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A 1000 Watt VHF Linear Amplifier
60 Watt VHF Amplifier
HF Transistor Amplifier Design
Lowpass Filter Design

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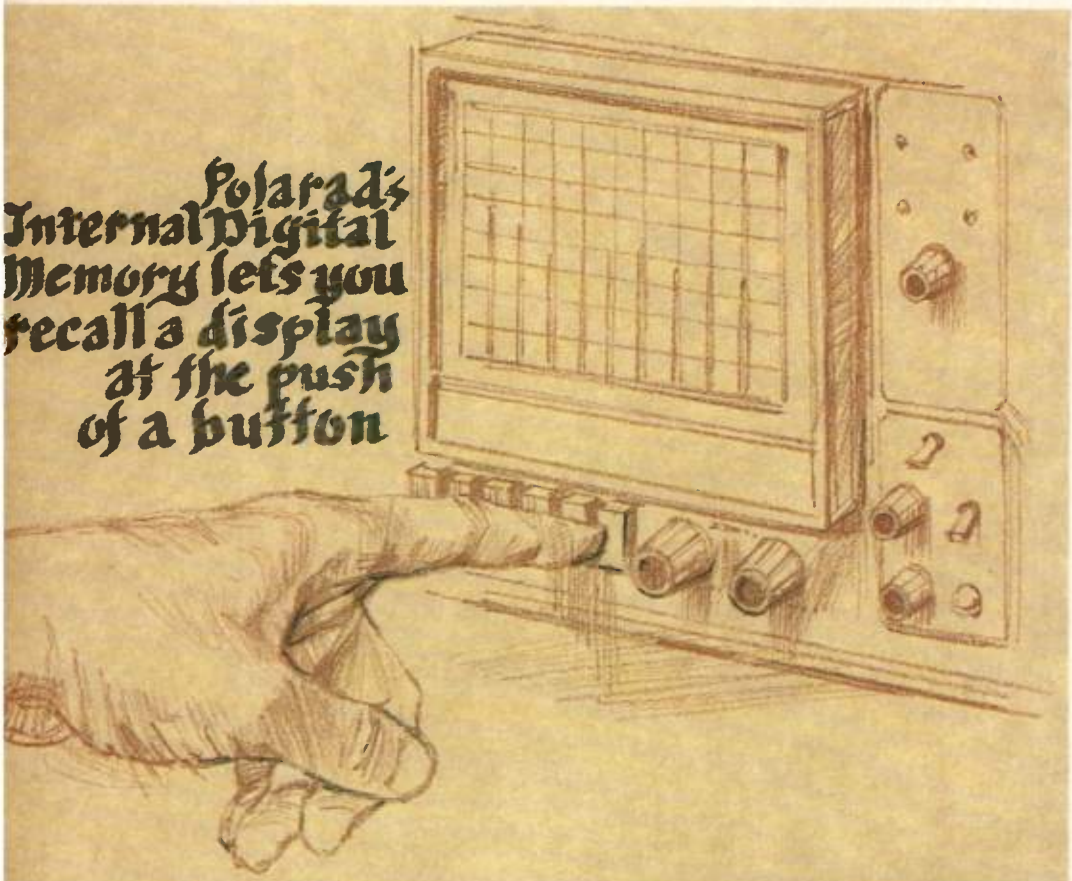


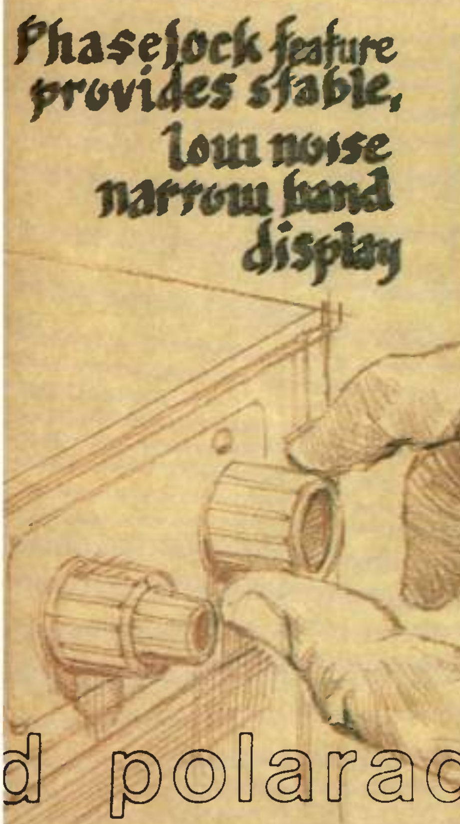


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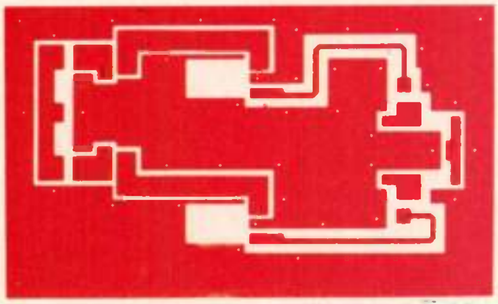
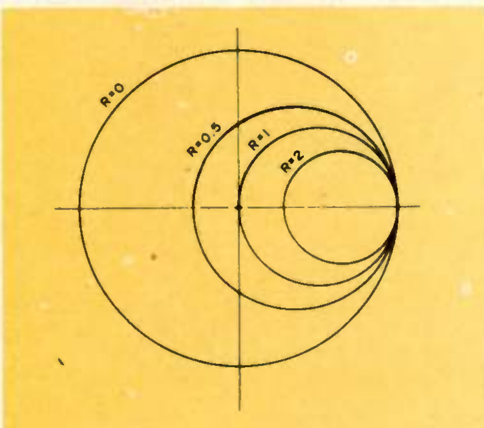


60 Watt Amplifier




HF Transistor Amplifier

Lowpass Filters


**September/October Cover** View of the RF Power amplifier module of ENI's Model 3200L, solid-state power amplifier with linear power exceeding 75 watts over a frequency range of 25 kHz to 150 MHz. **1**

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Design approaches that utilize S-Parameter, Smith Chart graphics and computer-aided techniques are considered.

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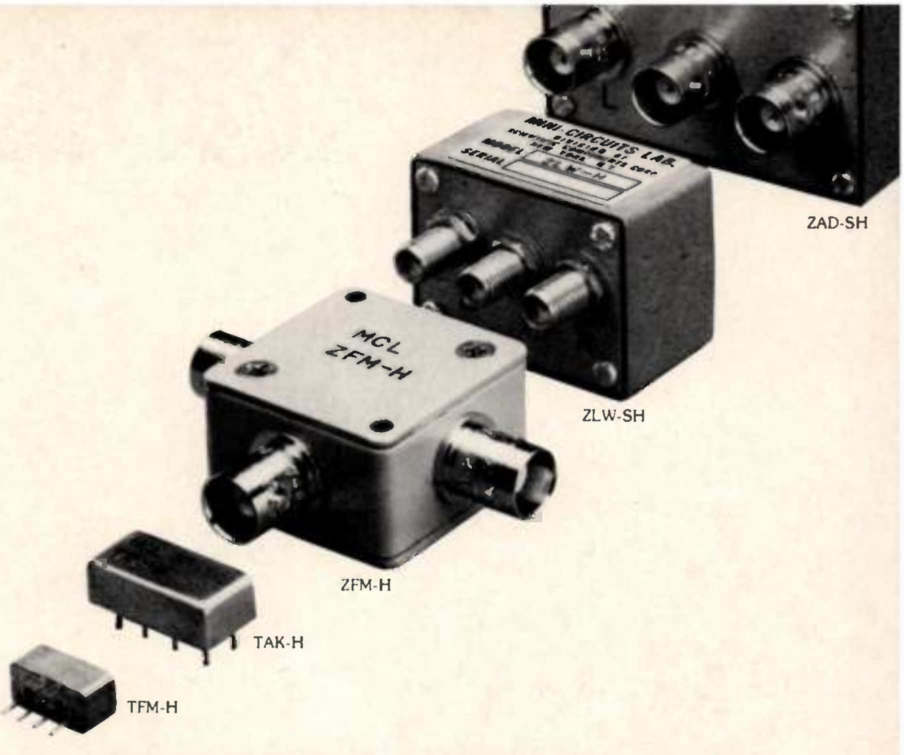
September/October 1981, Volume 4, No. 5. **r.f. design** (ISSN 0163-321X) is published bi-monthly by Cardiff Publishing Company, a subsidiary of Cardiff Communications, Inc., 3900 S. Wadsworth Blvd., Denver, Colo. 80235. (303) 988-4670.  Second Class postage paid at Denver, Colorado.  Please address subscription correspondence and Postmaster, please send PS form 3579 to P.O. Box 1077, Skokie, ILL. 60077. Subscriptions: Domestic $10, Canada & Mexico $15 per year; foreign $20 per year. Please make payment in U.S. funds only. Single copies available at $3 each.

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For subscription inquiries please contact Cardiff Publishing Circulation Service Center, P.O. Box 1077, Skokie, IL 60077



# Amplifier Design, WESCON '81

This issue has three amplifier design articles. They cover different power levels, frequency ranges and design concepts.

