanced	28
Image Reject	28

Modulators	
AM	28
FM/PM	30
FSK	30
QPSK	30
Pulse	30
Noise Sources	30
Oscillators	
Crystal	32
Clock	34
High Stability	34
VCO	34
SAW	36
LC	36
Phase Locked	36
Resonators	
SAW	36
Ceramic	38
LC	38

System Blocks and Subassemblies

Amplifiers	
IF	38
Log	38
LNA	38
RF Power	38
High Dynamic Range	40
Frequency Converters	40
Frequency Synthesizers	41
Motion/Proximity Sensors	42
Receivers	42
RF Modems	42
Tone Encoders/Decoders	42
Transmitters	44

Transmission Components

Antennas	
HF	44
VHF	44
UHF	45
EMC/EMI	45
Broadband	45
Antenna Positioners	45
Attenuators	
Coaxial	46
PC Mount	48
PIN	48
Variable	48
Programmable	48
Step	48
High Power	49
Coaxial Cable	49

Directional Couplers	50
Equalizers	50
Filters	
Cavity	50
Coaxial	51
Helical	51
Tunable	52
Highpass	52
Lowpass	53
Bandpass	54
Notch	54
Isolators/Circulators	55
Loads	55
Multicouplers	55
Multiplexers/Diplexers	56
Phase Shifters	56
Power Dividers/Combiners	56
Relays	
Vacuum	58
Mechanical	58
Multi-pole	58
Solid State	59
RF Connectors	59
Switches	
Electromechanical	60
Solid State	60
Terminations	60
Tuners	61

CAD/CAE Software and Systems

Software	
Circuit Analysis/Design	61
System Analysis/Design	61

continued on page 6

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Transmission Line.	62
EMC/TEMPEST.	62
Time Domain Analysis.	62
PC Board Design.	62
Circuit Optimization.	62
Systems - Hardware.	62

Materials and Hardware

Cabinets and Enclosures.	63
Ferrites & Other Magnetics.	63
Ceramics.	63
Heat Sinks.	63
Coatings Environmental.	63
Insulating Materials.	64
Components & Hybrid Packages.	64
PC Boards.	64
Crystal Blanks.	64
Sockets.	64
Electronic Hardware.	64
Substrates.	64
Equipment Racks.	64
Tools.	64

Test Equipment

Amplifiers (Lab).	65
Power Meters.	68
Antenna Measurement Systems.	68
Probes.	68
Calibrators/Standards.	68
Pulse Generators.	69
Dummy Loads.	69
RLC/Impedance Bridges.	70
Field Intensity Meters.	70
Signal Level Meter.	70
Frequency Counters.	70
Signal/Sweep Generators.	70
Modulation Meters.	72
Spectrum Analyzers.	72
Network Analyzers.	72
SWR Analyzers.	72
Noise Figure Meters.	72
Test Fixtures.	72
Noise Generators.	75
Wave/Distortion Analyzers.	75
Oscilloscopes.	75

Services

Antenna Measurement.	75
Equipment Calibration.	75

Circuit Assembly.	75
Instrument Rental/Leasing.	76
Consultants.	76
PC Layout/Fabrication.	76
Custom Hybrids.	76
Plating/Coating.	76
Custom ICs and ASICs.	79
Software Development.	79
EMC Testing.	79
Training/Education.	79

EMC Products

Absorbers.	80
Shielding	
Magnetic.	80
Foil.	80
Laminate.	80
Coatings.	80
AC Line Filters.	80
Anechoic Chambers.	80
Conductive Adhesives/Lubricants.	82
Conductive Coatings.	82
ESD Control Devices.	82
Surge/Transient Protectors.	82
Gaskets/Finger Stock.	82
Test Cells.	82
Hum Stop Filters.	85
Test Equipment & Systems	
Antennas.	85
Field Strength Meters.	85
Hazard Meters.	85
EMI/EMC Receivers.	85
EMI Generators.	85
Test Software.	85
Shielded Enclosures.	85

Design Guide

Noise Bandwidth	
Calculation.	87
Phase Difference Networks.	89
Theorems for Today's Networks.	94
Oxide Isolated Integrated Circuits	
for High Frequency Applications.	97

Company Index. 99

Advertisers' Index. 130

RFdesign

a Cardiff publication Established 1978

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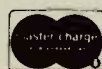
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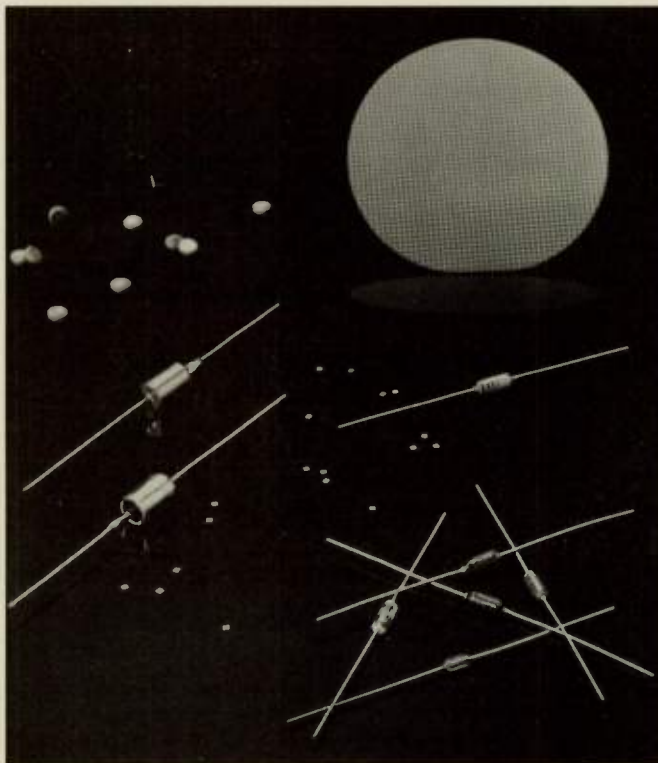
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Knox's unique design and processing techniques have been employed in this extension of devices, continuing to set new standards in customer satisfaction.

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SPECIFICATIONS

KSV14xx

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TR C2/C10 @ 1MHz, min. 10:1

Q @ 2V/1MHz, typical 500

KSV2xx

CT @ 4V from 20 to 200 pF

TR C4/C20 @ 50MHz, min. 6:1

Q @ 4V/50MHz, typical 500

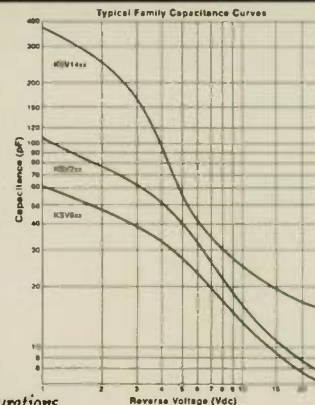
KSV8xx

CT @ 2V from 4 to 450 pF

TR C2/C20 @ 50 MHz, min. 6:1

Q @ 3V/50 MHz, typical 300

Available in DO-7, SMD and chip configurations



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DC current capacity: 100 mA
2, 3, 4 and 8 line styles (4 of each)
Kit D101 \$65

Common Mode Line Chokes

Current: .25 - 9 amps RMS
Inductance: 508 μ H - 10.5 mH
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Kit P202 \$100

Current Sensors

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Freq. resp.: 1 - 100 kHz, 50 - 400 Hz
Transformer and sensor-only versions
8 styles (15 total pieces)
Kit P203 \$50

Base/Gate Driver Transformers

Inductance: 1.5 mH Min.
Frequency: 10 - 250 kHz
2 single, 2 double section (2 of each)
Kit P204 \$50

Power Filter Chokes

Current: 3, 5, 10 amps
Inductance: 5 - 300 μ H
18 styles (48 total pieces)
Kit P205 \$75

Axial Lead Power Chokes

Current: .03-4.3 amps
Inductance: 3.9 μ H - 100 mH
60 styles (2 of each)
Kit P209 \$150

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RF SELECTOR GUIDE

When it comes to RF, Philips Components - Signetics is a World leader. The product line offers solutions for Cellular Radio, Cordless Phones, Two-way Communications, High Performance Receivers and a variety of other applications. The first company to offer a complete Cellular Chip Set anywhere in the world.

RF Components

NE605 Low Power Single Chip FM System

- 0.22uV sensitivity at 45MHz
- 90 dB dynamic range RSSI
- IF frequencies up to 25MHz
- 5.7mA supply current
- Mixer/Osc and IF Section Included
- >500 MHz RF Input
- New SSOP (Shrink small Outline Package)

NE604A Low Power FM IF

- 0.22uV sensitivity at 45MHz
- 90 dB dynamic range RSSI
- IF frequencies up to 25MHz
- 3.3mA supply current

NE602A Double Balanced Mixer

- Suitable for HF/VHF/UHF mixing
- 2.4mA supply current
- On Board Oscillator will function up to 300MHz (typ)
- >500 MHz RF Input
- 17dB conversion gain at 45MHz
- 5.0dB noise figure at 45MHz

Amplifiers

NE5209 Wideband Variable Gain Amplifier

- Bandwidth of 800Mhz (3dB)
- Variable Gain of 30dB
- 7dB Noise Figure
- High Impedance Differential Inputs
- 50 Ohm Differential Output

NE5205 Wideband High Frequency Amplifier

- Bandwidth of 600Mhz (3dB)
- 20dB of Insertion Gain
- 4.8dB ($Z_0=75$ Ohm) Noise Figure
- Input and output impedance matched to 50 and 75 Ohm systems

Comparator Components

NE578 Low Power Comparator

- Low Power 1.4mA @ 3.6V
- Programmable Unity Gain (0dB) 10mV to 1Vrms
- Power Down Mode ($I_{cc}=200\mu A$)
- Low external components count
- Multiple external summing capability
- 600 ohm drive capability
- Functional down to 2V

NE577 Low Power Comparator

- Low Power 1.4mA @ 3.6V
- Programmable Unity Gain (0dB) 10mV to 1Vrms
- Low External components count
- Functional down to 2V

NE575 Low Voltage Comparator

- Operates down to 3V
- Two Built in op-amps in each channel
- 600 ohm drive capability
- 0.12% typical THD
- New SSOP (Shrink Small Outline Package)

NE572 High Performance Comparator

- 110 dB Dynamic range
- Independent attack and recovery time
- Low output noise: 6uV typical
- Temperature compensated
- Very Low distortion (0.05% typical at 1kHz)

NE570/571 High Performance Comparator

- 110 dB Dynamic range
- Temperature compensated
- Built in op amps
- Excellent for high quality audio

Audio Processors

NE5750 Audio Processor - Companding and Amplifier Section

- Adjustable Gain Preamplifier
- Audio Companding
- Open Collector VOX output
- Power Down Mode
- Drivers for Speaker and Earpiece

NE5751 Audio Processor - Filter and Control Section

- Switched Capacitor Bandpass Filter
- Programmable DTMF generator
- Digitally Controlled Volume Control
- Power Down Mode
- I²C Bus Controlled

Signetics

Philips Components



PHILIPS

Document	853-1401
ECN No.	097959
Date of Issue	October 25, 1989
Status	Product Specification
Application Specific Product	

NE/SA605

High Performance Low Power Mixer FM IF System

DESCRIPTION

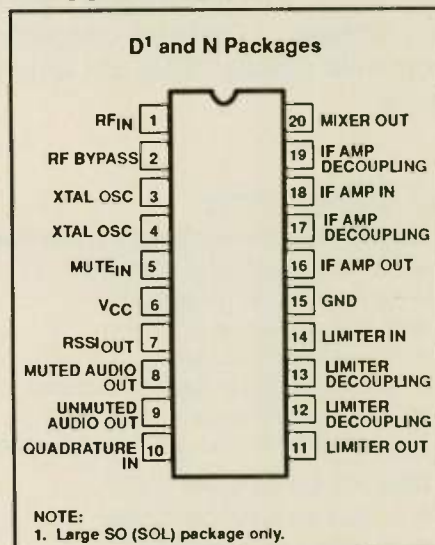
The NE/SA605 is a high performance monolithic low-power FM IF system incorporating a mixer/oscillator, two limiting intermediate frequency amplifiers, quadrature detector, muting, logarithmic received signal strength indicator (RSSI), and voltage regulator. The NE/SA605 combines the functions of Signetics' NE602 and NE604A, but features a higher mixer input intercept point, higher IF bandwidth (25MHz) and temperature compensated RSSI and limiters permitting higher performance application. The NE/SA605 is available in 20-lead dual-in-line plastic and 20-lead SOL (surface-mounted miniature package).

The NE/SA605 and NE/SA615 are functionally the same device types. The difference between the two devices lies in the guaranteed specifications. The NE/SA615 has a higher I_{CC} , lower input third order intercept point, lower conversion mixer gain, lower limiter gain, lower AM rejection, lower SINAD, higher THD, and higher RSSI error than the NE/SA605. Both the NE/SA605 and NE/SA615 devices will meet the EIA specifications for AMPS and TACS cellular radio applications.

FEATURES

- Low power consumption: 5.7mA typical at 6V
- Mixer input to >500MHz
- Mixer conversion power gain of 13dB at 45MHz
- Mixer noise figure of 4.6dB at 45MHz
- XTAL oscillator effective to 150MHz (L.C. oscillator to 1GHz local oscillator can be injected)
- 102dB of IF Amp/Limiter gain
- 25MHz limiter small signal bandwidth
- Temperature compensated logarithmic Received Signal Strength Indicator (RSSI) with a dynamic range in excess of 90dB
- Two audio outputs – muted and unmuted
- Low external component count; suitable for crystal/ceramic/LC filters
- Excellent sensitivity: 0.22 μ V into 50 Ω matching network for 12dB SINAD (Signal to Noise and Distortion ratio) for 1kHz tone with RF at 45MHz and IF at 455kHz
- SA605 meets cellular radio specifications
- ESD hardened

PIN CONFIGURATION



APPLICATIONS

- Cellular radio FM IF
- High performance communications receivers
- Single conversion VHF/UHF receivers
- SCA receivers
- RF level meter
- Spectrum analyzer
- Instrumentation
- FSK and ASK data receivers
- Log amps
- Wideband low current amplification

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
20-Pin Plastic DIP	0 to +70°C	NE605N
20-Pin Plastic SOL (Surface-mount)	0 to +70°C	NE605D
20-Pin Plastic DIP	-40 to +85°C	SA605N
20-Pin Plastic SOL (Surface-mount)	-40 to +85°C	SA605D

Document	
ECN No.	
Date of Issue	March 26, 1990
Status	Objective Specification
Application Specific Product	

NE/SA/SE5209

Wideband Variable Gain Amplifier

DESCRIPTION

The NE5209 represents a breakthrough in monolithic amplifier design featuring several innovations. This unique design has combined the advantages of a high speed bipolar process with the proven Gilbert architecture.

The NE5209 is a linear broadband RF amplifier whose gain is controlled by a single DC voltage. The amplifier runs off a single 5 volt supply and consumes only 40mA. The amplifier input is high impedance ($1k\Omega$) and differential. The output is 50Ω differential. Therefore, the 5209 can simultaneously perform AGC, impedance transformation, and the balun functions.

The noise performance is excellent (7dB noise figure) over the 800MHz bandwidth. Furthermore, the noise performance degrades at a comparatively slow rate as the gain is reduced. This is an important feature when building linear AGC systems.