The first entitled "A 1000 Watt VHF Linear Amplifier, Part 2" continues from the theoretical development started in Part 1 (July-August '81) with a detailed system description. Operation of individual and combined power amplifier and control modules is explained along with performance criteria for the power supply and divider/combiner sections. Considerable detail is devoted to the Linear Amplifier's Protection System that includes VSWR, Excessive Drive Power and Over-Temperature Protection.

"60 Watt VHF Amplifier Using Splitting/Combining Techniques" is a two transistor 6 dB gain, 150-175 MHz, 60 watt output amplifier design. Complete schematic, printed circuit board photo master, component placement and performance data are included for the benefit of engineers who wish to experiment with this circuit.

"HF Transistor Amplifier Design" is a back-to-basics design article that starts off with a Smith chart review (with examples), provides a clear introduction to s-parameters and the realistic need for them in RF design and measurements, and concludes with a straightforward discussion of transistor amplifier design techniques. Included in this last category are feedback and reactive mismatch techniques.

Ed Wetherhold of Honeywell concludes his article on "Table Of Precalculated Chebyshev-Lowpass Filters with Inductive Input and Output" with his detailed construction and testing of design #31. Once again, enough design information has been included to enable any interested engineer to construct and evaluate the filter for himself.

68,024 was the acknowledged attendance figure at Brooks Hall in San Francisco for this year's WESCON '81. Our roving staff photographer, Mary Lou Norman, has scores of exhibitors and thousands of participants on film. I regret not being able to reproduce them on this page.

Finally, after 13 issues (since January, 1980) of *r.f. design* this Editor/Associate Publisher is leaving Cardiff Publishing, this being my last issue. Nonetheless, please do not doubt for a minute that my belief in the importance and relevance of the RF field has waivered. It has *not.* It has a past, present and very viable future. I take pleasure in the fact that I was able to take part in the dissemination of some RF information to 25,000+ designers over the past two years.

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Part 2: System Description and Operation.

# A 1000 Watt VHF

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*1414 Madison St.*
*Hollywood, FL 33019*

## System Description

Having discussed the ECFB operation on a per module basis, the manner in which the modules are combined as a system will be described. Figure 6 is a block diagram of the completed transmitter showing the various interconnections.

As was explained previously, each PA module has its own Control Module and the gain is adjusted for 10 dB with 5 watts of drive. Seven PA modules are combined in a 7 way combiner which has a combining efficiency of 85 percent.

Excellent phase and amplitude tracking are achieved since the output envelope of each PA is corrected to the same input by its Control Module. There is therefore negligible additional loss over the modulation envelope dynamic range due to amplitude or phase mismatch. Three hundred watts of carrier is achieved with each module delivering 50 watts.

An eighth module is used as a driver, its power being divided into seven outputs by the input divider. A 1 dB transmission line attenuator at the output adjusts the power delivered to the inputs of each of the seven output PA's to 5 watts. Since all amplifiers initially are adjusted with 5 watts of drive for a gain of 10 dB, the power levels as a system must be the same, otherwise the feedback loops will keep the output power constant but change the amplifier gain. The result of not including the 1 dB attenuator is to overdrive the final PA's with a consequent distortion on modulation peaks. This is caused by a loss of control by the ECFB as saturation is approached (as discussed earlier).

## Power Supply

Each PA module has a series voltage limiter which limits its B+ to 28V. With an input voltage of 25-30V, the limiters are fully turned on and only have a voltage drop of 1.5V, resulting in minimal dissipation. The ECFB loop prevents any power supply variation, including ripple and noise, from affecting the transmitter output. Above 30V the limiter acts as a voltage regulator and dissipation increases accordingly. A low voltage regulated supply provides power for the sensitive control circuits. This design philosophy meets the intent of the goals set forth at the initiation of the project by providing minimum regulation with adequate protection.

## Divider/Combiner

A schematic diagram of the Divider/Combiner is shown in Figure 7. A quarter-wave transmission line is employed in which a two section matching transformer transforms the 50 ohm output/input impedance down to one-seventh the value. In this way, the seven quarter-wave lines connected to the isolation star can be made with standard RG-188/U 50 ohm cable. The cables are coiled and solve the geometry problem normally associated with this type of Divider/Combiner. The matching transformer is constructed using a combination of stripline and microstrip techniques within minimum size. Some VSWR is introduced due to this technique but is tuned out with shunt inductance. The Divider/Combiner has an isolation between ports exceeding 35 dB and a return loss of 20 dB into any port across the 118-136 MHz band. Its loss over theoretical is .65 dB of which

# Linear Amplifier

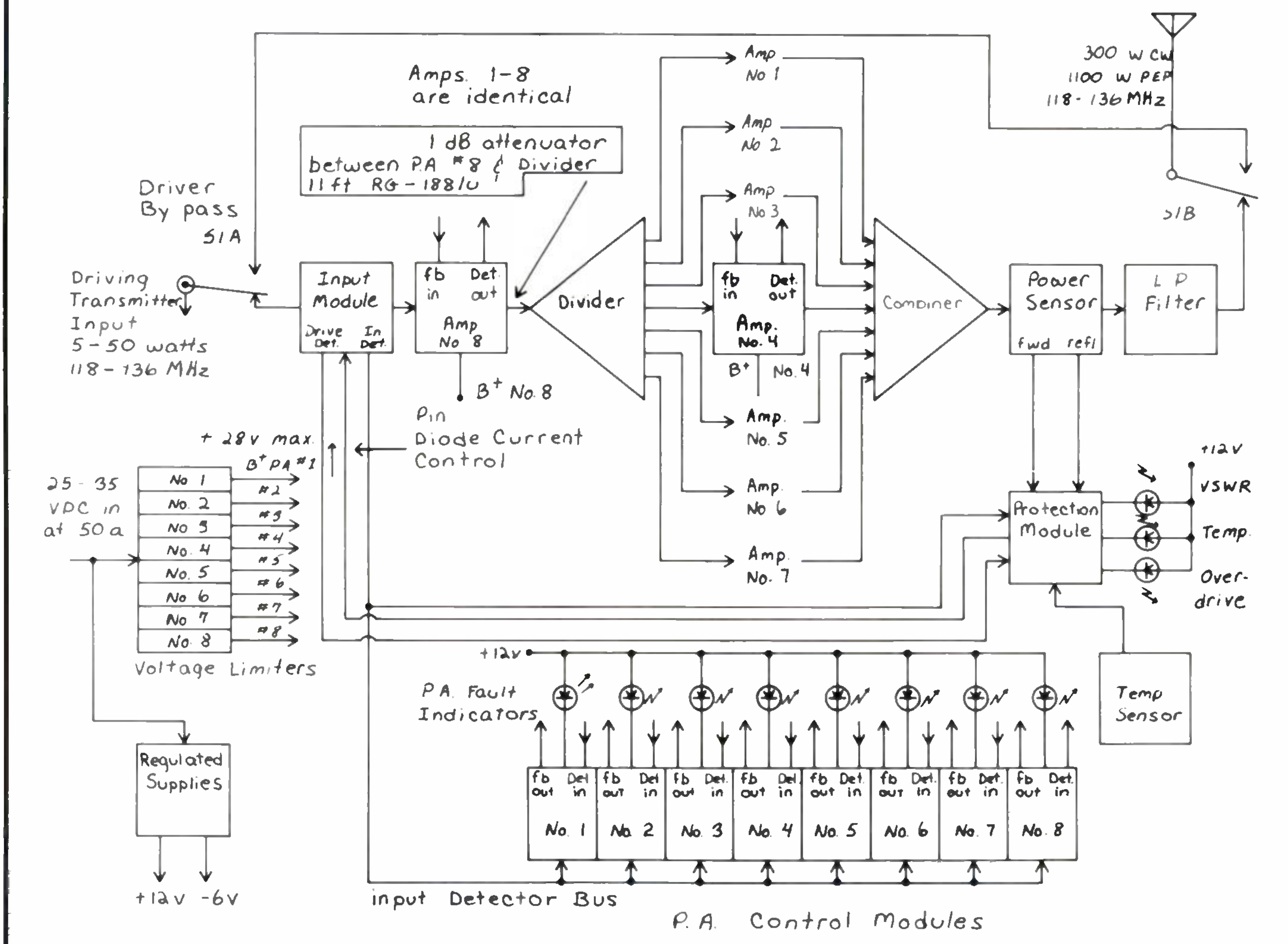


Figure 6. 1000 Watt VHF Linear System Block Diagram.

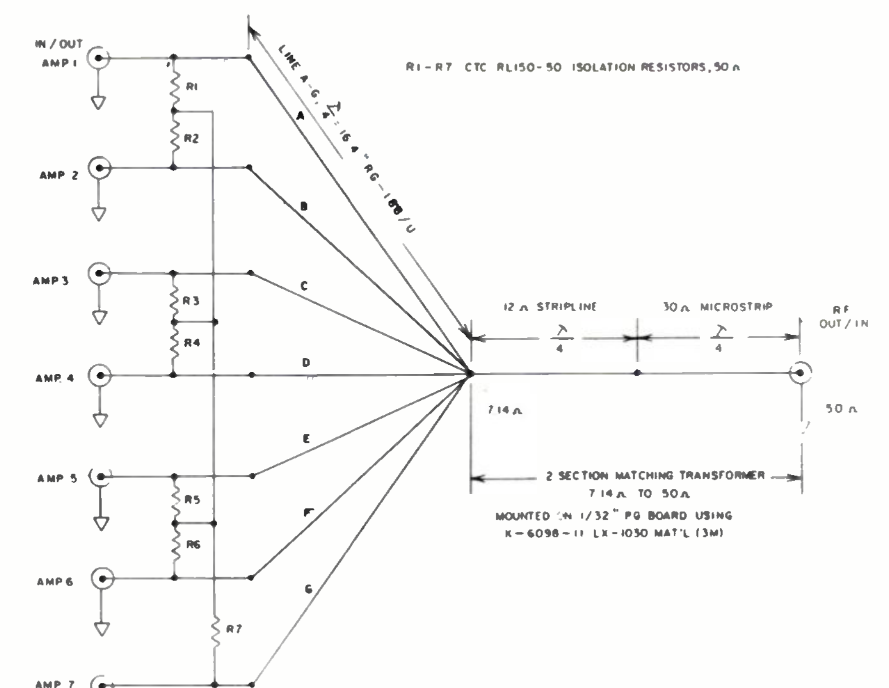

Figure 7. Combiner/Divider.

.35 dB is due to the RG-188/U cable lengths involved.

## Overall Protection System

The Linear Amplifier is protected from the following faults: 1) Antenna VSWR in excess of 3:1. 2) Excessive power from the driving transmitter. 3) Blower failure or other conditions which result in an excessive heat sink temperature.

A functional block diagram of the protection system is shown in Figure 8. The heart of the system is the pin-diode attenuator housed in the Input Module, shown schematically in Figure 3. During normal operation, the Attenuator is operated near its minimum attenuation level by current supplied by current driver Q3 in the Protection Module. The amount of current and, therefore, the degree of attenuation is determined by the setting of the $P_0$ potentiometer in the Protection Module shown schematically in Figure 9.

The pin-diode attenuator, input detector, active low pass filter (IC3)

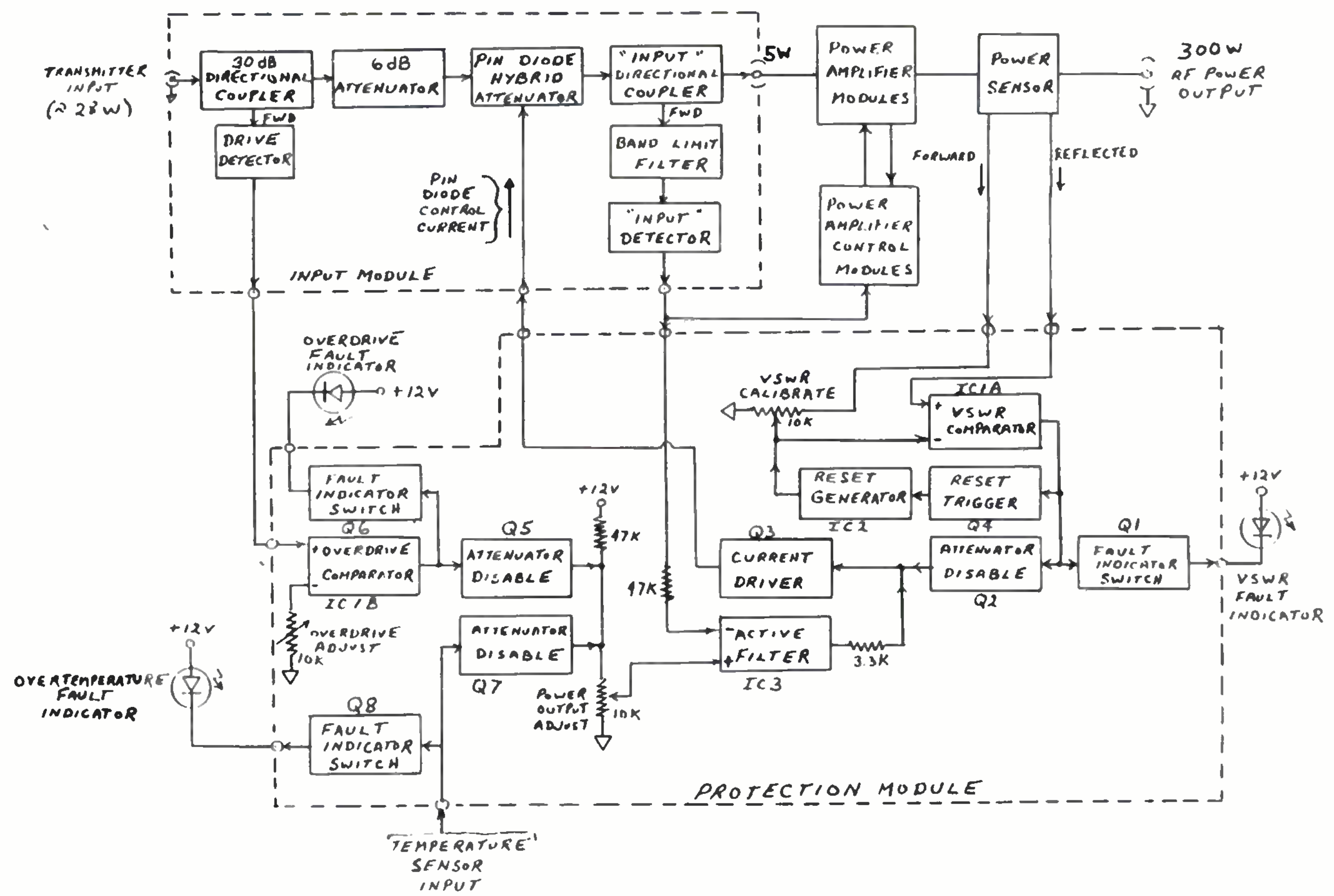


Figure 8. Protection System Functional Diagram.

and current driver (Q3), form a feedback loop which permits a linear attenuation adjustment over a 26 dB range with the $P_0$ adjustment potentiometer. This is an averaging feedback loop and does not respond to modulation or transient changes. Without this feedback arrangement, the nonlinear characteristics of the pin-diodes result in a very non-linear and critical adjustment of the output power.

While the attenuation is adjustable over a 26 dB range, the input power levels to the Attenuator must be close enough to that required to drive the transmitter to its specified output power without using more than 3-4 dB of attenuation. At attenuation levels greater than this, the attenuator diodes begin to create excessive envelope distortion. This is the reason for the 6 dB fixed attenuator which reduces the transmitter driving power to 7 watts.

The pin-diode attenuator is then used as a fine adjustment in setting the output power. Its full range can be used, however, in varying the output power for maintenance and testing purposes as long as the envelope distortion is not a factor.

## VSWR Protection

The forward and reverse outputs of the Power Sensor are connected to the inputs of Comparator IC1A, located in the Protection Module. Under normal operating conditions (VSWR less than 3:1), the output of IC1A is in its low state and Q1 and Q2 are off. When a VSWR of greater than 3:1 occurs, the positive input exceeds the negative input to IC1A and its output becomes high, thereby turning Q1 and Q2 on.

Q1 turns on the VSWR indicator light on the front panel and Q2 turns off the current flowing through Q3. This causes the pin-diode attenuator to assume its maximum attenuation state and reduces the transmitter output power to essentially zero.

A 47K ohm resistor between the output of IC1A and the positive input provides hysteresis in order to insure turn off once the 3:1 VSWR condition is exceeded. Once the VSWR is exceeded and the transmitter is turned off, there is no output power sample and IC1A is latched in its high output state. In order to provide automatic reset action, a reset technique is used similar to that used in the ECFB protection circuits. During the transition of the output of IC1A from its low to its high state, a pulse from Q4 initiates a 100 millisecond delay in one-half of IC2. At the end of this delay, a 200 microsecond reset pulse