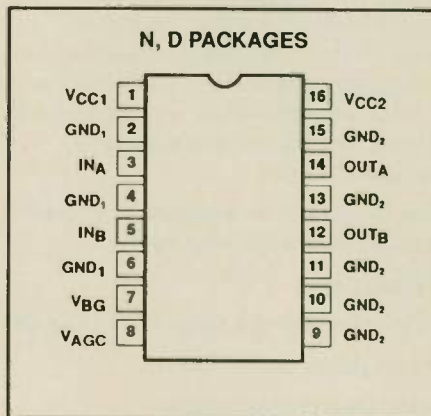
FEATURES

- 800MHz bandwidth
- High Impedance differential input
- 50Ω differential output
- Single 5V power supply
- 0–1V gain control
- 0–30dB gain
- Exceptional $V_{CONTROL} / V_{GAIN}$ linearity
- 7dB noise figure
- Full ESD protection
- Military version available
- Easily cascadable

APPLICATIONS

- Linear AGC systems
- Very linear AM modulator
- RF balun
- Cable TV multi-purpose amplifier
- Fiber optic AGC
- RADAR
- User programmable fixed gain block

PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic SO	0°C to +70°C	NE5209D
16-Pin Plastic DIP	0°C to +70°C	NE5209N
16-Pin Plastic SO	-40°C to +85°C	SA5209D
16-Pin Plastic DIP	-40°C to +85°C	SA5209N
16-Pin Plastic DIP	-55°C to +125°C	SE5209N

Signetics

Document	
ECN No.	
Date of Issue	May 24, 1990
Status	Preliminary Specification
Application Specific Product	

NE/SA577

Unity Gain Level Programmable
Low Power Compandor

DESCRIPTION

The NE/SA577 is a unity gain level programmable compandor designed for low power applications. The NE577 is internally configured as an expander and a compressor to minimize external component count.

The NE577 will be available in a 14-pin plastic DIP and SO packages.

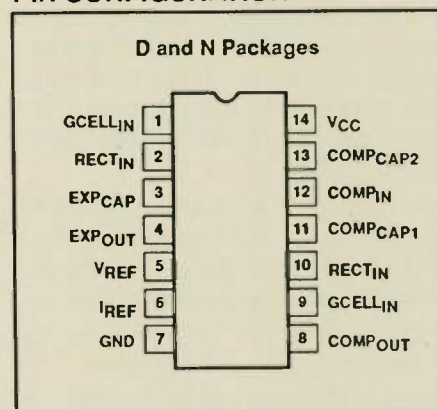
FEATURES

- Operating voltage range from 2V to 7V
- Low power consumption
- 0dB level programmable
- Wide input/output swing capability
- Low external component count
- SA577 meets cellular radio specifications
- ESD hardened

APPLICATIONS

- High performance portable communications
- Cellular radio
- Cordless telephone
- Consumer audio
- Wireless microphones
- Modems
- Electric organs
- Hearing aids

PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
14-Pin Plastic DIP	0 to +70°C	NE577N
14-Pin Plastic SO	0 to +70°C	NE577D
14-Pin Plastic DIP	-40 to +85°C	SA577N
14-Pin Plastic SO	-40 to +85°C	SA577D

Philips Components



PHILIPS

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(See our ad on page 35)
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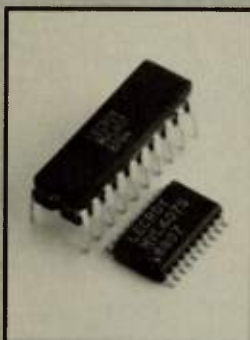
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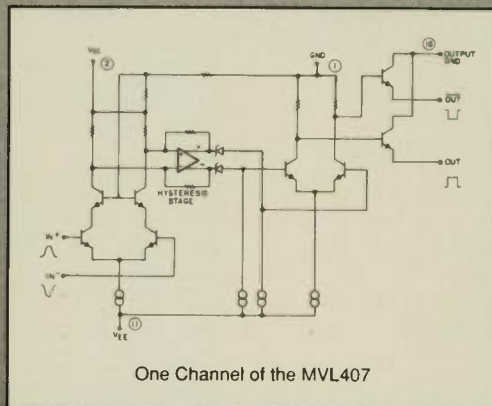
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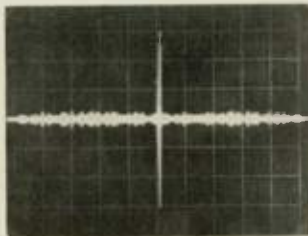
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Hazeltine has combined ASIC and Surface Acoustic Wave (SAW) technology to provide programmable phase coded matched filters that are now key processors in many tactical radar and spread-spectrum communications systems. These filters provide the processing gain necessary for jam-resistant operation and the electronic programmability essential for a secure system. Small size, low cost and operation in a military environment are inherent features of Hazeltine's programmable matched filters. Standard and custom designs are available.



Programmable SAW Matched Filter
Model: PSWMF-128-80-12.8 MSK



Correlated Output for 128-Chip MSK
Input Signal (Note 20-dB sidelobe
level.) Scale: 2 μ s/cm

MODEL SPECIFICATIONS

NUMBER OF TAPS	128
CENTER FREQUENCY	80 MHz
TAP SPACING/CHIP RATE	78.125 ns/12.8 Mb/s
INSERTION LOSS	15 dB
MODULATION	MSK
PROGRAMMING RATE	To 15 MHz
MISMATCH LOSS	Less than 0.5 dB
MAXIMUM INPUT LEVEL	+20 dBm
POWER DISSIPATION	2 watts
TEMPERATURE RANGE	-55°C to +125°C
SIZE	1.75 x 0.75 x 0.125 in

CONTACT HAZELTINE TO DISCUSS YOUR REQUIREMENTS
FOR SAW CHIRP AND PHASE CODED MATCHED FILTERS.



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• MITEQ's long-term warranty and the long list of custom-selectable options, you will realize that MITEQ's bipolar amplifiers offer unparalleled bargains. Study the data and make your own decision.

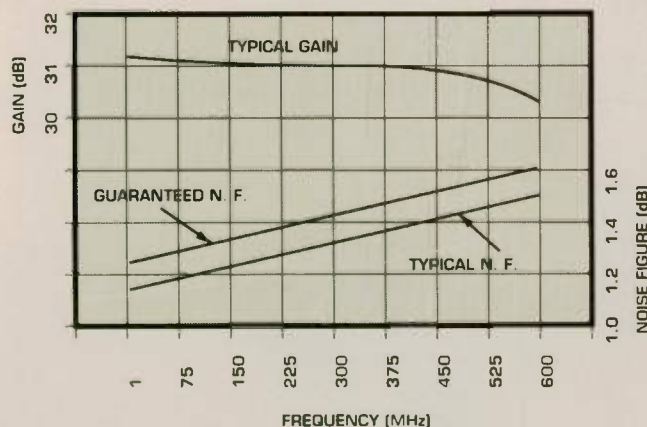
OPTIONS

- Integral input limiter
- Bias supplied via the output port
- Low-loss input "bias tee" for photo diodes
- Low frequency operation to below 10 kHz
- Fast recovery from input overloads
- Amplitude and phase matching
- Lower guaranteed noise figure

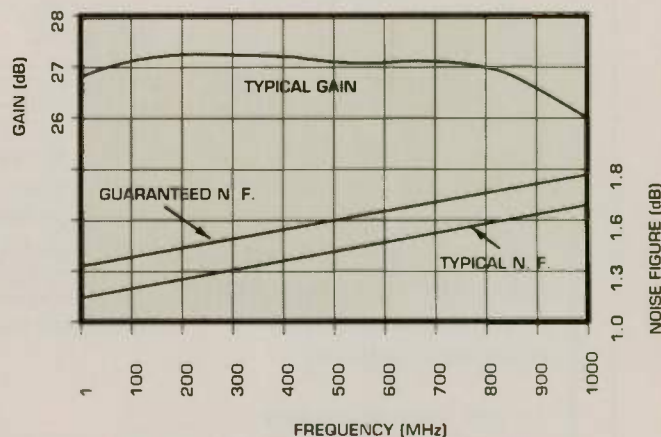
Model Number	Frequency (MHz)	Gain (Min.) (dB)	Gain Var. (Max.) (±dB)	Noise Figure (Max.) (dB)			VSWR (Max.)	Dynamic Range 1 dB Gain Comp. Output (Min., dBm)	Nom. DC Power (+15V, mA)
				Low End	Mid Band	High End			
AU-1054	1-500	30	0.5	1.25	1.4	1.5	2:1	+8	55
AU-2A-0150	1-500	30	0.5	1.25	1.4	1.5	2:1	+8	55
AU-3A-0150	1-500	45	0.5	1.25	1.4	1.5	2:1	+10	67
AU-1149	10-180	15	0.5	4.5	5.5	6.5	2:1*	+18	75
AU-1291	.015-500	60	0.75	1.25	1.4	1.5	2:1	+8	90
AM-1300	.01-1000	25	0.75	1.4	1.6	1.8	2:1	+6	50
AM-1052	1-1000	26	0.75	1.4	1.6	1.8	2:1	+6	50
AM-2A-000110	1-1000	27	0.75	1.4	1.6	1.8	2:1	+8	50
AM-1299	1-1000	38	0.75	1.4	1.6	1.8	2:1	+9	75
AM-3A-000110	1-1000	37	0.75	1.4	1.6	1.8	2:1	+9	75

* 75 ohm impedance level

AU-1054 and AU-2A-0150



AM-1052 and AM-2A-000110



For additional information, please call Bill Pope at (516) 436-7400, on extension 282.



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Come rain or shine, Trilithic will have the attenuator you need.

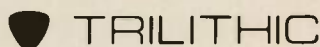
There are some things you're sure of. Trilithic is one of them. We have one of the largest selections of fixed attenuators. The combinations of power, frequency, and connectors are almost endless. And, in most cases, the attenuator you need is in-stock and ready for delivery.

To get the attenuator you need, call Toll-Free: 800-344-2412

An example of the range covered by just a few of our attenuators:

	FREQUENCY	POWER
FP50	DC-2GHz	1 Watt
FP18	DC-18GHz	2 Watts
HFP50	DC-8GHz	10-15 Watts
HFP51	DC-4GHz	20 Watts
HFP52	DC-18GHz	10-20 Watts
HFP53	2-18GHz	50 Watts

Additional models available.



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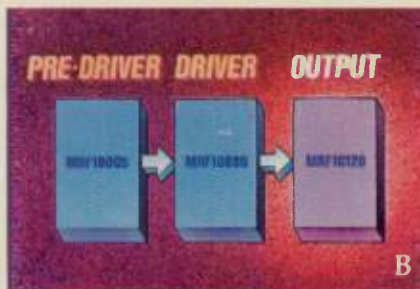
RF Design News



High gain power modules for mobile cellular radios.

From the industry leader in power modules comes a new high gain improvement. The industry standard, the MHW806A, now comes in a 0-dBm version as the MHW807 Series. Instead of the previous power requirement of 30 mW, the new series requires only 1 mW to obtain 6W of output power, thanks to two new gain stages.

The MHW807 Series is perfect for all cellular radio applications. They offer controllable, stable performance over more than the 35 dB range in P_o that's needed. Two different frequency models are available: 820 to 850 MHz and 870 to 905 MHz.



Long pulse microwave power transistor.

Motorola continues to expand its long pulse microwave power transistor portfolio with the introduction of its new 120 watt L-Band transistor, the MRF10120. This output device completes the lineup consisting of the pre-driver MRF10005 and the MRF10030 driver.

The MRF10120 operates on a power supply of 36 volts and delivers 120 watts of peak power for typically less than 15 watts of peak RF input power. It's designed for common base amplifier applications such as JTIDS (military) and Mode S (commercial) transmitters. The frequency of operation extends from 960 to 1215 MHz.



Satellite microwave power transistors.

Three new microwave power transistors are available for large-signal output and driver amplifier stages for satellite up/down links. The MRA1600-2, MRA1600-13 and MRA1600-30 are designed for Class C, common base amplifiers that operate in the 1600-1660 MHz frequency range. They provide 2.2, 12.7 and 30 watts of minimum power respectively.

These devices offer the highest in reliability and performance. They feature gold metalization, diffused ballast resistors and internal compensation for impedance matching control. All this is offered in a low-cost microwave package for cost efficiency.



A new breed of workhorse.

Now you can have the same rugged workhorse you've enjoyed for your high-power applications in a medium-power, broadband amplifier. The PAA series of broadband amplifiers are bred to outwork and outlast the competition.

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For high-reliability performance you can depend on in a medium-power amplifier, Motorola's PAA Series is the answer.

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To get more information on any of the Motorola products shown here, contact your local Motorola sales office, complete and return the coupon below to Motorola Semiconductor Products, Literature Distribution Center, P.O. Box 20912, Phoenix, AZ 85036. Or call toll-free any weekday, 8:00 a.m. to 4:30 p.m. (MST) 1-800-521-6274.



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 D ☐ PAA Broadband Amplifiers

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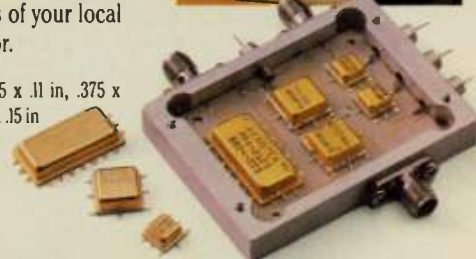
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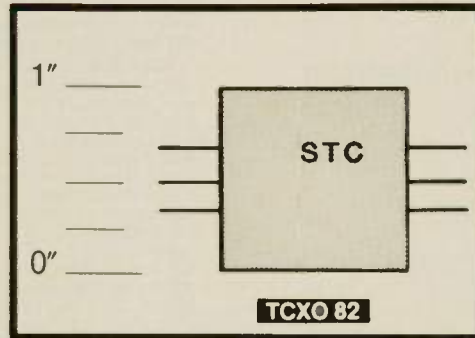
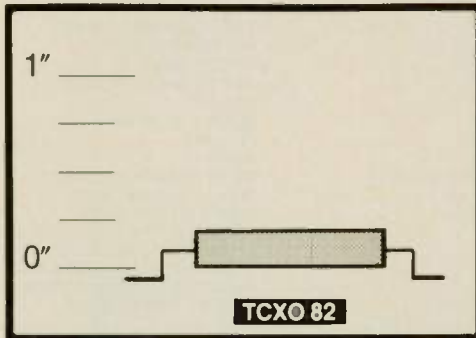
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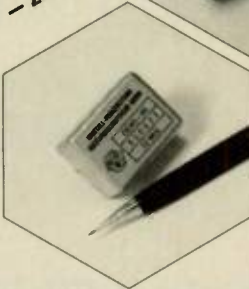
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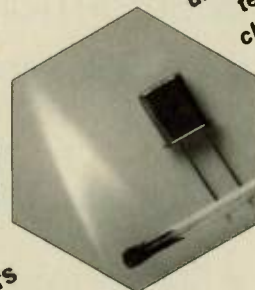
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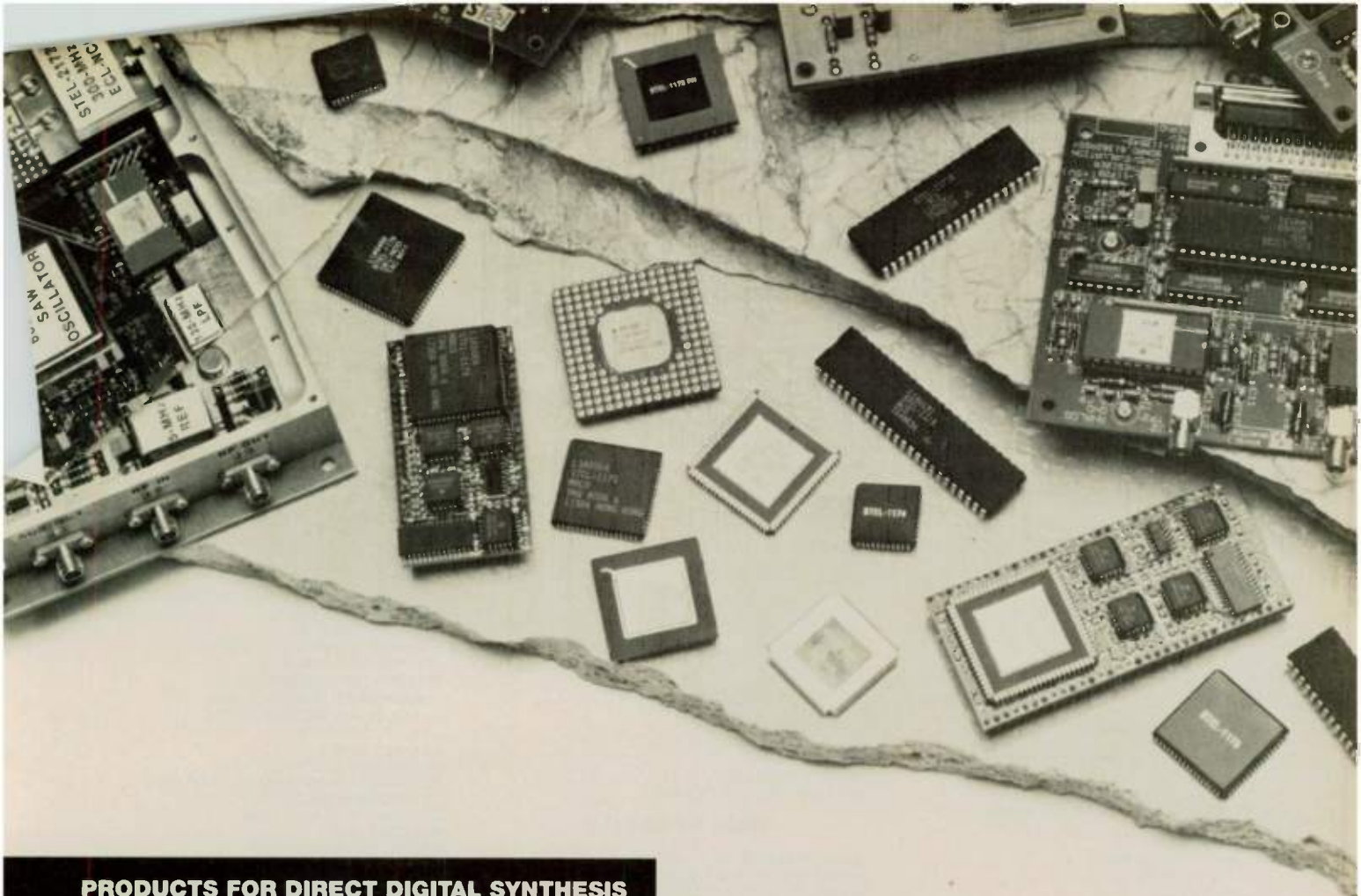
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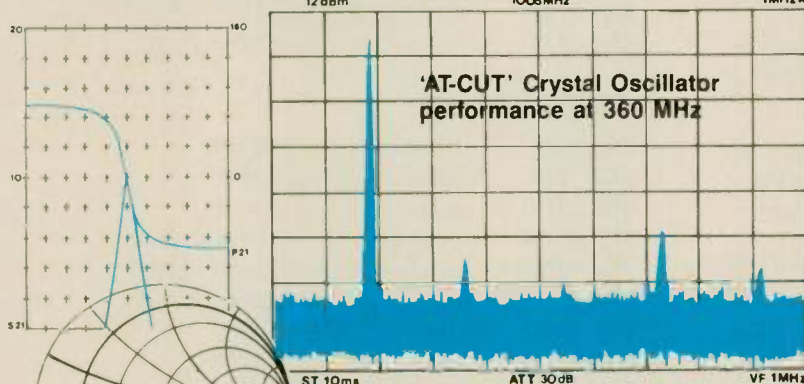
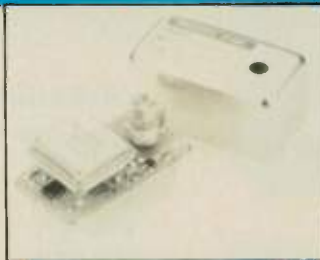
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ECL compatible crystal oscillators at frequencies up to 600MHz are offered in 14 pin DIP resistance welded packages. These oscillators are also compatible with 100K logic outputs and have

a temperature stability of 50 to 100ppm over the temperature range of 0°C to 70°C. Complementary outputs are standard while single output, voltage controlled units are optionally available. A sine wave output is also optionally available. These ECL compatible oscillators are designed for high resolution graphics, microprocessor clock and communications applications.

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Dielectric Resonators (DR)-Resonators

Murata Erie offers a complete line of dielectrics with high permittivity that make excellent resonators for microwave oscillators and filters.

FEATURES

- 400 MHz to 30 GHz Frequency Range
- High Unloaded Q ■ High Dielectric Constant
- Wide Temperature Coefficient Range: -4 to +10 ppm/°C

TE mode resonators are offered in both disc and coaxial cylinder configurations while TEM mode units are offered in 1/4 and 1/2 wavelength models. Substrates offer high Q's and dielectric constants making them ideal for applications in Microwave Integrated Circuits (MIC).

Dielectric Resonator Oscillators (DRO)

A complete line of Dielectric Resonator Oscillators, utilizing Murata Erie resonators, is offered.

FEATURES

- 3 GHz to 11.5 GHz Frequency Range
- Typical Frequency Stability of ± 1 MHz from -40°C to +60°C
- Typical Power Output of +7 dBm into a 50 Ohm Load
- Case Sizes Ranging

from 20 mm x 12 mm x 9.5 mm to 60 mm x 30 mm x 18 mm. These DRO's are ideal for application in Microwave Communications, DBS, TVRO, CATV, Network Analyzers and Satellite Communications.

Voltage Controlled Oscillators (VCO)

These VCO's also are based on state-of-the-art Murata Erie dielectric materials.

FEATURES

- 400 MHz to 1.5 GHz Frequency Range
- Typical Frequency Stability of ± 2 MHz from -35°C to +80°C
- Typical Tuning Range of ± 12 MHz

These devices find wide application in Cellular Mobile Telephone (CMT), Data Transmission, Global Positioning System (GPS) and Spread Spectrum Products.

Isolators

These dielectric resonator devices are used as interstage matching devices to absorb antenna mismatching and maintain stable transmission.

FEATURES

- Power Ratings from 2 to 10 Watts
- 15 dB min. Isolation
- Less than 1 dB Insertion Loss
- Sizes Ranging From 6.9 mm x 6.8 mm x 4 mm to 15 mm x 15 mm x 9 mm
- Surface Mount Packages Available

Applications include Cellular Mobile Telephone (CMT), Microwave Test Equipment, etc.

Dielectric Microwave Filter (Gigafil®)

These microwave filters are based on Dielectric Resonator technology and cover the frequency range from 400 MHz to 2 GHz. They offer extremely good temperature stability (± 5 ppm/°C), high selectivity and low insertion loss.

Applications include CMT, GPS, Data Transmission, Navigation Systems, Cordless Telephones, Land Mobile Radio (LMR) and Spread Spectrum Systems.

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Wide Band / Low Noise Amplifiers										
	W40C	1MHz—40MHz	42	± 5	1.0 Typ 1.2 Max	+ 5	2:1	+ 15	20	C/SMA
	W50ETC	10KHz—50MHz	24	± 5	5.3 Typ 6.0 Max	+23	2:1	+ 15	125	E-75/BNC
	W50ATC	10KHz—50MHz	50	± 5	1.3 Typ 1.5 Max	+ 5	2:1	+ 15	25	C-75/BNC
	W110F	5MHz—110MHz	55	± 5	1.1 Typ 1.2 Max	+ 15	2:1	+ 15	80	C/SMA
	W110H	5MHz—110MHz	30	± 5	1.2 Typ 1.4 Max	+ 5	2:1	+ 15	30	C/SMA
	W500K	1KHz—500MHz	30	± 1	1.7 Typ 2.2 Max	+ 3	2:1	+ 15	25	C-75/BNC
	W500C	5MHz—500MHz	40	± 5	1.4 Typ 1.6 Max	+ 10	2:1	+ 15	50	C/SMA
	W500EF	5MHz—500MHz	60	± 5	1.3 Typ 1.4 Max	+ 20	2:1	+ 15	190	A/SMA
	W500H	5MHz—500MHz	33	± 5	1.2 Typ 1.4 Max	+ 5	2:1	+ 15	25	C/SMA
	W1G2M	10KHz—1000MHz	30	± 1	2.0 Typ 3.0 Max	+ 5	2:1	+ 15	35	C-75/SMA
	W1G2H	5MHz—1000MHz	30	± 5	1.3 Typ 1.5 Max	+ 5	2:1	+ 15	40	C/SMA
	W2GH	500MHz—2000MHz	22	± 1	4.0 Typ 4.5 Max	+ 5	2:1	+ 15	30	C/SMA
	WFR1-4GA-14	100MHz—4000MHz	28	± 1	3.5 Typ 4.0 Max	+ 14	2:1	+ 15	100	A-75/SMA
Medium Power Amplifiers										
	P150D	35KHz—150MHz	27	± .5	5.0 Typ	+ 30	2:1	+ 24	400	H/SMA
	P150M	500KHz—150MHz	26	± 5	5.0 Typ	+ 30	2:1	+ 24	600	H/BNC
	P150ML	400KHz—150MHz	24	± 1	11 Typ	+29.5	2:1	+ 24	600	H/BNC
	P500A	2MHz—500MHz	37	± 5	4.5 Typ	+ 30	2:1	+ 24	500	H/SMA
	P500L	5MHz—500MHz	17	± .7	10 Typ	+ 30	2:1	+ 24	420	H/BNC
	P500ML	2MHz—500MHz	16	± 1	11 Typ	+ 28	2:1	+ 24	600	H/BNC
	P1GB	50MHz—1000MHz	30	± 1	5.5 Typ	+ 30	2:1	+ 20	800	A-S/SMA
	P1000M	5MHz—1000MHz	20	± 5	6 Typ	+ 21	2:1	+ 20	200	H/SMA
	P2GF-2	10MHz—2000MHz	32	± 1	7.5 Typ	+ 30	2:1	+ 15	1000	FW1/SMA
	P42GA-29	.5GHz—4.2GHz	30	± 1.5	6.5 Typ	+ 29	2:1	+ 20	1200	FW75/SMA

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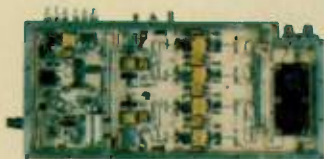
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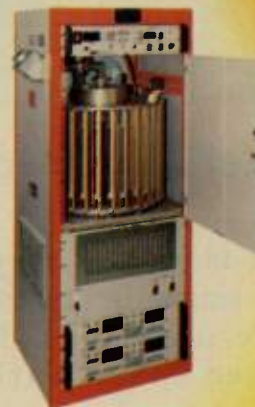
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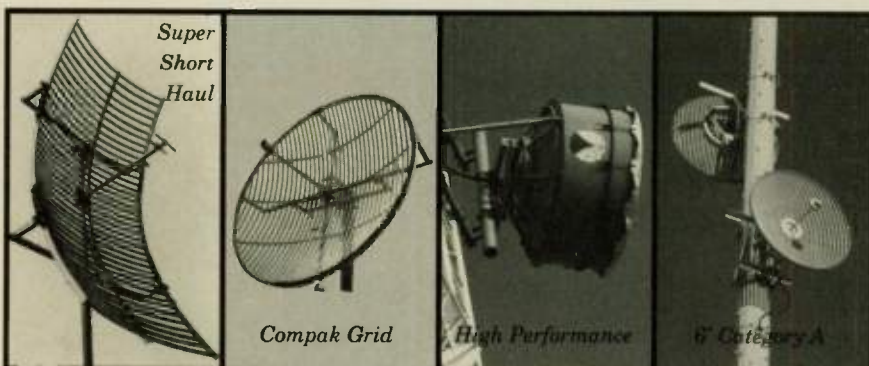
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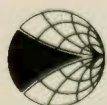
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- LOW LOSS
- HIGH ISOLATION
- EXTERNAL TERMINATIONS

4 WAY COMBINER MODEL

D2650	54-88 MHz
D2685	88-108 MHz
D2655	174-216 MHz

TYPICAL SPECIFICATIONS

INS. LOSS	0.15db
ISOLATION	25db
VSWR	1.2:1
POWER	2Kw

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input N
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1/839	50Ω	DC-1GHz	0-22.1dB	1dB
847	75Ω	DC-1GHz	0-102.5dB	1dB
870	75Ω	DC-1GHz	0-132dB	1dB
4440	50Ω	DC-1.5GHz	0-130dB	10dB
4450	50Ω	DC-1.5GHz	0-127dB	1dB
1/4450	50Ω	DC-1GHz	0-16.5dB	1dB
4467	75Ω	DC-1GHz	0-31dB	1dB
0/400	50Ω	DC-500MHz	1-13dB	—
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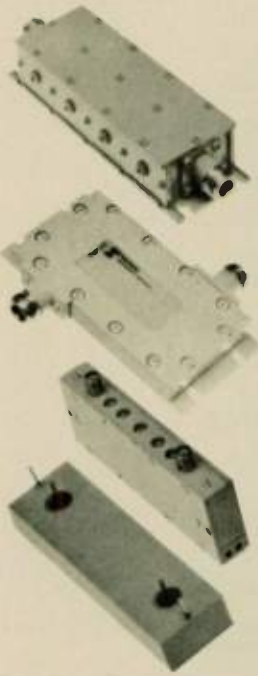
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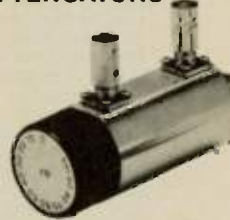
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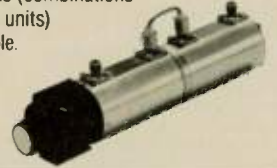
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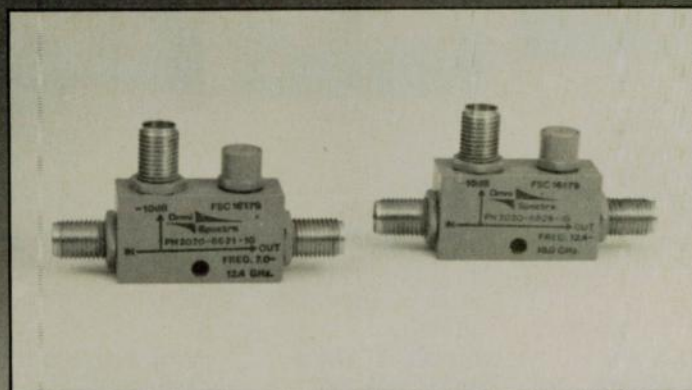
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ESCO ASSOC.
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FILTRON CO. INC.
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GAMMA-F CORP.
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KDI/TRIANGLE ELECTRONICS, INC.
KW MICROWAVE CORPORATION
L & M ENGINEERING SOUTH
LARK ENGINEERING CO., DIV. OF BAIER & BAIER INC.
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MICROWAVE RESOURCES, INC.
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RDL, INC.
REACTEL, INC.
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RLC ELECTRONICS, INC.
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TELEMET/GDC
TELEWAVE INC.
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TRILITHIC, INC.
TTE INC.
TUCKER ELECTRONICS
UHF ASSOCIATES
UNITED TRANSFORMER CORP.,
SUB. OF OPT INDUSTRIES
VANGUARD ELECTRONICS CO.
VIEWSONICS
WAVETEK RF PRODUCTS, INC.
WAVETEK SAN DIEGO, INC.
WESTERN MICROWAVE INC.
WI-COMM ELECTRONICS INC.
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CAPTOR CORP.
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COILCRAFT
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COMARK COMMUNICATIONS, INC.
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DAYTON-GRANGER, INC.
DECIBEL PRODUCTS, INC.
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DYMA ENGINEERING
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EATON CORP., COMPONENT & SEMICONDUCTOR
EATON CORP., MICROWAVE PRODUCTS DIV.
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ESC ELECTRONICS CORP.
ESCO ASSOC.
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FERRETEC, INC.
FILTRON CO. INC.
FOX ELECTRONICS
FUSION ELECTRONICS, INC.
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GAMMA-F CORP.
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HUBER AND SUHNER AG
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HUGHES AIRCRAFT CO. MICROWAVE PRODUCTS DIV.
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K & L MICROWAVE, INC.
KDI/TRIANGLE ELECTRONICS, INC.
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KW MICROWAVE CORPORATION
L & M ENGINEERING SOUTH
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RAMAR COMMUNICATIONS, INC.
RDL, INC.
REACTEL, INC.
REMEC
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RF PROTOTYPE SYSTEMS
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TAUBER-DREYER CORP.
TDK CORP. OF AMERICA
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TELEWAVE INC.
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TEMEX ELECTRONICS INC.
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TTE INC.
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COAXIAL DYNAMICS, INC.
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HUBER AND SUHNER AG

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KDI/TRIANGLE ELECTRONICS, INC.
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MICROLAB/FXR
MICRONETICS INC.
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RCD COMPONENTS INC.
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RFDC INC.
RLC ELECTRONICS, INC.
ROHDE AND SCHWARZ GMBH CO
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SOLITRON DEVICES, INC. & SOLITRON/MICROWAVE
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TELONIC BERKELEY, INC.
TEST EQUIPMENT CORP.
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TROMPETER ELECTRONICS
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DAJEM TECHNOLOGY CO.
DECIBEL PRODUCTS, INC.
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ELCOM SYSTEMS INC.
ENI
F S Y MICROWAVE INC.
GAMMA-F CORP.
INTERAD LTD.
KDI/TRIANGLE ELECTRONICS, INC.
L & M ENGINEERING SOUTH
ELECTRONIC RESEARCH LABS
LOCUS INC.
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LORCH ELECTRONICS DIV. OF VERNITRON CORP.
LUCAS EPSCO INC.
M/A-COM GOVERNMENT PRODUCTS, INC.
(Micro-Tel and Reaction Insts.)