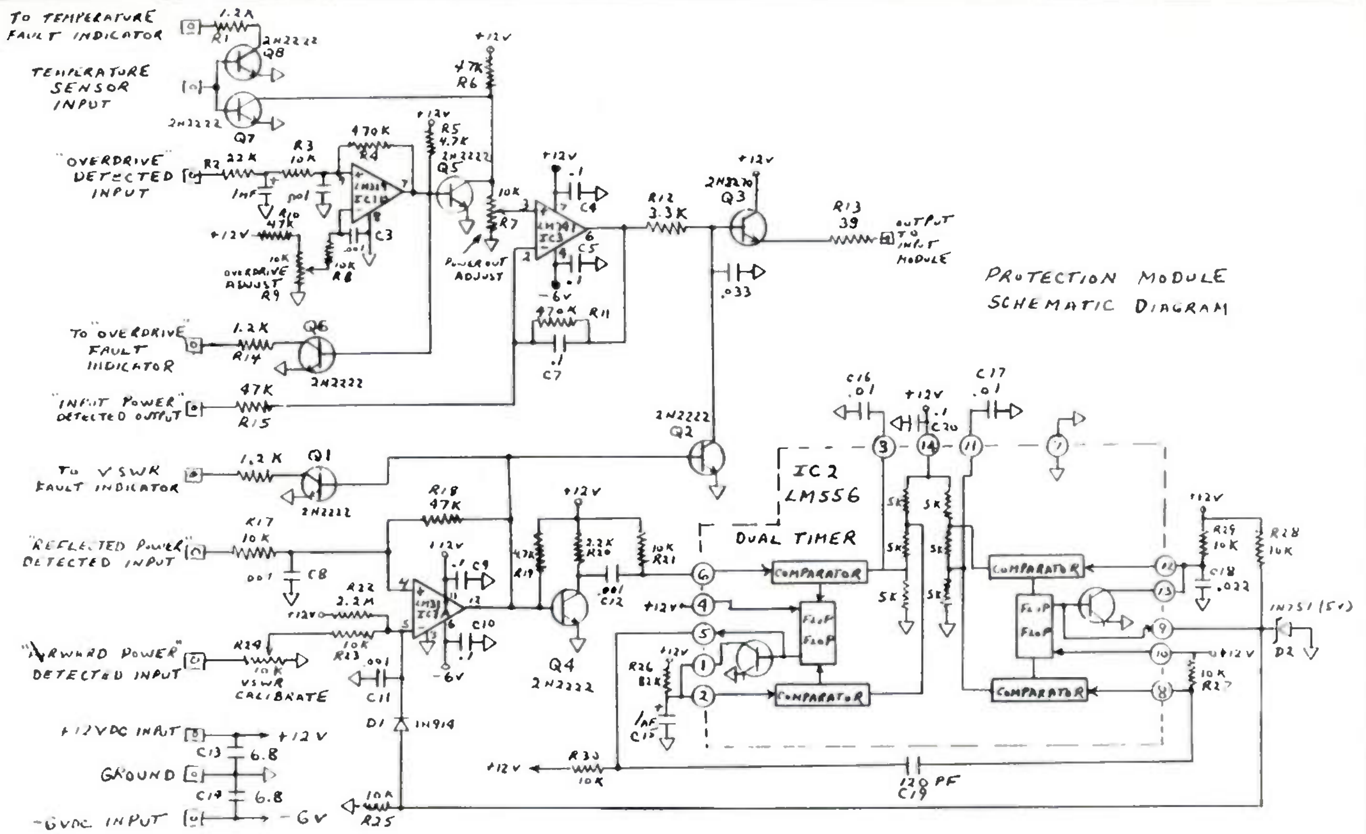

Figure 9.



is generated in the second half of IC2 and forces the negative input of IC1A positive. This thereby causes its output to change to the low state for the length of the reset pulse.

The attenuator is returned to its minimum insertion loss and the power sensor samples the condition of the VSWR. If it still remains greater than 3:1, IC1A latches in its high output position, thereby turning off the transmitter. The sampling cycle will continue until the high VSWR condition is removed, at which time normal transmitter output will be restored.

The sampling time and duty cycle are low enough to prevent transistor failures under any conditions of mismatch. The fast shut down response will protect the transistors even from such transient mismatch conditions as antenna relay actuations. This shut down technique has been used in lieu of other methods that operate on an averaging basis reducing the output power as the VSWR increases, for two reasons:

1. There are unpredictable reactions of solid-state PA's which are combined using isolated combiner techniques. Such combiners lose their isolation properties under high VSWR conditions, creating a condition whereby the amplifiers can interact, often resulting in catastrophic failures.

2. The ECFB technique results in considerable power dissipation in the transistors even if the output is reduced by as much as 10 dB. This is caused by the increase in current required to achieve linearity at low input power. By turning the drive power off with the pin-diode attenuator below the ECFB threshold point, no average current flows through the transistor except that created by the VSWR sampling scheme, which is about 0.2 percent of the normal operating current.

## Excessive Drive Power

Protection from excessive drive power is provided by Comparator IC1B located in the Protection Module. The output of the drive detector, which is housed in the Input Module, is integrated to remove the modulation and connected to the positive input of IC1B.

The overdrive adjust potentiometer is set for the amount of overdrive considered indicative of a fault. If the drive power exceeds this value, IC1B changes to a high output state, thereby turning on Q5 and Q6. Q6 turns on the front panel light indicator and Q5 causes IC3 to change to its low output state. The pin-diode attenuator is thereby placed in its maximum attenuation condition and the transmitter is turned off.

About 1 dB of hysteresis is established in IC1B to prevent erratic fault indications for drive power increases just at the fault level. This protection has been incorporated to prevent excess stress on the Linear Power Amplifier by either a failure in the driving transmitter which results in high CW drive power, or by applying a driving transmitter of too high an output power.

## Over-Temperature Protection

A temperature sensor is mounted on the Power Amplifier heat sink and its output switches to a high state when the temperature indicating a fault is reached. When this fault condition occurs, Q7 and Q8 in the Protection Module are turned on. Q8 turns on the light indicator on the front panel and Q7 causes the output of IC3 to assume its low state. Current driver Q3 is turned off, and the pin-diode attenuator is placed in its maximum attenuation position, thereby turning the transmitter off. Hysteresis is incorporated in the temperature sensor, thus allowing the heat sink to cool to a predetermined temperature before restoring the transmitter to normal operation.

## Summary

The ECFB technique has resulted in a 1000 watt linear amplifier design which fully meets the criteria established at the beginning of the project. The most difficult problems encountered were in conjunction with the protection circuits and are possibly the reason why ECFB has not been used more extensively. From this development effort it is believed that the advantages are worth the additional circuitry required to solve the problems.

The writer in no way claims this to be the only path to high power design but has simply presented one complete approach which has achieved the goals set forth. Comments and suggestions from those involved in similar work would be greatly appreciated. □

This work was performed for Aerocom, Miami, Florida and is presented with their permission.

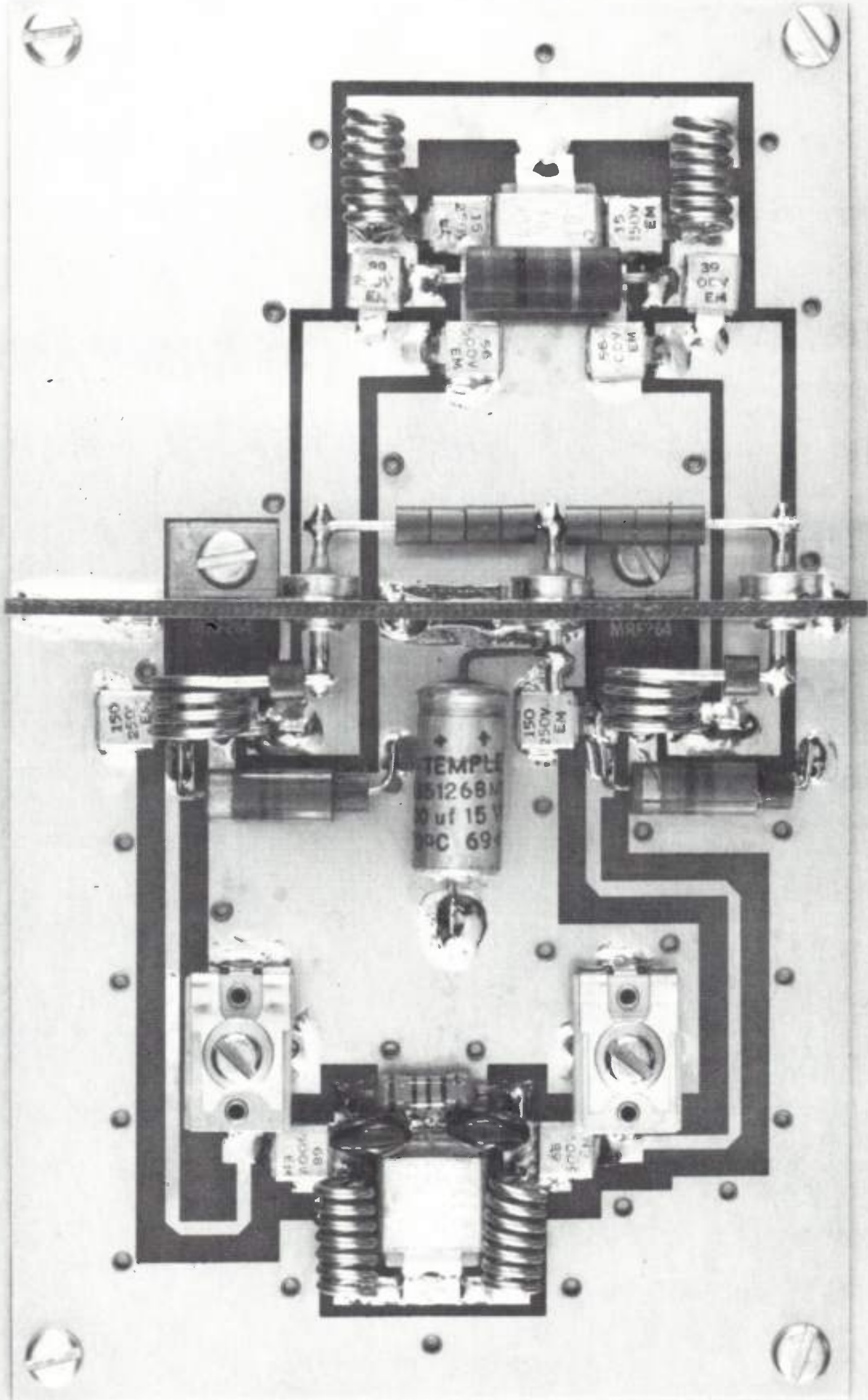

FIGURE 2 - Amplifier Layout-Top View

# 60 Watt VHF Amplifier Using Splitting/Combining Techniques *

*Ken Dufour*
*Motorola Semiconductors, Inc.*
*Phoenix, AZ*

Using proven combining techniques to obtain higher output power or added reliability at VHF can be accomplished with excellent results. Simple matching networks and power transistors featuring moderate gain capability can produce a level of performance comparable to that of a single-stage amplifier using a larger, more expensive device. Though not the ultimate answer in VHF amplifier design, the splitter/combiner method does have distinct advantages over designs that brute force the transistors into a parallel configuration. Current hogging and reduced impedance level problems associated with that technique are minimized. The exotic materials or expensive board layout required to produce a true push-pull design operating at VHF again makes combining techniques more appealing.

This 60 W amplifier operates from 150 to 175 MHz and features two, Motorola MRF264 transistors. These devices are designed for operation at VHF and individually produce 30 watts of rated output power and 6.0 dB of gain with a 12.5 volt supply. The amplifier design makes use of a modified Wilkinson combiner technique to produce 60 watts output with a drive level of 15 watts.

## Design Considerations

Experimental work with 90° (quadrature) couplers proved unsuitable for this application. Generally, they are sensitive to mismatch and tend to create instability and loss of power when used in an amplifier. In-phase (Wilkinson) couplers provide an adequate solution to this problem. They are relatively insensitive to phase changes and offer good bandwidth characteristics.

Printed transmission lines for the frequency of interest can become somewhat cumbersome on standard circuit board material. Therefore, lumped reactances (L1, 2, 9, 10 and C1, 2, 3, 14, 15, 16, Figure 5) are used to simulate 70.7 ohm 1/4 wave transmission lines, the main element in the couplers. This approach not only conserves board space, but provides a means to compensate for small variations in associated component values.

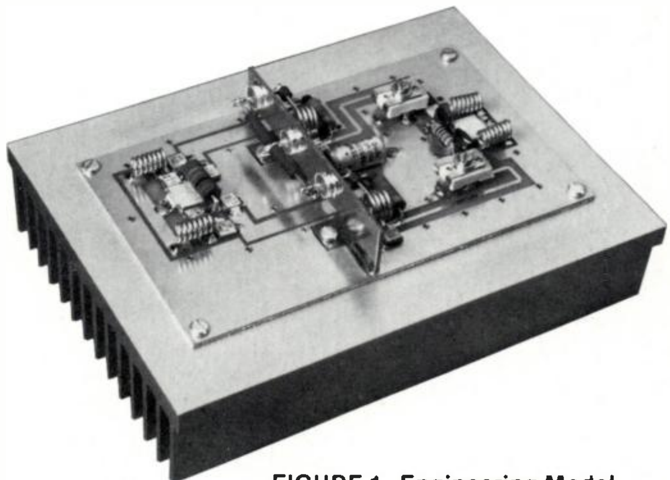

**FIGURE 1 · Engineering Model**

Microstrip techniques are incorporated in the amplifier networks to balance RF performance and promote reproducibility. Because of the lower circulating currents and reduced component heating in the collector circuitry of low-powered stages, smaller capacitors can be used in the networks at that point than would be required for a single-ended 60 watt design. Separating the major heat producing devices to two areas on the heatsink produces a more even heat transfer to the ambient air. The combined amplifier presented here has

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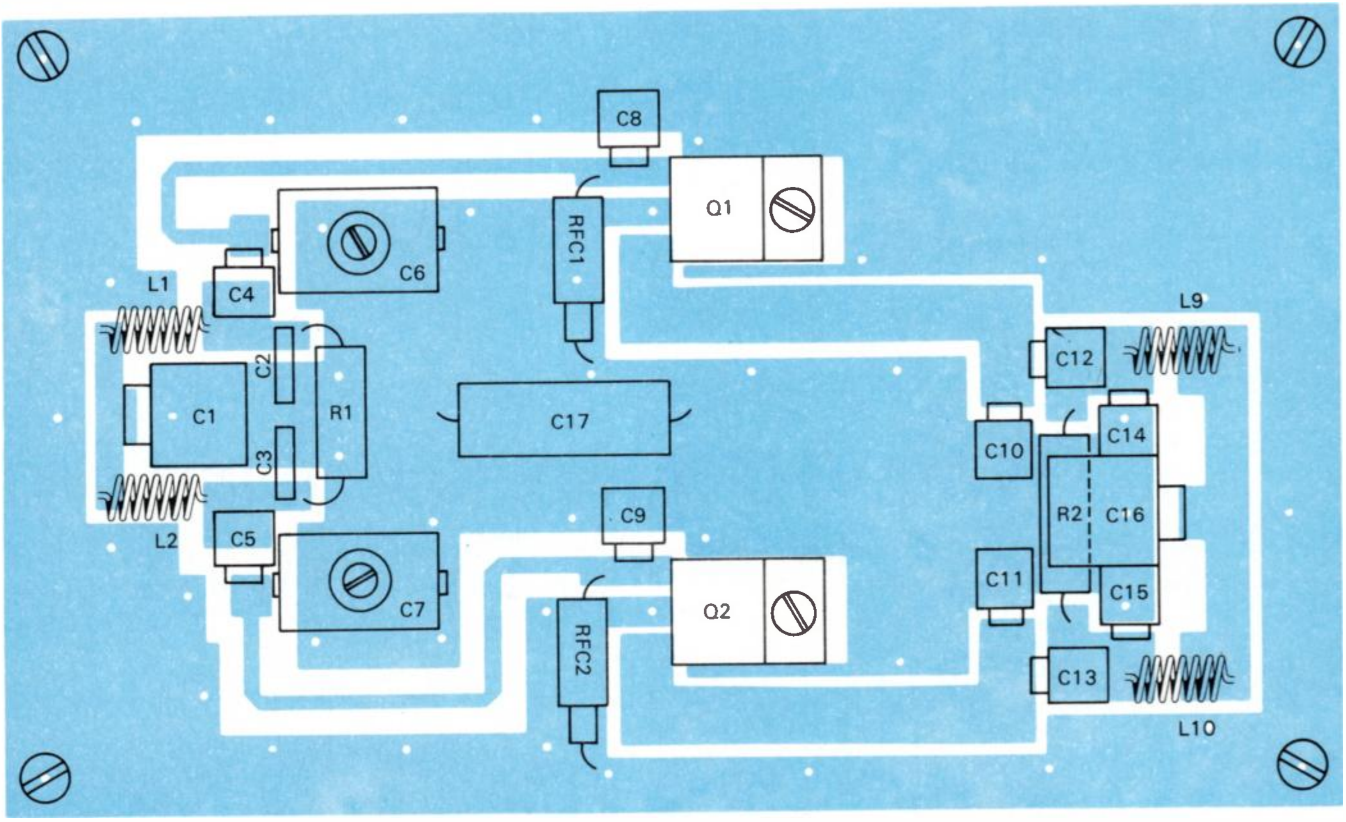
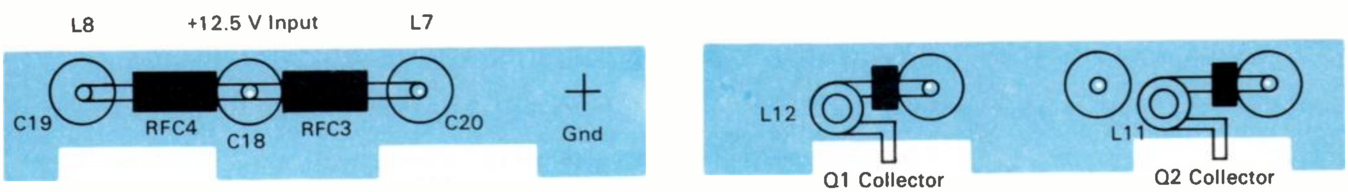



FIGURE 3 — Component Placement




a. Output Side

b. Base Side

FIGURE 4 — PCB Bridge Details

good harmonic suppression. A low-pass filtering effect is noticeable with the Wilkinson combiners.

## Construction and Alignment

A 1:1 photomask of the circuit is provided in Figure 8. Double-sided G-10 fiberglass board with two-ounce copper cladding is recommended for construction. The ground points are indicated on the PCB photomask.

The inductors required for the splitter/combiner are constructed by winding the appropriate number of turns (closewound) on a temporary 1/8 inch form and then separating the individual turns by 0.020 inch. An Xacto number 11 knife blade was used for this purpose and provides the correct turns spacing. The 100 ohm isolation resistors, R1 and R2, must be noninductive and carbon composition resistors proved to be entirely adequate. In a properly-tuned and balanced amplifier these resistors should remain fairly cool to the touch during normal operation. Each amplifier and coupler input and output port is designed to be terminated in 50 ohms to facilitate testing into a 50 ohm system.

A PCB bridge (Figures 2, 4 and 8) is used to carry all of the DC feed circuitry. It acts as a continuation of the ground plane and enhances circuit stability. Solid copper (0.027 inch) and double-sided circuit board were used as a construction medium and no difference in performance was noted with either material.