RF transmission components

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MERRIMAC INDUSTRIES, INC.
MICON, INC.
MICRO COMMUNICATIONS, INC.
MICROLAB/FXR
MICROWAVE RESOURCES, INC.
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MU-DEL ELECTRONICS, INC.
NARDA MICROWAVE CORP.
NORSAL INDUSTRIES INC.
NORTH HILLS ELECTRONICS, INC.
OAKBURY COMPONENTS LTD.
OLEKTRON CORP.
OLEKTRON SYSTEMS, INC.
P.G. ELECTRONICS LTD.

PENSTOCK INC.

QUAL-TECH, DIV. OF FLEXCO MICROWAVE INC.
REACTION INSTRUMENTS, INC., A DIV OF ADAMS-RUSSELL
REMEC
RF COMPONENTS
RF PRODUCTS, INC.

ROHDE AND SCHWARZ GMBH CO

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TELE-TECH CORP.
TELEMET/GDC
TELEWAVE INC.
TRANSCO PRODUCTS, INC.
VORTEX MICROWAVE CORP.

WERLATONE, INC.

WI-COMM ELECTRONICS INC.
WIDE BAND ENGINEERING CO. INC.
XETRON CORP.

L & M ENGINEERING SOUTH
LARK ENGINEERING CO., DIV. OF BAIER & BAIER INC.
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TELEDYNE MICROWAVE
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TOKO AMERICA, INC.
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TTE INC.
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VIEWSONICS
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MICRO COMMUNICATIONS, INC.
MICROLAB/FXR
MICRONETICS INC. /NEM
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MICROWAVE TECHNIQUES
MIDWEST MICROWAVE INC.
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NARDA
NORSAL INDUSTRIES INC.
OLEKTRON SYSTEMS, INC.
P&H LABORATORIES
PACIFIC MONOLITHICS
PASCALL ELECTRONICS LTD.
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PULSAR MICROWAVE CORP.
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REMEC
RHG ELECTRONICS LABORATORY - A M/A-COM CO
RICHARDSON ELECTRONICS, LTD./RF GAIN, LTD.
ROBINSON LABS
SAGE LABORATORIES, INC.
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SYNERGY MICROWAVE CORP.
SYSTRON DONNER CORP., MICROWAVE DIV.
TELE-TECH CORP.
TELEDYNE MICROELECTRONICS
TELEDYNE RYAN ELECTRONICS
TEST EQUIPMENT CORP.
TRANSCO PRODUCTS, INC.
VENTRONICS MICROWAVE CORP.
VORTEX MICROWAVE CORP.
WAVELINE INC.
WERLATONE, INC.

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ANGHEL LABORATORIES, INC.
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APPLIED SPECIALTIES, INC.

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CAIN FORLAW CO.

CALIFORNIA EASTERN LABORATORIES

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CHESTERFIELD PRODUCTS INC.
COM DEV
COMAD COMMUNICATIONS LTD.
COMMUNITRONICS
CORVUS MICROWAVE, INC.
CIR-O-TEL MICROWAVE INC.
DADEN ASSOCIATES, INC.
DAJEM TECHNOLOGY CO.
DECIBEL PRODUCTS, INC.
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KAPTRON INC.
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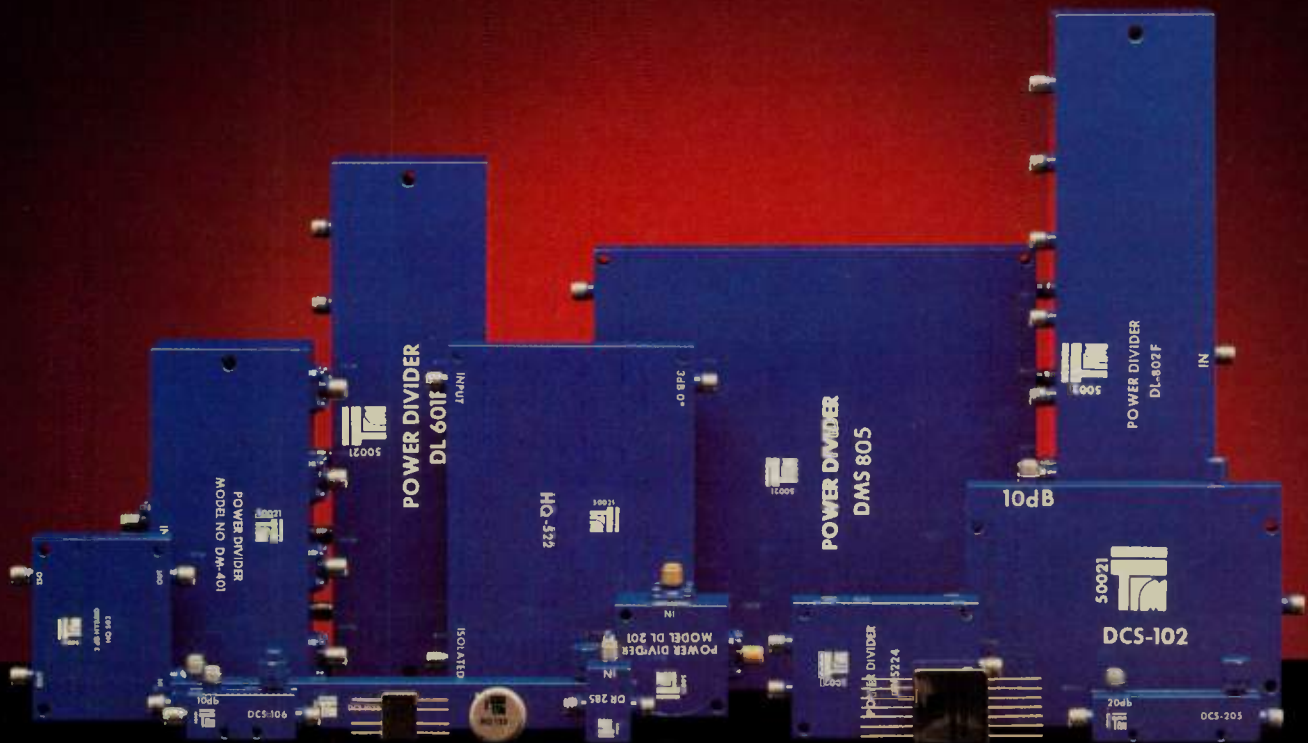
PHASE SHIFTERS

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DMS 221	2-18	1.00	20	1.50:1	1.50:1	0.40	6	30	1
DMS 347	6-18.5	1.00	15	2.00:1	1.70:1	0.50	5	5	3
DMS 421	2-18	1.75	20	1.50:1	1.50:1	0.50	8	30	2
DMS 447	6-18.5	1.25	15	1.60:1	1.60:1	0.30	8	30	1

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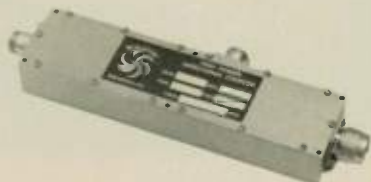


TYPICAL SPECIFICATIONS MODEL D2599

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.

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TYPICAL SPECIFICATIONS MODEL C2523

FREQUENCY RANGE 100-400 MHz
COUPLING 30db
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The model C2523 features exceptional coupling linearity vs input power and non-destructive precision stainless steel connectors.

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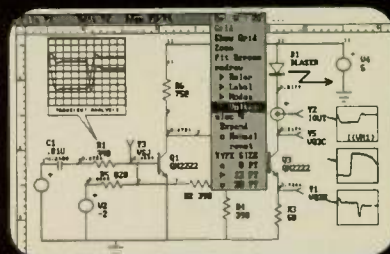
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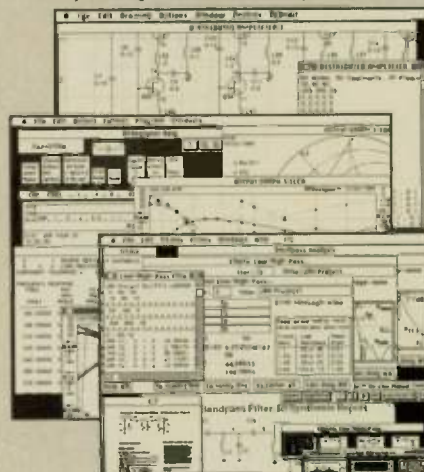
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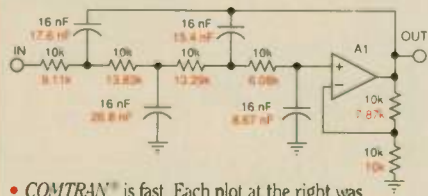
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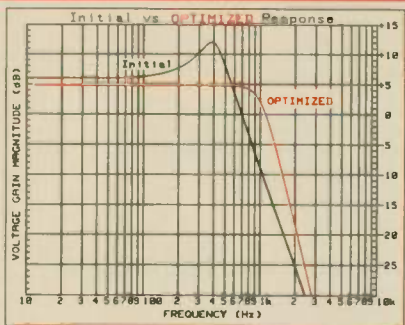
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Feature	Fluke 6082A	HP 8642B	HP 8644A/002
Frequency Range	100 kHz-2.112 GHz	100 kHz-2.115 GHz	252 kHz-2.060 GHz
SSB Phase Noise @1 GHz			
20 kHz offset	-131 dBc/Hz	-134 dBc/Hz	-128 dBc/Hz
1 kHz offset	-94 dBc/Hz	No Specification	-91 dBc/Hz
Non-harmonic Spurious @1 GHz	-100 dBc	-100 dBc	-100 dBc
Residual FM @1 GHz	1.5 Hz	2 Hz	2 Hz
Modulation	AM, FM, ϕ M, Pulse	AM, FM, ϕ M, Pulse	AM, FM, ϕ M, Pulse
Pulsed RF			
Rise/fall time	15 ns	400 ns	100 ns
On/off ratio	80 dB	40 dB to 80 dB	35 dB to 80 dB
Programming Compatibility	HP 8642A/B Fluke 6060, 6070	HP 8642A/B	HPSL
Panel Height	5 1/4 inch (133 mm)	5 1/4 inch (133 mm)	7 inch (178 mm)
US List Price	\$20,950	\$38,200	\$24,350

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The PSA-65A covers frequencies thru 1000 MHz in one sweep with a sensitivity greater than -95 dBm at narrow spans. The PSA-65A is ideally suited for 2-way radio, cellular, cable, LAN, surveillance, educational, production and R&D work. Options include frequency extenders to enable the PSA-65A to be used at SATCOM and higher frequencies, audio demod for monitoring, log periodic antennas, 10 KHz filter for .2 MHz/DIV range, carrying case (AVSAC), and more.