Initial alignment is accomplished by driving the

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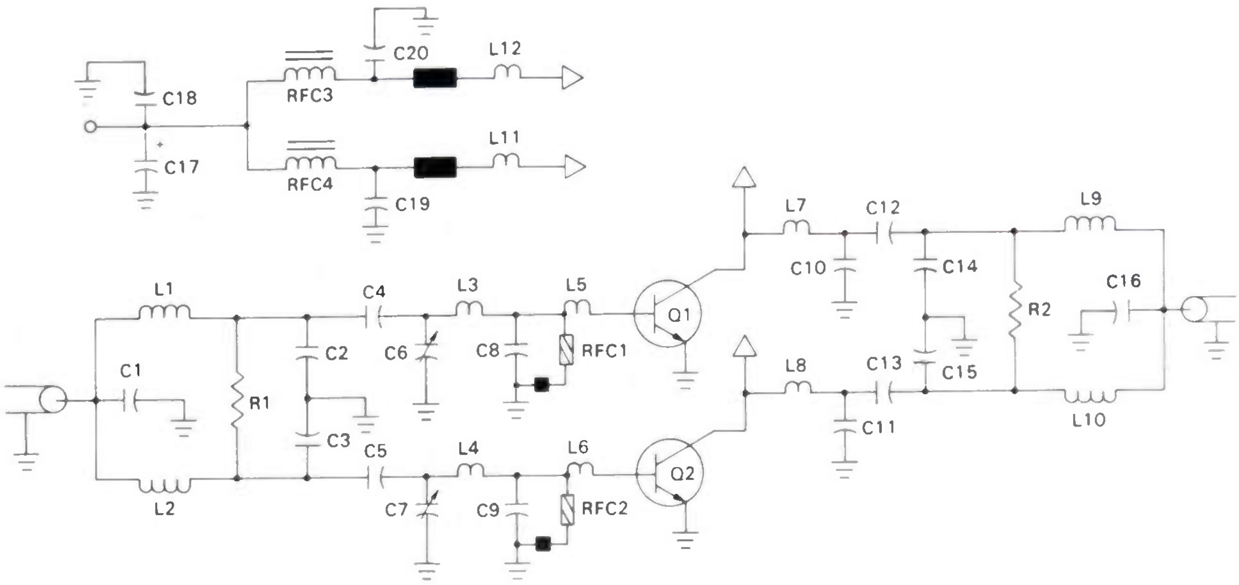



C1, C16 — 25 pF Unelco (J101)
C2, C3 — 15 pF CM04 Mica
C4, C5 — 68 pF Standex
C6, C7 — Arco 404 Variable
C8, C9 — 150 pF Standex
C10, C11 — 56 pF Standex
C12, C13 — 39 pF Standex
C14, C15 — 15 pF Standex
C17 — 100 μF @ 16 V Electrolytic
C18, C19, C20 — 680 pF Allen Bradley Feedthru

L1, L2 — 7 Turns #18, 0.125" ID
L3, L4, L5, L6 — Printed Inductors
L7, L8 — Printed Inductors
L9, L10 — 7 Turns #18 AWG, 0.125 ID
L11, L12 — 4 Turns #18 AWG, 0.250 ID w/ Bead
Q1, Q2 — MRF264
RFC1, RFC2 — 0.15 μH Molded Choke w/ Bead, Ferroxcube 56-590-65/3B
RFC3, RFC4 — 4 Ferrite Beads each on #18 AWG
R1 — 100 Ω 1/2 W Carbon
R2 — 100 Ω 2.0 W Carbon

**FIGURE 5 — Schematic - 60 W Amplifier**

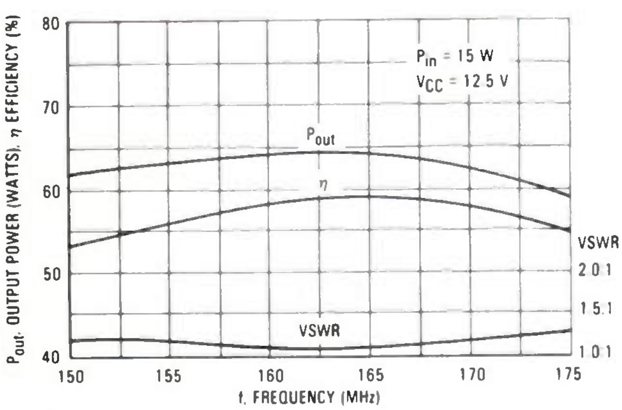



**FIGURE 6 — Output Power, Efficiency, and Input VSWR versus Frequency**

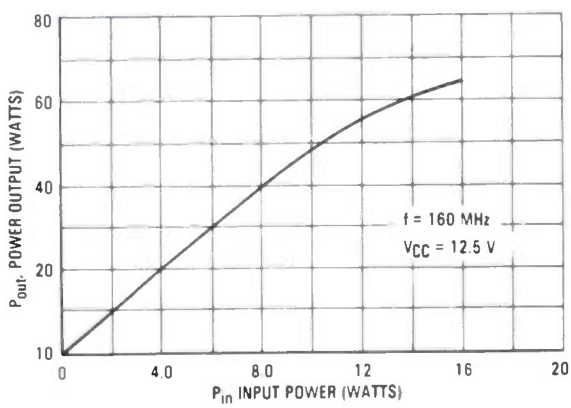



**FIGURE 7 — Output Power versus Input Power**

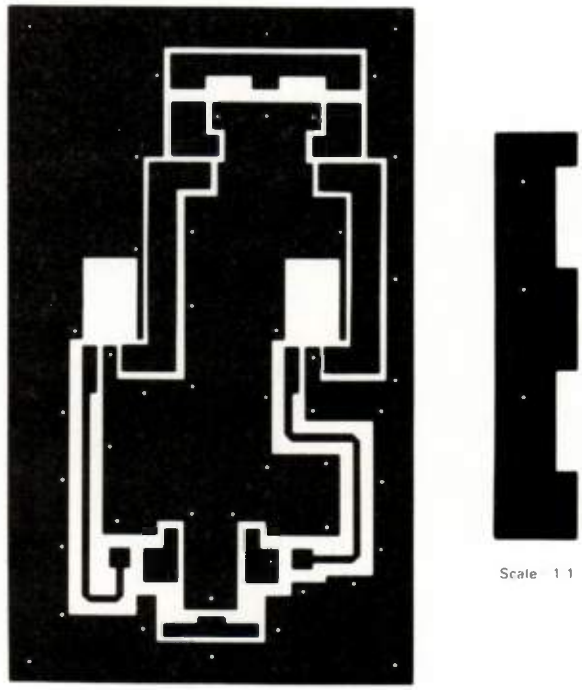
amplifier with a 5 watt CW source at approximately 160 MHz. The applied voltage is set at 12.5 volts and the variable capacitors, C4 and C5, are adjusted in an alternating manner to provide maximum output power. Full drive (15 watts) is then applied and the capacitor adjustments are repeated. At this point, the circuitry should be delivering 60 watts or more to the 50 ohm load with the 15 watts input. After the final adjustments are made, the isolation resistor temperature in either coupler should be relatively cool to the touch and the input VSWR should be at a minimum. Best results will be obtained if the transistors are beta-matched (± 10 percent prior to installing them in the circuit.

## Additional Comments

This amplifier has been extensively tested for ruggedness and reproducibility. The 15 watt input level makes it compatible with the EB-90 two-stage VHF amplifier as a driver. Together they form a chain requiring 200 mW of input power for 60 watts or more of output. ☐

### References


1. Lawrence R. Laveller; "Two Phased Transistors Shortchange Class C Amps," *Microwaves,* Pg. 48-54, February, 1978.
2. Ernest J. Wilkinson; "An N-Way Hybrid Power Divider," *PGM TT Transactions,* Pg. 116-118, January, 1960.

Scale 1:1

FIGURE 8 — PCB Photomaster

# HF Transistor Amplifier

Design approaches that utilize S-Parameter, Smith chart graphics and computer-

*Marty Jones, P.E.*
*Scientific Communications*
*Garland, Texas*

High frequency amplifier design is not magic. This article illustrates, by analysis and example some general techniques which may be adapted for immediate use in design of amplifiers throughout the RF frequency range.

The design approaches considered utilize a combination of S-parameter techniques, graphical manipulations using the Smith chart, and computer-aided design. This procedure has proven to be accurate and cost-effective.

Development and use of the Smith chart is treated in detail. This material is included because most articles on graphical design techniques seem to assume previous knowledge of the chart, as well as the graphical behavior of common circuit elements. In addition to providing algorithms for determining type and value of components, the reader is provided with an intuitive feel for the physical significance of the chart.

## Smith Chart Review

Consider an arbitrary impedance, Z. (Capital Z represents actual impedance, z represents an impedance which has been normalized to a system of characteristic impedance $Z_o$, i.e. $z = Z/Z_o$) Z may have both real and imaginary parts of either sign: $Z = \pm R \pm jX$. Figure 1 illustrates three typical circuit elements, and Figure 2 shows their impedances as graphed on a rectangular coordinate plane.