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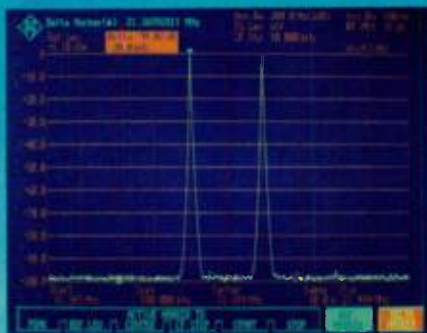
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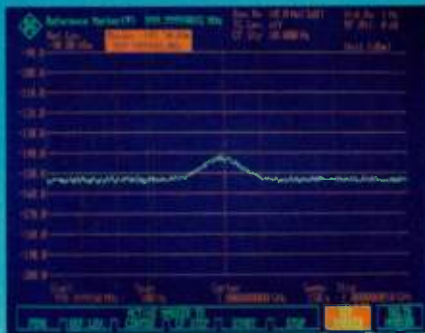
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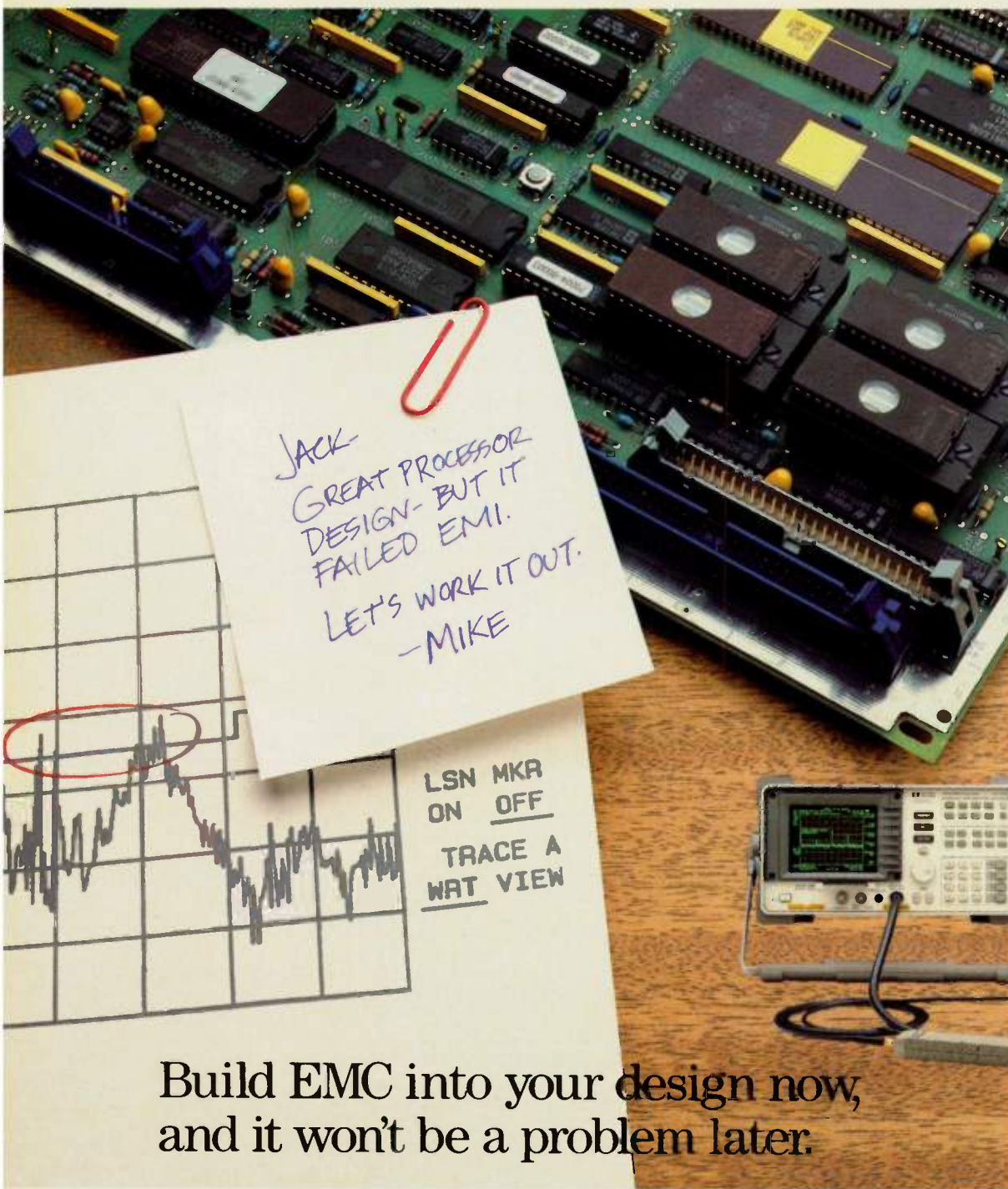
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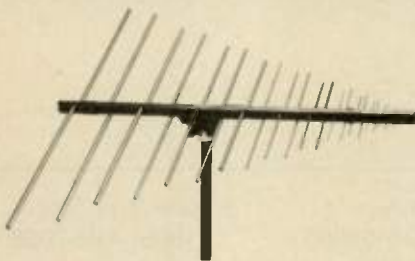
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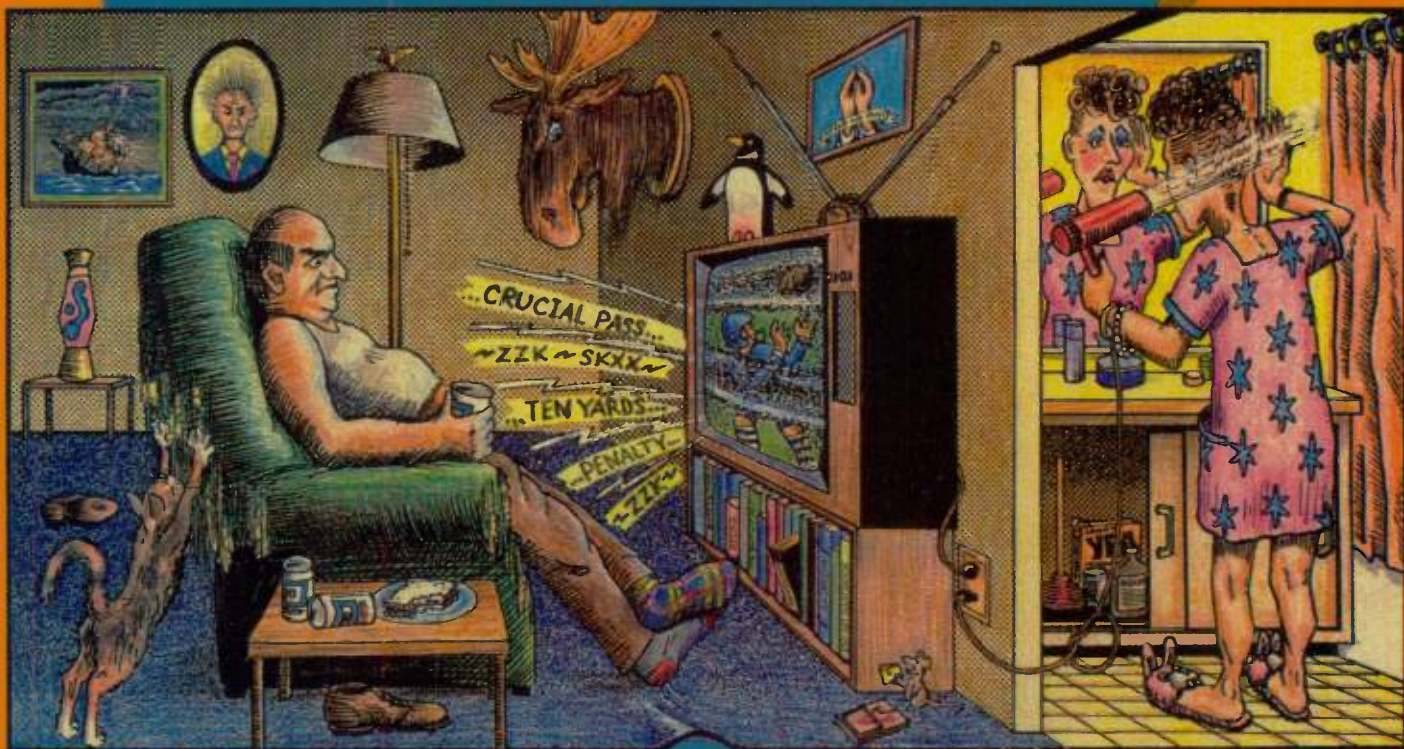
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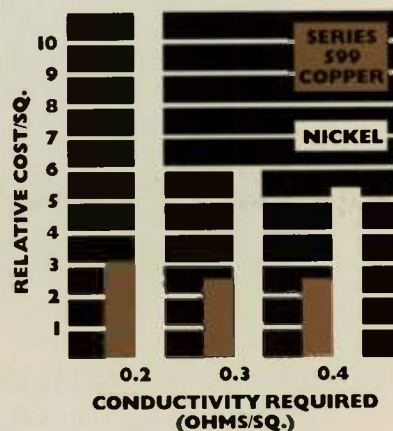
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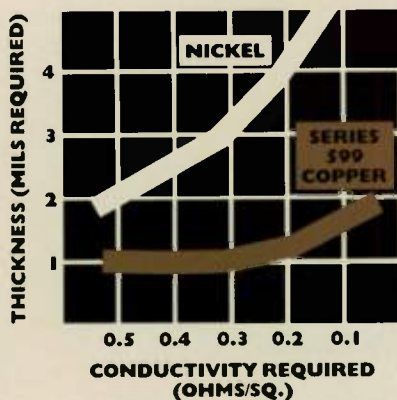
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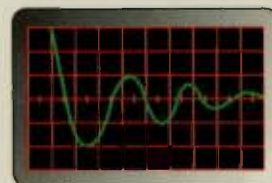
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Noise Bandwidth Calculation

By Richard Bain
E-Systems, ECI Division

When analyzing or predicting the performance of a communications system, it is sometimes necessary to determine the noise bandwidth of the filter(s) used to set the selectivity of the system. In order to predict the bit-error-rate (BER) performance of a receiver system for example, the noise bandwidth (B_n) of the IF filter and the carrier-to-noise ratio (C/N) can be used to determine energy-per-bit to noise density ratio (E_b/N_0), which in turn is used to predict the system BER. This article explains what noise bandwidth is and several methods by which it can be determined, including computer aided graphical integration.

The noise bandwidth of a filter is the bandwidth of an ideal rectangular filter response that will pass the same noise power as the filter under consideration. By definition, the passband insertion loss of the rectangular filter is equal to the insertion loss of the filter under consideration at the point of minimum loss in its passband. This relationship is shown in Figure 1. For the most common filter types, the noise bandwidth is approximately equal to the 3 dB bandwidth of the filter.

One way to find the noise bandwidth of a filter is to determine the area under the filter curve by integrating the square of its voltage transfer function. Equation 1 shows this integration expressed mathematically. Papers (1) and articles (2) have been written describing this process and giving tables of noise bandwidth to 3 dB bandwidth ratios for common filter types. However, if you are not using

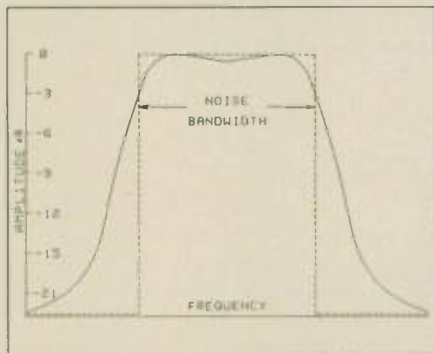


Figure 1. Noise bandwidth overlaid on filter passband.

one of the common types of filter, and you cannot construct the transfer function of the filter, then other means must be used to determine the noise bandwidth of the filter.

$$B_n = \frac{1}{K} \int_{-\infty}^{\infty} |H(\omega)|^2 d\omega \text{ rad/sec} \quad (1)$$

Where: K is the maximum magnitude of $H(\omega)^2$.
 $H(\omega)$ is the magnitude of the transfer function at a frequency of ω radians.

Another method to determine the noise bandwidth of a filter is graphical integration, i.e., dividing the passband into finite slices and summing them to determine the power represented by the area under the curve. Figure 2 shows a filter passband which is divided as just described. The amplitude of each rectangle represents the average power per Hertz under the bandpass curve over the frequency increment corresponding to the width of the rectangle. To obtain the noise bandwidth of the filter, multiply the amplitude of each increment by the width in Hertz, add the products and divide the sum by the power per Hertz at the point of minimum insertion loss in the filter passband. This process is expressed mathematically in Equation 2.

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

Where: P_{ref} is the maximum power level in the passband
 k is the number of frequency

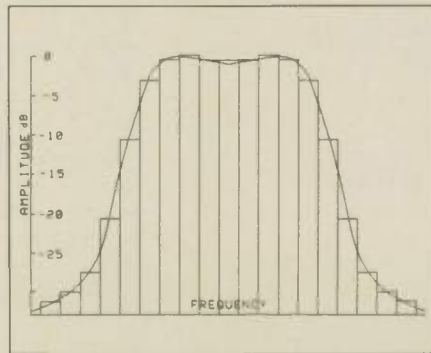


Figure 2. Graphical integration of a filter passband.

increments

f_k is the lower frequency of the increment

f_{k+1} is the upper frequency of the increment

P_k is the power density at f_k

P_{k+1} is the power density at f_{k+1}

If you have a filter in hand for which you need the noise bandwidth, and if you happen to have a computer controlled signal generator and spectrum analyzer, it is not difficult to write a program that will implement Equation 1 to determine the noise bandwidth of any filter. If you do not have this equipment, or if you have only a plot of the filter response, there is another way to obtain the noise bandwidth.

The graphical integration technique mentioned earlier can be applied quite literally to a filter response plot using a pencil, ruler, and calculator, but this is rather tedious. A program called NBW was written to speed and simplify the graphical integration process. The program requires as inputs the start and stop frequencies, the total number of frequencies, and the amplitude of the filter response at each frequency. The program then requests the filter attenuation at each frequency and calculates and displays the noise bandwidth of the filter. A program option enables the user to print out a list of the frequencies and amplitudes after the noise bandwidth calculation. Two other options allow an amplitude to be changed and the noise bandwidth recalculated. Figure 3 shows the attenuation and noise bandwidth printed out by one of the program options.

The program performs the same steps a person would when manually integrating the filter response. First, the filter response is divided into a number of equal width segments. Second, the average amplitude in absolute units of each segment is calculated and multiplied by the segment width. Third, the segments are summed. Since P_{ref} has been normalized to one, the numerical value of the sum of the attenuations is the noise bandwidth.

The accuracy of the calculated noise bandwidth is limited by the number of

N	FX(N)	ATTEN(N)	NORMALIZED
	dB	dB	ATTEN(N)
0	10.000	10.000	9.000
1	11.000	6.000	5.000
2	12.000	3.000	2.000
3	13.000	1.000	0.000
4	14.000	3.000	2.000
5	15.000	6.000	5.000
6	16.000	10.000	9.000

NOISE BANDWIDTH = 3.020263 MHz

Table 1. Frequency, attenuation and noise bandwidth printout.

increments used to calculate the noise bandwidth. The increments that the filter response is divided into are rectangular and so represent the average power in that frequency segment, which may be slightly less or more than the actual power. For a large number of very small increments, the inaccuracy becomes negligible. In most cases, 25 increments are an adequate amount. Another source of error is the truncation of the energy in the skirts of the filter. As a rule, including points down to the 25 dB

attenuation points of the filter is adequate. A filter with very broad skirts - such as a Gaussian response - might require including points down to the 30 or 40 dB attenuation points. The error involved can be checked by adding a few more frequency points and re-running the program. The increase in noise bandwidth is an indication of the size of the truncation error. It is possible to use this process iteratively with the program to determine the 99% power bandwidth of a filter as well. This is the bandwidth that 99% of the power resides in if broadband noise is being transmitted through a filter.

BER probability curves are generally plotted versus E_b/N_0 rather than C/N. As was noted in the introduction, C/N and NBW can be used to determine E_b/N_0 . This relationship is shown in Equation 3. Note that C/N is not in dB, but rather is the ratio of the absolute powers.

$$\frac{E_b}{N_0} = \frac{C}{N} (B_n)(T) \quad (3)$$

Where: E_b is the energy per bit
 N_0 is the noise density

C/N is the carrier-to-noise absolute power ratio
 B_n is the noise bandwidth
 T is the bit length

The noise bandwidth of a filter can be measured quickly and easily with computer aided measurements, but the program described in this article will ease the burden of hand calculations if such equipment is not available. It also provides a means of determining noise bandwidth if only the response curve of the filter is available. **RF**

References

1. R.D. Shelton and A.F. Adkins, "Noise Bandwidth of Common Filters", IEEE Transactions in Communications Technology, Dec, 1970
2. William A. Geckle, "Noise Bandwidth Program Eases Filter-Function Analysis", EDN, May, 1970

About The Author

Richard Bain is a Senior Engineer at the ECI Division of E-Systems. He can be reached at Box 12248, St. Petersburg, FL 33733.

```

*NOISEBN1: Program to calculate the noise bandwidth of a filter
*Uses the upper and lower bandwidth frequencies and attenuations to
*calculate the noise bandwidth. 2:1990
DIM FX(200), ATEN(200), FWR(200), ATEN(200)
RESTART: 'Come back here to re-run program
CLS
PRINT "ENTER ATTENUATION AS A POSITIVE VALUE IN dB"
PRINT
INPUT "SELECT UNITS: 1=MHz 2=kHz 3=Hz ", UNIT
IF UNIT = 1 THEN UNIT% = "MHz"
IF UNIT = 2 THEN UNIT% = "kHz"
IF UNIT = 3 THEN UNIT% = "Hz"
INPUT "ENTER LOW FREQUENCY SELECTED ", F1
INPUT "ENTER HIGH FREQUENCY ", F2
INPUT "ENTER ATTENUATION AT LOW FREQUENCY POINT ", ATDBF1
INPUT "ENTER ATTENUATION AT POINT OF MINIMUM LOSS ", ATMIN
INPUT "ENTER NUMBER OF FREQUENCIES TO BE USED ", NFRD
NINCR = NFRD - 1 'Calculate number of freq increments
PRINT
CLS
FSUM = 0
FINCR = (F2 - F1) / NINCR
ATEN(0) = ATDBF1
ATEN(1) = ATDBF1 - ATMIN
FWR(0) = 10 * (-ATEN(0) / 10)
FX(0) = F1
FOR N = 1 TO NINCR
FX(N) = FX(N - 1) + FINCR
PRINT
PRINT "ENTER THE ATTENUATION AT A FREQUENCY OF "; FX(N); UNIT%
INPUT "ATTENUATION="; ATEN(N)
ATEN(N) = ATEN(N) - ATMIN
FWR(N) = 10 * (-ATEN(N) / 10)
FWR(N) = (FWR(N) + FWR(N - 1)) / 2
FSUM = FSUM + FWR
NEXT N
'Noise bandwidth below assumes that power/increment at top is 1.0
'FSUM*FINCR gives a power bandwidth, watts/hertz or whatever unit is used
NBW = FSUM * FINCR
PRINT
PRINT "NOISE BANDWIDTH="; NBW; UNIT%
PRINT
PRINT "Come back here"
PRINT
PRINT "OPTION MENU"
PRINT
PRINT "1 = PRINT ATTENUATION VS FREQUENCY ON SCREEN"
PRINT "2 = PRINT ATTENUATION VS FREQUENCY ON PRINTER"
PRINT "3 = CHANGE AN ATTENUATION VALUE"
PRINT "4 = RECALCULATE NOISE BANDWIDTH"
PRINT "5 = START PROGRAM AT BEGINNING"
PRINT "6 = QUIT PROGRAM, RETURN TO DOE"
PRINT
INPUT "PLEASE ENTER YOUR CHOICE ": CHOOZ
IF CHOOZ = 1 THEN 'Print to screen
CLS
PRINT
PRINT "N" FX(N) ATEN(N) NORMALIZED
PRINT "UNIT:" " dB dB"
PRINT
FOR N = 0 TO NINCR 'Loop to print atten vs freq
PRINT USING "### " ; N;
PRINT USING "###.### " ; FX(N);
PRINT USING "###.### " ; ATEN(N);
PRINT USING "###.### " ; ATEN(N);
IF N = 10 OR N = 30 OR N = 50 OR N = 70 OR N = 90 THEN
INPUT "PRESS RETURN TO SHOW NEXT PAGE", DUMY
END IF
NEXT N
PRINT
PRINT "NOISE BANDWIDTH="; NBW; UNIT%
PRINT
END IF
IF CHOOZ = 2 THEN 'Print to printer
LPRINT "N" FX(N) ATEN(N) NORMALIZED
LPRINT "UNIT:" " dB dB"
FOR N = 0 TO NINCR 'Loop to print atten vs freq
LPRINT USING "### " ; N;
LPRINT USING "###.### " ; FX(N);
LPRINT USING "###.### " ; ATEN(N);
LPRINT USING "###.### " ; ATEN(N);
NEXT N
LPRINT
LPRINT "NOISE BANDWIDTH="; NBW; UNIT%
LPRINT
END IF
IF CHOOZ = 3 THEN 'Change an attenuation
INPUT "ENTER N AT WHICH TO CHANGE ATTENUATION VALUE ", N
INPUT "ENTER NEW ATTENUATION IN dB ", NEWVAL
ATEN(N) = NEWVAL - ATMIN
END IF
IF CHOOZ = 4 THEN 'Recalculate noise bandwidth
FSUM = 0
FWR(0) = 10 * (-ATEN(0) / 10)
FX(0) = F1
FOR N = 1 TO NINCR
FX(N) = FX(N - 1) + FINCR
PRINT
FWR(N) = 10 * (-ATEN(N) / 10)
FWR(N) = (FWR(N) + FWR(N - 1)) / 2
FSUM = FSUM + FWR
NEXT N
NBW = FSUM * FINCR
PRINT
PRINT "NOISE BANDWIDTH="; NBW; UNIT%
PRINT
END IF
IF CHOOZ = 5 THEN GOTO RESTART
IF CHOOZ = 6 THEN GOTO Menu 'Otherwise fall off the end
END

```

Figure 3. Computer program to calculate noise bandwidth of a filter.