All values of Z which correspond to passive networks

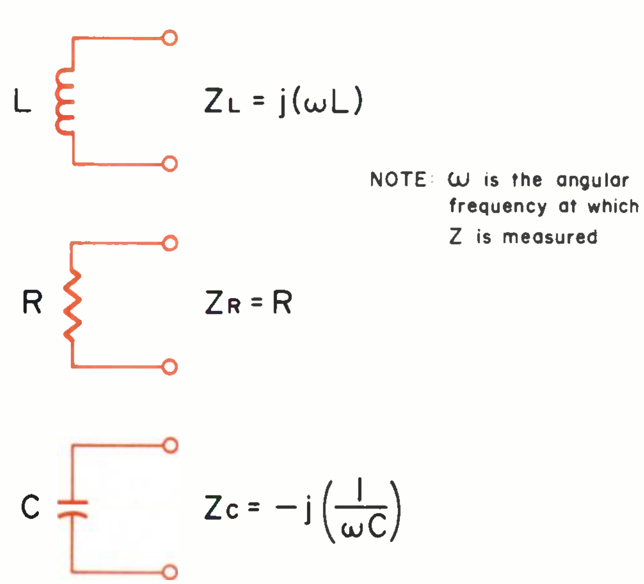



**Figure 1. Impedances of Circuit Elements**

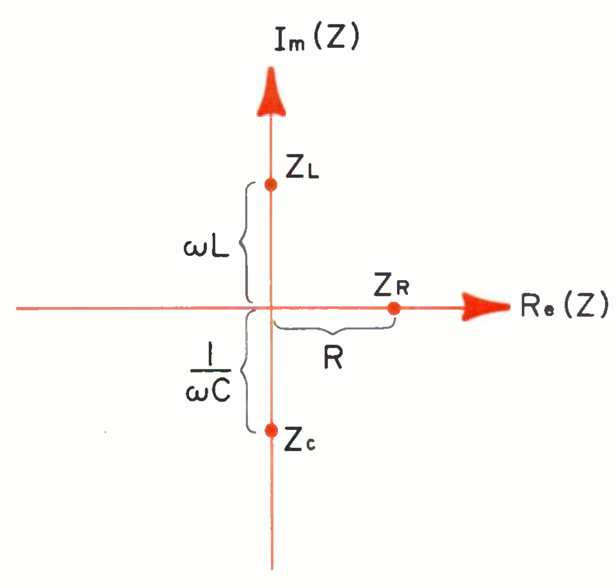



**Figure 2. Rectangular Impedance Plot**

# Design

aided techniques are considered.

are plotted on or to the right of the imaginary axis of the Z plane (since a negative real part would imply that the network is capable of supplying energy).

Utility of the rectangular plot is severely limited in that, to display the impedance of all possible passive networks, it must extend to infinity in three directions. The Smith chart overcomes this limitation by plotting the complex reflection coefficient, Γ

$$\Gamma = \frac{z - 1}{z + 1}$$

This equation shows that, for all passive impedances z, the magnitude of Γ will be between 0 and 1.

Since $|\Gamma| \leqslant 1$, the entire right half of the z plane may be mapped onto a circular area on the Γ plane. This circle has a radius of 1, and a center at $\Gamma = 0$ (corresponding to $z = 1$ or $Z = Z_0$). The following two examples graphically illustrate the transferral of impedances from the Z plane to the Γ plane.

## Example 1 — Lines of Constant Re(z)

The rectangular plot of Figure 3A shows four lines of constant resistance. For example, any impedance with a real part Re (z) = 1 will lie on the R = 1 line. Those with an inductive component of z will fall above the real axis, while capacitive impedances will fall below. Figure 3B illustrates the location of these impedances after mapping onto the Γ plane. Notice that lines of Re (z) become circles of Re (Γ). Inductive impedances are still transferred to the portion of the circle above the horizontal axis and capacitive impedances to the portion below. The major difference is that the lines no longer extend to infinity. The infinity points now all meet on the Γ plane, at a distance of 1, to the right of the origin. This implies that $\Gamma = 1$ for $z = \infty$; whether real, inductive, or capacitive. Substituting $z = \infty$ and $z = \pm j\infty$ into the defining equation for Γ,

$$\Gamma = \frac{z - 1}{z + 1}$$

verifies that this is indeed true.

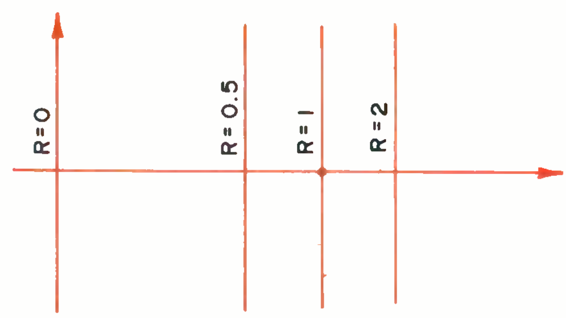



Figure 3B. Circles of Constant Re (Γ)

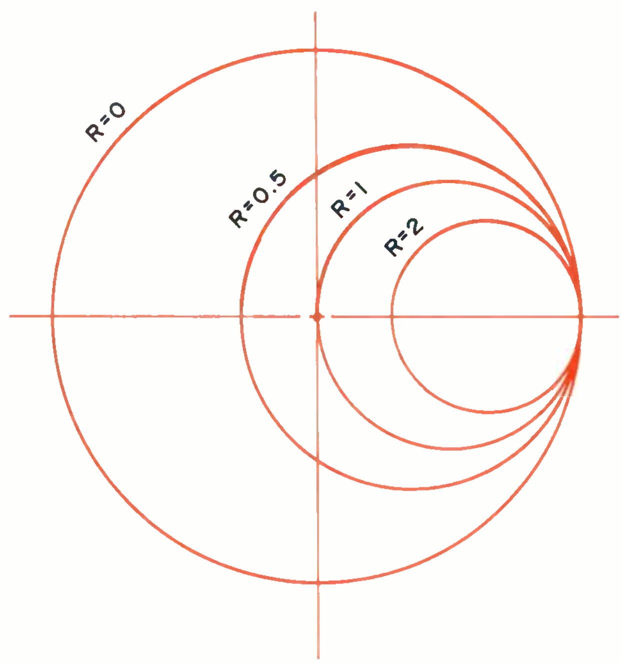



Figure 3A. Lines of Constant Re (Z)

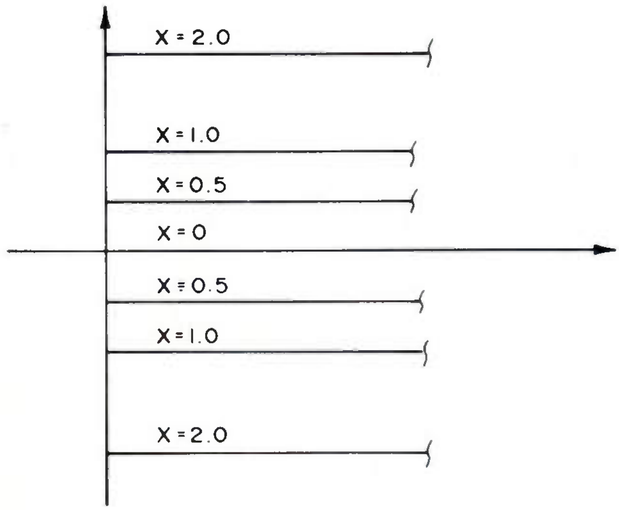



**Figure 4A. Lines of Constant Im (Z)**

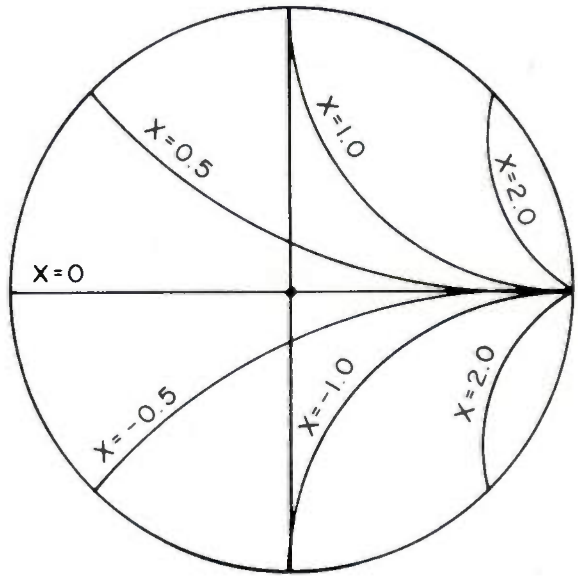



**Figure 4B. Curves of Constant Im (Γ)**

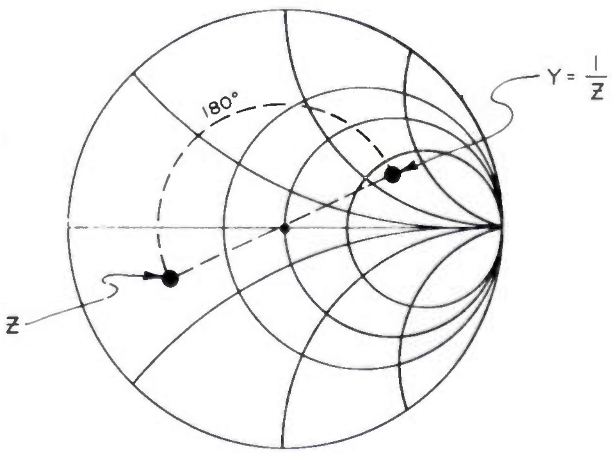



**Figure 5. Impedance Inversion**

## Example 2 — Lines of Constant Im (z)

Figures 5 and 6 similarly illustrate the behavior of Lines of Im (z) when mapped onto the Γ plane. Again we see the points of infinite magnitude meet at Γ = 1. We can visualize the entire rectangular z plane "curling" to the right, and its three axes (which previously extended infinitely) meeting at the interesection of the |Γ| = 1 circle and the horizonal Γ axis. The left half-plane values of z are, of course, mapped into the area outside the unit circle on the Γ plane. Having developed our chart, let us now review some basic graphical techniques.