Phase Difference Networks

By W.G. Beauregard
Olektron Corporation

Of the many fundamental design problems facing the electronics engineer, probably none is as difficult as that of developing networks having a constant phase difference over a broad band of frequencies. Furthermore, this is by no means a new problem. Hartley (1), in a patent dated January 19, 1925, described the use of filters having different cutoff frequencies as phase difference networks.

Throughout the ensuing years, there have been many papers covering the subject. In January 1950, Darlington (2) published a paper describing a method of design for phase tracking networks using elliptic functions. In July of 1950, Saraga (3) published a comprehensive paper on the same subject but, in addition to the elliptic function approach, Saraga also explored techniques for the algebraic solution to the problem. In May, 1969, Albersheim and Shirley (4) published a paper which appears to expand on Saraga's earlier work on algebraic methods. They demonstrated in their paper that, for the unique case of 90 degrees phase difference, the design solutions can be accomplished by purely algebraic means and presented a number of examples. Not only did this make possible the use of a computer for quick calculations of the root parameters, the calculations can be performed just as rapidly with a hand held programmable calculator. Recently, a paper by Keely (5) demonstrated the practical application of Darlington's work by providing a number of examples of network solutions using elliptic functions.

In most of these papers, the emphasis was on RC circuits, since inductors were, at best, very inefficient elements. In his paper, R.B. Dome (6) devotes a full paragraph to the problems of inductors. Keely, in his paper, does provide one example of a four pole lattice network using LC elements. It is the intention of this paper to expand on this area and to present a number of alternative circuit approaches which make LC phase tracking networks a very attractive choice for many low loss signal

processing applications.

Network Realization

Assume a pair of networks to be designed for a 10:1 band and having a four-pole characteristic.

Darlington, as an alternative to solving an equation in elliptic functions also provides the solution as a Fourier series expansion of Theta functions.

$$P = \omega_0 \frac{\cos \lambda \sigma + q^2 \cos[3\lambda \sigma] + q^6 \cos[5\lambda \sigma]}{\sin \lambda \sigma - q^2 \sin[3\lambda \sigma] + q^6 \cos[5\lambda \sigma]} \quad (1)$$

$$\lambda \sigma = \frac{\sigma 180 - \frac{\beta_a}{2}}{n} \quad (2)$$

for n odd

and

$$\lambda \sigma = \frac{\sigma 180 - \frac{180 - \beta_a}{2}}{n} \quad (3)$$

for n even

for $\sigma = 0, 1, 2 \dots (n-1)$

For $\beta_a = 90$ degrees, either expression for $\lambda \sigma$ yields the same results. $q = e^{-\pi K/K'}$ a complex parameter of the Jacobi theta function.

Further conditions of the problem are

$$\omega_0 = 1 \quad \text{normalized cutoff frequency}$$

$$\omega_1 < \omega_0 < \omega_2 \quad (4)$$

$$\omega_0 = \sqrt{\omega_1 \omega_2} \quad (5)$$

$$\omega_2 = \sqrt{10} \quad (6)$$

$$\omega_1 = \frac{1}{\sqrt{10}} \quad (7)$$

Darlington defines the modulus, k of the system as

$$k = \frac{\sqrt{\omega_2^2 - \omega_1^2}}{\omega_2} \quad (8)$$

From the theory of elliptic integrals, k is the modulus of the first kind and

$$k' = \sqrt{1 - k^2} \quad (9)$$

is the complementary modulus.

Further,

$$k' = \frac{\omega_1}{\omega_2} \quad (10)$$

For our problem,

$$k' = 0.1 \quad (11)$$

and

$$k = 0.994987 \quad (12)$$

Then, from a table of complete elliptic integrals, the value of K and K' corresponding to the values of k and k' are obtained, where:

K = complete elliptic integral of the first kind of the modulus k

K' = complete elliptic integral of the first kind of the modulus k'.

and K = 3.695788 K' = 1.57476

$$q = e^{-\pi \frac{K}{K'}} = 0.262207 \quad (13)$$

Keely presents an alternative and simpler method of calculating q as follows;

$$q_0 = \frac{1}{2} \left(\frac{1 - \sqrt{k'}}{1 + \sqrt{k'}} \right) = 0.2597469 \quad (14)$$

and

$$q = q_0 + 2q_0^5 + 15q_0^9 + 150q_0^{13} + \dots \quad (15)$$

$$q = 0.262196$$

The values of $\lambda \sigma$ for the given problem are:

$$\lambda \sigma = \frac{\sigma 180 - \frac{\beta_a}{2}}{4} \quad (16)$$

$$n = 4 \quad \beta_a = 90^\circ$$

$$\sigma = 0, 1, 2, 3$$

$$\lambda_0 = 11.25$$

$$\lambda_1 = 56.25$$

$$\lambda_2 = 78.75$$

$$\lambda_3 = 123.75$$

Then, from equation 1

$$\begin{aligned} P_0 &= 6.60526 & P_1 &= 0.59702 \\ P_2 &= -0.15139 & P_3 &= -1.67498 \end{aligned}$$

Algebraic Solution

Saraga provides, in his paper, all of the necessary information required to perform the algebraic solution to the problem. Albersheim and Shirley expand on Saraga's work, making the manipulation much easier

Saraga:

$$k = \frac{\omega_1}{\omega_2} \quad (17)$$

$$k_2 = \frac{2\sqrt{k}}{1+k} \quad (18)$$

$$k_4 = \frac{2\sqrt{k_2}}{1+k_2} \quad (19)$$

Albersheim and Shirley:

$$R = \frac{\omega_2}{\omega_1} \quad (20)$$

$$R' = \frac{\sqrt{R} + \sqrt{1/R}}{2} \quad (21)$$

$$R'' = \frac{\sqrt{R'} + \sqrt{1/R'}}{2} \quad (22)$$

It is important to point out that, while the elliptic function solution is applicable for any phase difference, the algebraic solution is useful only for the case of quadrature phase difference. Albersheim and Shirley demonstrate, by the process of pole reduction, that the expression for any odd or even pole function can be reduced algebraically to a single pole function. Restricting ourselves to even functions as before;

$$R = \frac{\omega_2}{\omega_1} = 10 \quad (23)$$

and $n = 4$.

The new reduced frequency range, R' , for the new two-pole is

$$R' = \frac{\sqrt{R} + \sqrt{1/R}}{2} = 1.73925 \quad (24)$$

and for the single pole

$$R'' = \frac{\sqrt{R'} + \sqrt{1/R'}}{2} = 1.038534 \quad (25)$$

The roots are then obtained as:

1. Single root parameter $C_1'' = 1$

2. Root parameters of the two-pole case

$$C_1' = \sqrt{R'} + \sqrt{1/R'} \quad (26)$$

and

$$C_2' = -\frac{1}{C_1'} \quad (27)$$

3. Root parameters for the four pole network are

$$C_1 = C_1' \sqrt{R'} + \sqrt{1 + (C_1')^2 R'} \quad (28)$$

$$C_2 = C_2' \sqrt{R'} + \sqrt{1 + (C_2')^2 R'} \quad (29)$$

$$C_3 = -\frac{1}{C_1} \quad (30)$$

$$C_4 = -\frac{1}{C_2} \quad (31)$$

The same technique can be further expanded to derive the root parameters for the eight-pole network

$$R''' = \frac{\sqrt{R''} + \sqrt{1/R''}}{2} \quad (32)$$

and deriving $C_1, \dots, C_8$

From our previous example of $n=4$ and $R=10$, the following values are obtained

$$\begin{aligned} R' &= 1.73952 & R'' &= 1.03853 \\ C_1' &= 2.446857 & C_2' &= -0.408687 \\ C_1 &= 6.605256 & C_2 &= 0.597021 \\ C_3 &= -0.1513945 & C_4 &= -1.674981 \end{aligned}$$

These results agree favorably with those obtained by the elliptic function solution. Figure 1 is a computer generated plot of phase versus frequency for two four-pole phase difference networks designed using the above root values and for the frequency band of 10 to 100 MHz.

Deviation from Quadrature

The degree to which the n -pole network pair deviates from the required

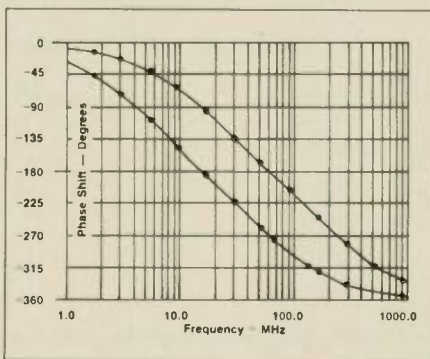


Figure 1. Plot of phase versus frequency for two four-pole, phase difference networks designed for 90 degrees phase difference over the 10 to 100 MHz band.

phase difference can be expressed in a number of ways.

From Darlington:

$$\delta = \pm 4 \sin[\beta_a] q^n \text{ radians} \quad (33)$$

where β_a = phase difference

$$q = e^{-\pi \frac{k}{k'}} \quad (34)$$

n = circuit complexity, i.e., number of poles per network. Then, from the earlier example for the 10:1 bandwidth networks

$$\begin{aligned} \beta_a &= 90 \text{ degrees} & \sin \beta_a &= 1 \\ q &= 0.262207 & n &= 4 \end{aligned}$$

and,

$$\delta = \pm 0.0189 \text{ radians} = + 1.0830^\circ \quad (35)$$

δ can also be determined directly from the bandwidth ratio, ω_2/ω_1 , and the value of n , provided that β_a is restricted to 90 degrees.

Calculating $R, R', R'', \dots$,

$$\delta = \pm \left(\arctan R' - \frac{\beta_a}{2} \right) \text{ for } n = 2 \quad (36)$$

and

$$\delta = \pm \left(\arctan R'' - \frac{\beta_a}{2} \right) \text{ for } n = 4 \quad (37)$$

etc.

Then for the four-pole case:

$$R'' = 1.03853$$

and

$$\delta = \pm 1.083 \text{ degrees}$$

Figure 2 is a plot of phase deviation versus frequency ratio for the commonly used values of $n = 2, 4$, and 8. Figure 3 shows the variation of β_a for the example of $n = 4$ and $R = 10$, specifically, for the networks of Figure 1.

LC Circuit Realization

As already noted, most of the early papers dealt with RC networks in high impedance circuits. In many signal processing applications, it becomes necessary to develop quadrature phase signals in 50 ohm systems while maintaining high return loss and low insertion loss on the networks. In general, LC networks meet these requirements. Consider the circuit of Figure 4.

Assume the following general conditions: $Z_0 = 50$ ohms, $f_1 = 10$ MHz, $f_2 = 100$ MHz. Then $\omega_0 = 2\pi \sqrt{f_1 f_2} = 1.98691 \times 10^8$.

A number circuit configurations will satisfy the requirement. We will investi-

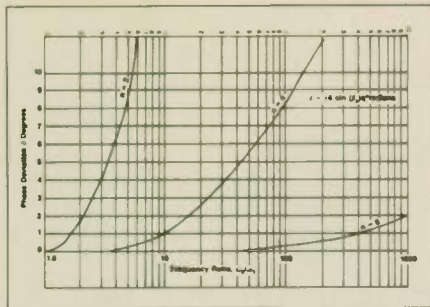


Figure 2. Phase deviation vs. bandwidth for 90 degree phase difference networks.

gate the balanced lattice and the equivalent unbalanced network.

1. Cascaded two-pole lattice network (Figure 5).

$$L_1 = \frac{R_0}{\omega_1} \quad (38)$$

$$L_2 = \frac{R_0}{\omega_2} \quad (39)$$

$$C_1 = \frac{1}{R_0 \omega_1} \quad (40)$$

$$C_2 = \frac{1}{R_0 \omega_2} \quad (41)$$

$$\omega_1 = C_1 \omega_0 \quad (42)$$

$$\omega_2 = C_2 \omega_0 \quad (43)$$

and similarly, for the second network, the elements are calculated using C_3 and C_4 .

Then

$$\begin{aligned} L_1 &= 38.1 \text{ nH} & L_2 &= 150 \text{ nH} \\ C_1 &= 15.3 \text{ pF} & C_2 &= 60 \text{ pF} \\ L_3 &= 422 \text{ nH} & L_4 &= 1662 \text{ nH} \\ C_3 &= 169 \text{ pF} & C_4 &= 665 \text{ pF} \end{aligned}$$

2. Four-pole lattice network.

The two-pole cascaded lattices can be combined to form the single four-pole lattice network of Figure 6.

Equating the elements of the four-pole lattice to the related elements of the two-pole cascaded lattices, we obtain the following:

$$L_a = L_1 + L_2 \quad (44)$$

$$L_b = \frac{L_1 L_2}{L_1 + L_2} \quad (45)$$

$$C_a = \frac{C_1 C_2}{C_1 + C_2} \quad (46)$$

$$C_b = C_1 + C_2 \quad (47)$$

Expanding these equations further,

$$L_a = \frac{R_0}{\omega_1} + \frac{R_0}{\omega_2} \quad (48)$$

$$\omega_1 = P_0 \omega_0 \quad (49)$$

$$\omega_2 = P_1 \omega_0 \quad (50)$$

$$\omega_{01} = \sqrt{P_0 P_1} \omega_0 \quad (51)$$

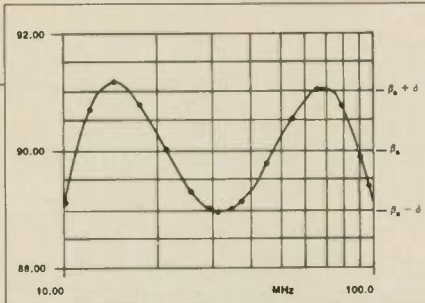


Figure 3. Phase deviation vs. frequency.

and L_a equates to

$$L_a = \frac{R_0}{\omega_{01}} \times k_m \quad (52)$$

where

$$k_m = \frac{P_0 + P_1}{\sqrt{P_0 \times P_1}} \quad (53)$$

which is a constant of the system.

Similarly,

$$C_a = \frac{C_1 C_2}{C_1 + C_2} \quad (54)$$

equates to

$$C_a = \frac{1}{R_0 \omega_{01}} \times \frac{1}{k_m} \quad (55)$$

Again, k_m is a system constant, not to be confused with modulus k or k' of the elliptic integrals

and

$$k_m = \frac{P_0 + P_1}{\sqrt{P_0 P_1}} = \frac{P_2 + P_3}{\sqrt{P_2 P_3}} \quad (56)$$

Then, for all of the elements of the four-pole lattice,

$$L_a = k_m L_{01} \quad (57)$$

$$L_b = \frac{L_{01}}{k_m} \quad (58)$$

$$L_{01} = \frac{R_0}{\omega_{01}} \quad (59)$$

$$C_a = \frac{C_{01}}{k_m} \quad (60)$$

$$C_b = C_{01} k_m \quad (61)$$

$$C_0 = \frac{1}{R_0 \omega_{01}} \quad (62)$$

The second network is calculated in the same manner, using the roots P_2 and P_3 .

Then

$$k_m = 3.626854, \omega_0 = 1.98692 \times 10^8 \quad (63)$$

$$\omega_{01} = \sqrt{P_0 P_1} \omega_0 \quad (64)$$

$$\omega_{02} = \sqrt{P_2 P_3} \omega_0 \quad (64)$$

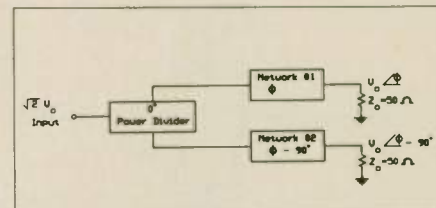


Figure 4. Typical phase tracking network in a 50 ohm system.

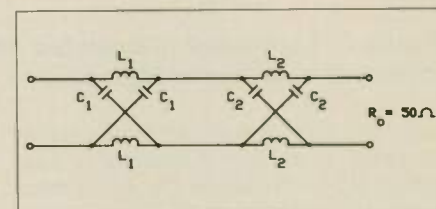


Figure 5. Cascaded two-pole lattice network.

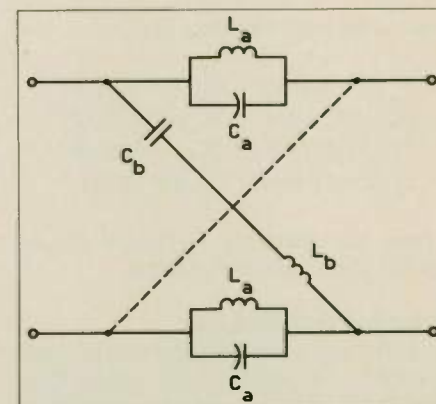


Figure 6. Four-pole balanced lattice network.

$$\begin{aligned} \omega_{01} &= 3.945646 \times 10^8 \\ \omega_{02} &= 1.0005487 \times 10^8 \end{aligned}$$

$$\begin{aligned} L_{01} &= 126.72 \text{ nH} & L_{02} &= 499.72 \text{ nH} \\ C_{01} &= 50.688 \text{ pF} & C_{02} &= 199.89 \text{ pF} \\ L_{a1} &= k_m L_{01} = 459.6 \text{ nH} & L_{a2} &= 1812.4 \text{ nH} \\ C_{a1} &= C_{01}/k_m = 13.975 \text{ pF} & C_{a2} &= 55.1 \text{ pF} \\ L_{b1} &= L_{01}/k_m = 34.94 \text{ nH} & L_{b2} &= 137.8 \text{ nH} \\ C_{b1} &= C_{01} k_m = 183.8 \text{ pF} & C_{b2} &= 724.97 \text{ pF} \end{aligned}$$

Equivalent Unbalanced Bridged Tee

In many instances, the balanced lattice network is not a convenient configuration to deal with. An unbalanced bridged tee network can be derived from the lattice network by an

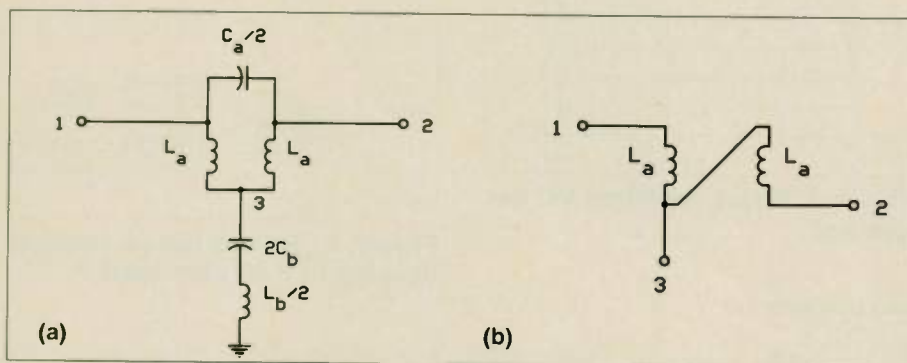


Figure 7. Equivalent bridged tee network (a) schematic representation, (b) actual transformer connection.

rem. The form presented is somewhat modified from the theoretical model presented by Guillemin (7), but it does yield very satisfactory results.

The element values of the network of Figure 7 are obtained directly from the elements of the four-pole lattice network as shown.

Values for both networks are:

$$\begin{aligned} C_{a1}/2 &= 6.98 \text{ pF} & C_{a2}/2 &= 27.6 \text{ pF} \\ 2L_{a1} &= 918 \text{ nH} & 2L_{a2} &= 3625 \text{ nH} \\ 2C_{b1} &= 367 \text{ pF} & 2C_{b2} &= 1450 \text{ pF} \\ L_{b1}/2 &= 17.5 \text{ nH} & L_{b2}/2 &= 69 \text{ nH} \end{aligned}$$

These are physically realizable values for the given frequency range.

Physical Realizability

Assuming we are dealing with lumped constant elements, there are certain limitations which must be considered when designing the networks. To varying degrees, all of the following will influence the design approach and will affect the ultimate performance of the design: frequency of operation; bandwidth; size limitations; quality of the magnetic circuit; i.e., inductor Q factor; interwinding capacity of the inductors; series inductance associated with the capacitors and inductor center taps; and parasitic capacitance to ground of any of the elements.

At the lower frequencies, i.e., from the kilohertz range to approximately 10 MHz, parasitic elements are generally not a problem. If size is not a consideration, then any of the proposed circuit configurations will yield comparable results. If size is important, then the unbalanced configuration of Figure 7 would yield the smallest package. If the networks are required to operate in a balanced circuit, then Figure 6 would

be more applicable. If because of bandwidth, the element values become too unwieldy, then the dual two pole networks of Figure 5 could improve the situation.

As the operating frequency is increased up to approximately 200 MHz to 300 MHz, the parasitic elements of the circuits have increasingly more effect. This problem is even more acute with very wide bandwidth networks. Consider the unbalanced configuration of Figure 7. The parasitic inductance of the center tap of the inverting transformer approaches the circuit value at these frequencies. The interwinding capacitance also becomes a major part of the circuit shunt capacity but, with proper design, both of these parasitic elements can be incorporated into the circuit. In general, capacitor Q being in the order of 1000 or more, has little or no effect on the performance of the circuit but care must be taken with the inductor elements. Inductor Q in the range of 25 will not seriously degrade performance. This has been proven by computer modelling as well as by many practical designs. These values can be readily obtained with wire-wound inductors built on small powdered iron toroids.

At frequencies up to 1000 MHz, the balanced lattices, fed by coaxial baluns, when required, provide very practical designs. The series inductors are usually air wound and the crossover inductors become no more than the wire leads connecting the capacitors to the circuit. At these frequencies, the dual two-pole lattices become more difficult because of the parasitic elements of the circuits, the crossover series inductance and the parasitic capacity shunting the series inductors. Using the four-pole balanced lattice design of Figure 6, these parasitic elements can easily be incorporated into the circuit.

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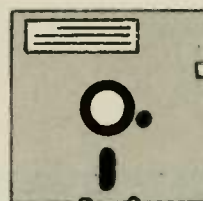
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Theorems for Today's Networks

By Dr. Harry E. Stockman

Today's students and engineers in the electric and electronic fields use network theorems as helpful tools, and in doing so often save valuable time and space. With reference to equivalent networks, this branch of electrical engineering, or rather physics, originated in Germany in 1853 with the contribution of an equivalent generator by Hermann von Helmholtz (1). The next big step forward did not occur until 1926, when Hans Mayer in Germany published his monumental paper on source transformation (2). In his paper he referred to Helmholtz' equivalent generator, without specifically mentioning Helmholtz' name. Essentially the same theorem was contributed, also in 1926, by E.L. Norton, but it only appeared in a patent communication and was not published until years later. Thanks to Mayer, today's workers in the electrical and electronics fields have at their fingertips the combination of Helmholtz' and Mayer's theorems. Extended with the concepts of initial conditions and dependent sources, amongst other things, this final or "combined" linear network theorem does just about everything in equivalent generator design.

Recently some new network theorems have appeared, providing highly useful additional network tools. A few of those will be discussed in the following with reference to networks containing both dependent and independent sources. We shall use the

simple resistive network in Figure 1 for the immediate discussion, however, the theorems are generally valid for networks containing both dependent and independent sources as well as storage elements, both in the transient state and in the steady state. Thus the network in Figure 1 may be provided with tuning elements, and if the dependent source, kl_2 , represents a stimulance (negative resistance or conductance) diode, we are actually discussing a microwave amplifier or oscillator. The conventional Kirchhoff's laws allow for the calculation of common quantities such as I_1 , I_2 , R_{IN} , R_{IN} , and R_{OUT} , as well as the transfer function,

$$\frac{I_2}{E} = \frac{1}{(1 - k)R_1 + R_2} = \frac{1}{R_3} \quad (1)$$

In the microwave field, the Classical Feedback Formula is often used, and we can then write for the actual amplification A_a , with $(R_1 + R_2) = R_{12}$,

$$\frac{I_2}{E} = A_a = \frac{A}{(1 - HA)}, \text{ or } \frac{1}{R_3} = \frac{1}{(R_{12} - H)}$$

which yields $H = kR_1$, where the feedback quantity k often is the variable with the critical value k_c for instability.

In the network in Figure 1, we wish to determine I_1 by the application of the conventional Superposition Theorem. It is immediately found that this theorem does not apply since there is only one independent source, namely E . Here the Dependent Source Superposition Theo-

rem dodges the difficulty by replacing the dependent source control variable, I_2 , by its derived inherent value I_{2i} , giving the dependent source all the characteristics of an independent source. Now we can apply the conventional Superposition Theorem because there are two independent sources. Actually, it is not necessary to use this dodge, since it would have been possible to use the unknown source kl_2 , but then there would be an unknown in the answer, and it would be necessary to start all over again. Thus, the real contribution of the cited theorem is that it derives all information from the given network right away, with the result that the answer truly is a final answer.

$$I_1 = \frac{(1 - k)E}{R_3} \quad (3)$$

As the next assignment, construct the equivalent Series-Form Generator to the network in Figure 1, looking to the right through the port. It is immediately found that the conventional Equivalent Generator Theorem does not apply since the network to be manipulated does not contain any independent sources. However, the Dependent Source Equivalent Generator Theorem teaches that indeed it is possible to construct the desired equivalent generator, but it will become a fictitious one. Again, make use of the inherent value I_{2i} , obtaining for the open-port voltage, the closed-port current, and the generator resistance

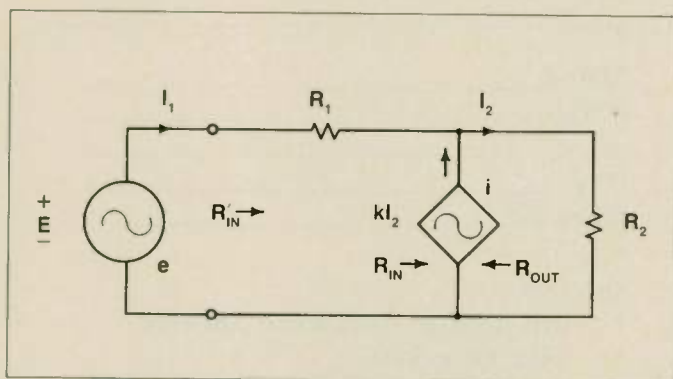


Figure 1. With tuning elements added and kl_2 a negative-conductance diode, this is a microwave amplifier.

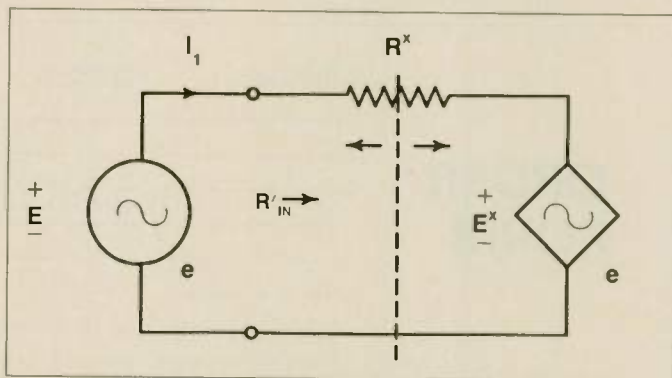


Figure 2. Partitioning at an indicated line of demarcation.

$$E^x = \frac{kR_2 E}{R_3} \quad (4)$$

$$I^x = \frac{kR_2 E}{R_{12} R_3} \quad (5)$$

$$R^x = \frac{E^x}{I^x} = R_{12} \quad (6)$$

The equivalent generator in Figure 2 is now fully determined. The relationship in equation 6 was contributed by Mayer, but Helmholtz' method could have been used to find R^x , passivating the network and looking in through the port. The network input resistance is $R_{IN} = R_1 + R_2/(1-k)$. Since I^x is available in equation 5, the Parallel-Form Generator could have been constructed as well.

Often it is required to determine the input and output impedances of a given network. The basic technique here is to use the Applied Source Method. At the input end, there is already a source and the input current should be calculated to find the input impedance. At the output, first passivate the network, and then replace the load impedance by a source, calculating the ensuing current. All this may represent considerable work. If the transfer function is known, the Port Immittance Theorem may give the two impedances almost directly. In the transfer function in equation 1, set the denominator R_3 to zero, and then solve for $-R_1$ to find R_{IN} and for $-R_2$ to find R_{OUT} . From that:

$$R_{IN} = -R_1 = \frac{R_2}{(1-k)} \quad (7)$$

$$R_{OUT} = -R_2 = (1-k)R_1 \quad (8)$$

If the transfer function is not known, only the network, the method is slightly different, and expressed by the Port Image-Immittance Theorem. To find R_{IN} , replace R_1 by $-R_{IN}$, and analyze the network for R_{IN} , obtaining equation 7. To find R_{OUT} , replace R_2 by $-R_{OUT}$ and analyze the network for R_{OUT} , obtaining equation 8.

In constructing the equivalent generators to a given network, it is common to use equations 4, 5 and 6 even if the network is extremely simple. However, for simple networks, Mayer's source transformation, when applicable, yields the equivalent generators much quicker. The technique is described by the Progressive Network-Reduction Theorem, in which the network is simply "unwound" from one end to the other. More specifically, in accordance with

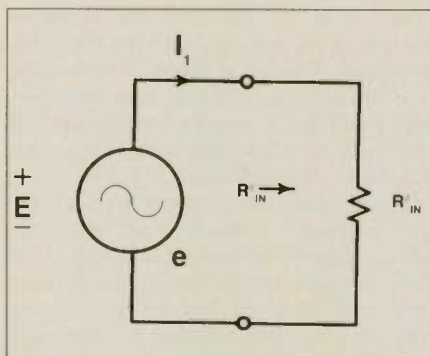


Figure 3. Use of the Compensation Theorem to find R'_{IN} .

Mayer, replace the Parallel-Form Generator (kI_2 , R_2) by the Series-Form Generator ($kR_2 I_2$, R_2) and right away we have the equivalent generator in Figure 2 with $E^x = kR_2 I_2$, which agrees with equation 4.

Many theorems are of interest simply because they provide a totally different solution, and thus, are invaluable for checking purposes. An example here is the General Immittance Theorem, derived from Tellegen's Theorem, best known from the statement that the total power sum in a given network is zero. To verify this with reference to the network in Figure 2, write the power from the two sources as $-EI_1$ and $-kR_2 I_2^2$, and the dissipated powers as $R_1 I_1^2$ and $R_2 I_2^2$. Using I_1/I_2 as obtained from equations 1 and 3, it is readily found that the summation yields zero. Time is saved instead by using the General Immittance Theorem, and with reference to the network in Figure 2, we shall claim that the total sum of all impedance is zero. Turning both sources into resistance, we obtain $-E/I_1 + E^x/I_1 + R^x = 0$, and it is easy to show that indeed this summation yields zero. Still another theorem, useful for checking, is the Partitioning Theorem, which pertains to a cut placed in the signal path, such as that indicated by the demarcation dotted line, cutting through the middle of the resistance R^x in Figure 2. The theorem states that looking left from the demarcation line one sees the negative of the power, respectively the impedance, that one sees when looking right. The theorem is readily verified.

A new theorem that saves considerable time is the Maximum Output Power Theorem, which states that once the transfer function is known, the maximum output power is predetermined and therefore also known, although the network may remain unknown. With refer-

ence to the network in Figure 1, its transfer function is given by equation 1, and the theorem then simply states that the maximum power that can be drawn from the given network is $P_{max} = R I_2^2$, where I_2 is the output current in the load R , and R the denominator of equation 1 less the load resistance R_2 . It is often desirable to "size up" the capabilities of a certain device, like an RF amplifier right away, and the maximum output power it can give is then often one of the critical evaluation quantities, and this theorem gives it almost directly. The theorem holds in general, whether or not storage elements and dependent sources are present, provided the rule for conjugate matching is observed. From equation 1 write $R = (1-k)R_1$ and obviously $I_2 = E/2R$, thus it is possible to write down almost directly $P_{max} = E^2/4(1-k)R_1$.

Instability is a key issue in amplifier design, not to mention that it is the key ingredient in oscillator design. It is surprising to find that the stability conditions can be obtained from a number of different theorems. We may ask if the network in Figure 1 can "take off." Certainly it can! Start by using the well-known Transfer Function Stability Criterion, which simply states that the denominator of the transfer function, set to zero, yields the instability conditions. Perhaps the most basic approach is that of Tellegen's theorem; the summation of all powers in the passivated network yielding the stability conditions. Still simpler, using the General Immittance Theorem with reference to the network in Figure 2, write with $E/I_1 = 0$, $R^x + E^x/I_1 = 0$, which yields the stability condition

$$k_c = \frac{R_{12}}{R_1} \quad (9)$$

The use of the Partitioning Theorem to find the stability condition merely offers a variation of above. When the Classical Feedback Formula is used, the Barkhausen Criterion, basic for the well-known Nyquist plot, yields the quickest solution with the criterion $H_c A = 1$, or $k_c R_1 A = 1$, see equation 2. If A is determined from equations 1 and 2, the result of equation 9 follows. Other very direct ways of obtaining k_c are available if the matrix and determinant solution methods are used.

Perhaps none of the common network theorems has undergone more recent changes than the Equivalent Generator Theorem. In the days of Helmholtz and Mayer, dependent sources were not in common use, thus not included in the

theorems, at least not specifically. Thus the theorems specified independent sources only. But for these networks, the new Progressive Network-Reduction Theorem generally yields the equivalent generator quicker than the Equivalent Generator Theorem, and is today the preferred way to construct an equivalent generator. Thus the old standby, the Equivalent Generator Theorem, is only of interest for dependent sources. But even here its position is being threatened, since it fails in cases of externally controlled dependent sources. Again, a new theorem is needed, and here the Anti-Coupling Equivalent Generator Theorem comes to the rescue (3). Merely replace the controlling variable by its inherent value as calculated from the undisturbed network.

One trend in modern network theorems is to further extend them into the time domain, with the idea that in the future, network theorems will be used just as often in the time domain as in the time-free domain. Even as late as 1926, the year of Mayer's revolutionizing theo-

rem, most of the common network theorems were formulated for the resistive case. Thanks to the ever increasing use of Laplace, this has now changed, and there are more and more cases of common network theorems in transient solutions. Nevertheless, if the entire problem could be readily solved without ever leaving the time domain, we would certainly give up Laplace and stay in the time domain. Reference 4 introduces a new calculus, in which the driving function, such as the step function $u(t)$, is treated as an algebraic entity. The storage elements are then interpreted in terms of algebraic switching functions, with the network result that the answer is obtained without leaving the time domain, and without the use of differential equations. The method is illustrated by means of worked problems. The solution time is about the same as for a Laplace solution. **RF**

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Oxide-Isolated Integrated Circuits for High Frequency Applications

By David W. Hughes
Georgia Tech Research Institute

High frequency integrated circuits are beginning to find increasing usage in all but the highest power systems. The insertion of such chips is expected to significantly reduce the packaging, assembly and testing cost overhead currently handicapping less highly integrated hardware. It is further anticipated that several standard, semiconductor wafer processing flows may ultimately serve as the workhorses for this marketplace.

Previous research work has pointed out that gallium arsenide (GaAs) integrated circuits are unlikely to be either appropriate or cost effective up through UHF (1). In contrast to GaAs, silicon-based circuitry deserves a serious look for high frequency systems. Of course, a careful reduction of on-chip parasitic impedance is necessary in order to achieve reasonable performance at elevated frequencies. Both dielectric isolation and oxide sidewall isolation hold promise for producing silicon chips capable of operating at the desired frequencies.

Dielectrically Isolated Integrated Circuits

Ultrahigh frequency integrated circuits are beginning to appear on dielectrically isolated silicon (2, 3). Fundamental to this process is the ability to prepare the substrate material using a process analogous to that outlined in Figure 1. These important operations rely on the use of a caustic etch to create V-grooves in <100>-oriented silicon. Subsequently, the material is oxidized to form the dielectric liner for the device islands and backfilled with a thick layer of polycrystalline silicon. Then, the majority of the original, single-crystal material is lapped away to form the indicated structure. The wafer is now available for circuit fabrication. Alternatively, at this juncture, Motorola prefers to cap the mate-

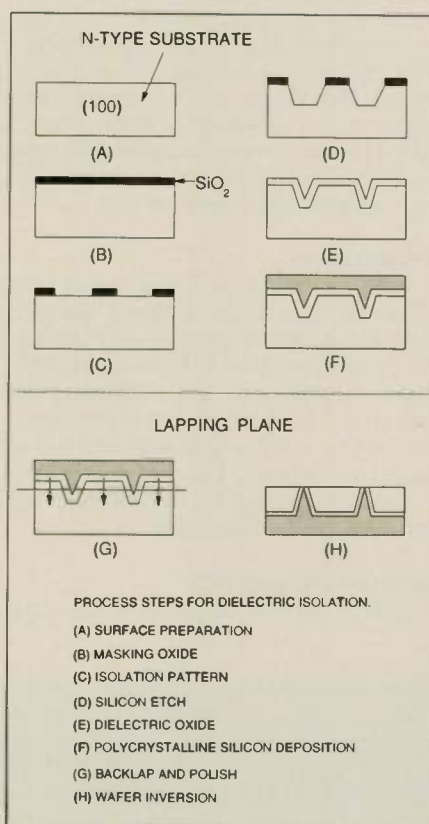


Figure 1. The basic processing steps for preparing dielectrically isolated material.

rial with an epitaxial layer for use in their dielectrically isolated, monolithic microwave integrated circuit DIMMIC™ process (2,3). In either event, standard semiconductor processing is ultimately employed to fabricate the desired circuit elements in adjacent islands. Figure 2 shows a simplified cross section through an NPN transistor formed on the DIMMIC process.

It is apparent that dielectric material features low parasitics and excellent isolation between islands and thus, provides the potential for high fre-

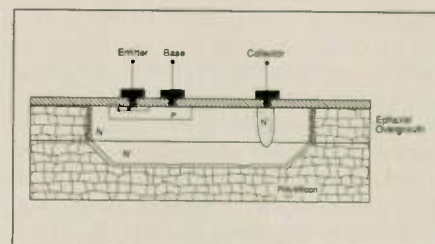


Figure 2. Simplified cross section through an NPN bipolar transistor formed on the DIMMIC process. (Courtesy of Pete Bachert, Motorola).

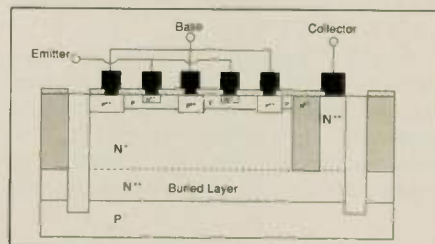


Figure 3. Simplified cross section through an NPN bipolar transistor formed on the ISOSAT process. (Courtesy of Craig Snapp, Avantek).

quency, low crosstalk, and high voltage operation. In addition, the associated confinement of any induced substrate current permits the corresponding circuit to enjoy good immunity from transient radiation upset. However, the silicon dioxide required for the island liners has only moderate thermal conductivity. Consequently, circuits formed on dielectrically isolated material will probably be limited to several tens of watts of power dissipation.

Available data show impressive operation beyond 1 GHz and it is expected that dielectrically isolated integrated circuits will find significant applications in the high frequency, low density, medium power marketplace (2).

Technology	Frequency of Operation	On-Chip Circuit Density	Power Dissipation
Dielectric Isolation	High	Low	Medium
Oxide Sidewall Isolation	High	High	Low

Table 1. Performance comparison of dielectrically isolated and oxide sidewall isolated integrated circuits.

Oxide Sidewall Isolated Integrated Circuits

Another wafer processing technology deserving of evaluation for UHF is that which relies on oxide regions to provide simple lateral isolation of the integrated circuit elements. Originally developed for digital applications, the various oxide sidewall isolation processes yield transistors which can operate well into the microwave regime. Perhaps, the earliest technology of consequence in this arena was the ISOPLANAR™ process developed by Fairchild. Other firms followed suit with their own variations and names like MOSAIC™ from Motorola are now commonly heard. AvanteK has developed a related process called ISOSAT™ specifically for high frequency integrated circuits MMICs (4,5).

Figure 3 presents a simplified cross section through a bipolar transistor formed on the ISOSAT process and illustrates the salient features of the oxide sidewall approach. Notice, in particular, the use of planar field oxide

and vertical trenches for device isolation. In addition, the emitter and base are self-aligned resulting in small devices and reduced parasitics. Consequently, the transistors enjoy extremely high cutoff frequencies for a silicon-based technology. With continued progress, it is believed that some version of the oxide sidewall isolated technologies will be useful for high frequency, high density, low power circuits.

Conclusion

It is apparent from Table 1 that both dielectric and oxide sidewall isolation hold considerable promise for the eventual proliferation of UHF integrated circuits. Furthermore, the escalating demand for UHF chips by the military, industrial, commercial and consumer sectors is expected to help accelerate the development of the requisite fabrication facilities and expertise.

Acknowledgements

It is a pleasure to acknowledge dis-

cussions with Pete Bachert and Craig Snapp during the preparation of this article. This work was partially supported by the United States Air Force, Rome Air Development Center. **RF**

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Advantest America, Inc.	71	KVG	32
Amplifier Research	79	LeCroy	17
Applied Engineering Products	49	M/A-COM Omni Spectra, Inc.	59
Avantek	29	M/A-COM Passive Component Division	51
Avcom	72	MCL, Inc.	22
Avtek Electrosystems Limited	68	Milcom International Inc.	2
Aydin-Vector	20	MITEQ	25
Bird Electronic Corp.	69	Motorola Semiconductor Products, Inc.	27
Cal Crystal Lab., Inc.	30	Murata Erie	35,37, 131-132
California Eastern Laboratories	19, 21	Penstock Engineering, Inc.	135
Coilcraft	12	Philips Components-Signetics	13-16
Communication Concepts, Inc.	9	Programmed Test Sources, Inc.	7, 42, 70
Comtran Integrated Software	62	QUALCOMM, Incorporated	40
Crystek Corporation	34	Quartztek	9
DAICO Industries, Inc.	5	Radiation Systems, Inc.	44
Electrical Insulation Suppliers, Inc.	81	Rantec Anechoic/Shielding Systems	86
The Electro Mechanics Company	78	RF Power Systems	58
ENI	136	Rohde & Schwarz	73
Hazeltine Corporation	24	SCITEQ	41
Hewlett-Packard Co.	74	SGS Thompson Microelectronics	43
High Voltage Components, Inc.	8	Spraylat	83-84
IFR Systems, Inc.	3	Stanford Telecommunications	33
ingSOFT Limited	61	STC Components, Inc.	31
Instruments for Industry	65	Surcom Associates, Inc.	9
INTECH Incorporated	41	Tecdia	8
Integrated Microwave	52	Telonic Berkeley, Inc.	47
International Crystal Manufacturing Co., Inc.	10	TRM, Inc.	57
Intusoft	61	Trilithic, Inc.	26
John Fluke Manufacturing Co., Inc.	66-67, 131-132	TTE, Inc.	53
Kay Elemetrics Corp.	46	Werlatone Inc.	45, 50, 58
		Wide Band Engineering Company, Inc.	65

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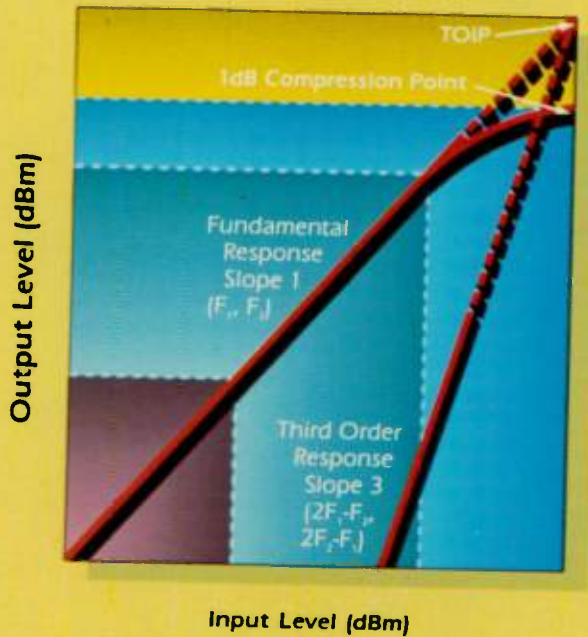
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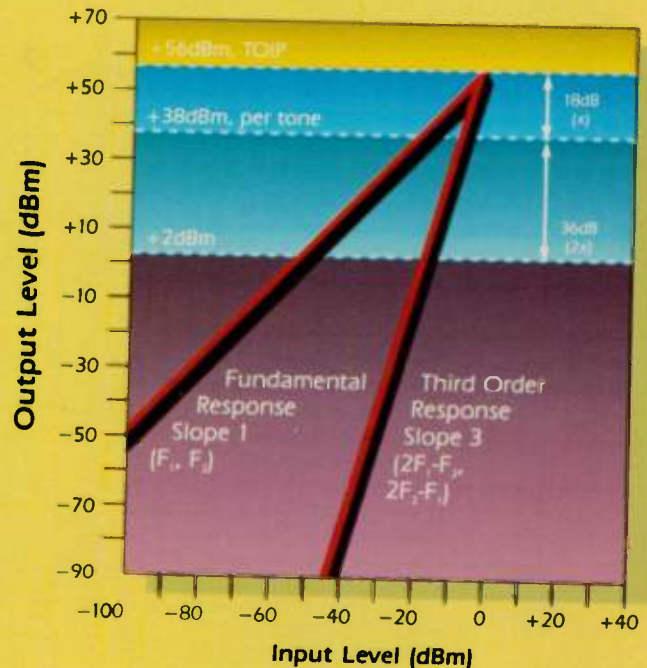
The determination of the Third Order Intercept Point (TOIP) is the best way to study an amplifier's linearity performance.

Linearity performance — the ability to accurately reproduce an input signal — is a major factor in the selection of any RF power amplifier. Usually, linearity is evaluated by measuring the amplifier's output level at 1dB compression. At ENI, we recognize that this method — a "snapshot" approach providing an indication of linearity at only **one** power level — has limited usefulness.

To evaluate linearity at **any** given power level, we measure two-tone intermodulation distortion, using the concept of the Third Order Intermodulation Intercept Point (TOIP) in the following formula:

$$\text{TOIP} = (\text{PEP Output Level} - 6\text{dB}) + 1/2 (3\text{IM})$$

We inject two signals with equal amplitude but slightly different frequencies into the amplifier. The third order intermodulation product (3IM) is the largest



Intercept plot for Model 525LA Amplifier (25W, 1-500MHz).

close-in distortion product present at the amplifier's output. The TOIP is the theoretical output signal level at which the undesired 3IM level is equal to one of the two desired output signal levels. The higher the TOIP, the more linear the amplifier.

Using the calculated TOIP of an amplifier, the relative linearity can be determined at any desired power level. To provide you with a complete picture of linearity performance, we include TOIP data with every ENI amplifier product specification.

Write or call us for our four-page Tech Note, "Evaluating Amplifier Linearity."

ENI manufactures a complete line of solid state RF power amplifiers for industrial, research, and medical applications. For more information, contact the nearest ENI office listed below:

ENITM

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