## Impedance Inversion (Admittance)

The mathematical inverse, y = 1/z of any impedance (or, more generally, of any complex number) may be found graphically using the Smith chart. Plot z on the complex Γ plane. Rotate this point by 180° about Γ = o. and read the corresponding y from the coordinates of the stopping point (see Figure 5). By rotating every point on the chart by 180°, we can develop a second set of coordinates which are an inverted mirror image of our original chart. Frequently this set of admittance coordinates will be superimposed on the same chart as the impedance coordinates (Figure 6). Using this combination chart, it is no longer necessary to invert impedances by rotation. Once Γ is plotted on the combination chart, either z or y may be read directly from the proper set of coordinates.

## Complex Conjugate

The complex conjugate may be easily determined by

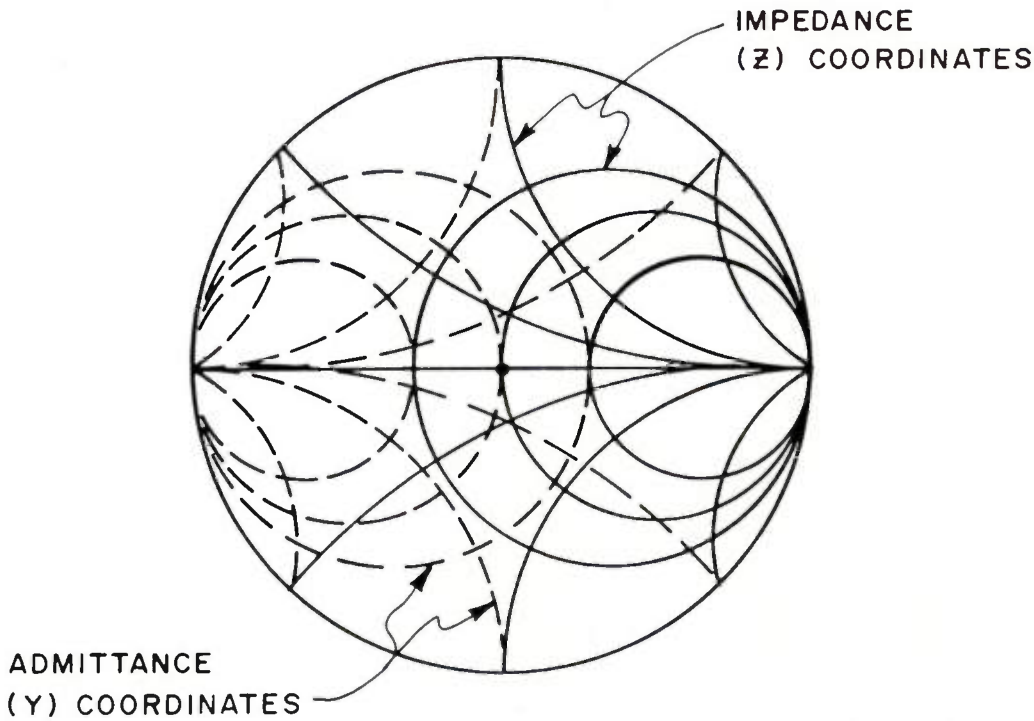



**Figure 6. Combination Impedance/Admittance Chart**

graphical means. This involves merely reversing the sign of the angle of Γ (note: Γ is usually written in polar form.) On the Smith chart, angles become more negative (phase lagging) as we rotate clockwise. 0° is on the right end of the real axis and −180°/+180° on the left.

As an example, consider $\Gamma = 0.5 \angle +150°$. The complex conjugate, $\Gamma^*$, is therefore $0.5 \angle -150°$. In Figure 7 we see that $\Gamma^*$ is found by "mirroring" Γ about the real axis.

## Impedance Transformation And Matching

We can now plot a given impedance on the Smith chart. Our next step will be to learn to use the chart in graphically determining circuit elements to modify these impedances. For example, we know that the impedance of a 25 ohm resistor is plotted at the intersection of the R = 0.5 circle and the X = o line. (We are normalizing to $Z_0$ = 50 ohms, so 25 ohms/50 ohms = 0.5.) What happens if we place an inductor in series with the resistor? Assume that the inductor has a reactance of +j25 ohms at our measurement frequency. Referring to Figure 8, we see that the real part of the series impedance remains unchanged at 0.5. We move along the circle representing constant R = 0.5 until we reach the point where X has increased by j0.5 (+j25 ohms/$Z_0$ = j0.5). This point is our new reflection coefficient.

The manner in which lumped circuit elements graphically modify impedances may be stated in a simple set of rules:

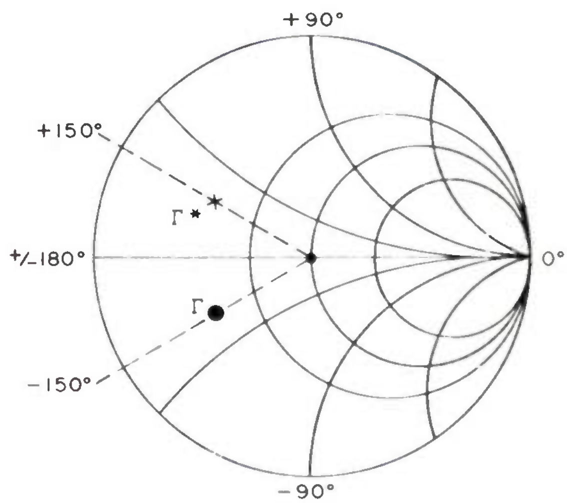
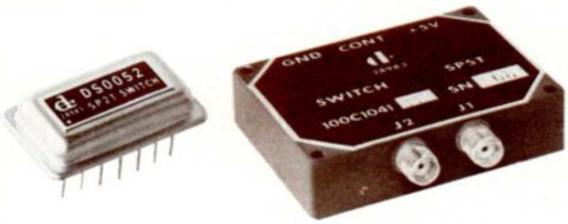



**Figure 7. Complex Conjugate**





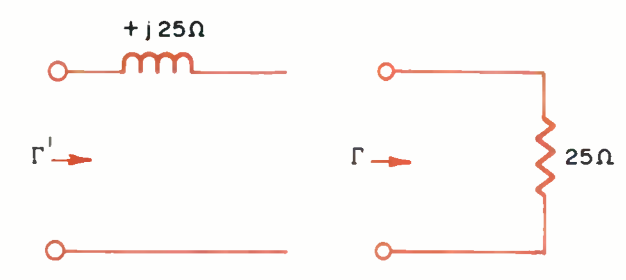


Figure 8A.

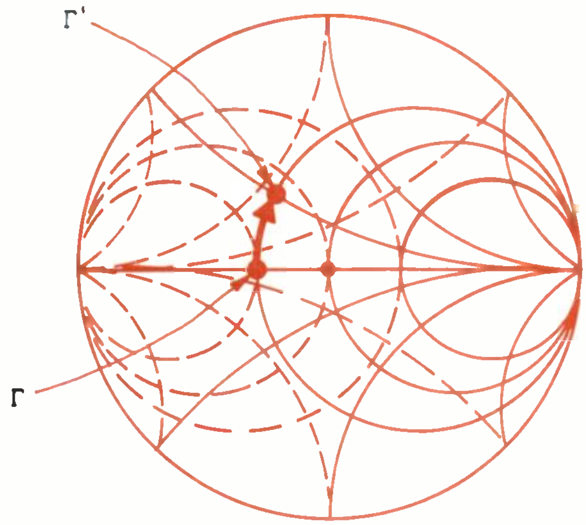


**Figure 8B. Impedance Transformation**

1. Series elements use impedance coordinates, shunt elements use admittance coordinates.

2. Inductors transform toward the upper portion of the chart along circles of constant Re (z) or Re (y). Capacitors transform toward the lower portion of the chart along these same circles.

3. Series resistors transform along constant Im (z) curves from left to right. Shunt (parallel) resistors transform along constant Im (y) curves from right to left.

Frequently we have the need to transform an arbitrary impedance to some other specified impedance such as $Z_o$. To design the transformation network graphically, we proceed as follows: First plot the beginning impedance on the combination impedance/admittance chart. Next find a path, along some combination of circles and curves, from this point to $Z_o$ (center of chart). Using the three rules previously given, determine the type and connection configuration of the elements required to transform along this path. By noting from the chart the change in normalized reactance, resistance, susceptance, or conductance ($\Delta X$, $\Delta R$, $\Delta B$, $\Delta G$), we can then calculate the actual element values. Formulas for calculating the element values from the normalized changes are given below (for $Z_o = 50$ ohms):

## Impedance of Series Elements

Resistor: R (ohms) = $50 (\Delta R)$

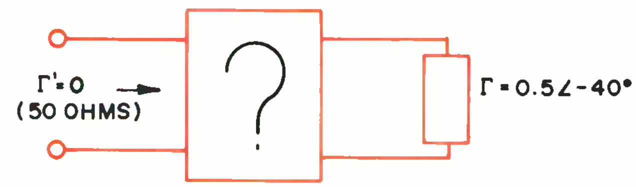



Figure 9A.

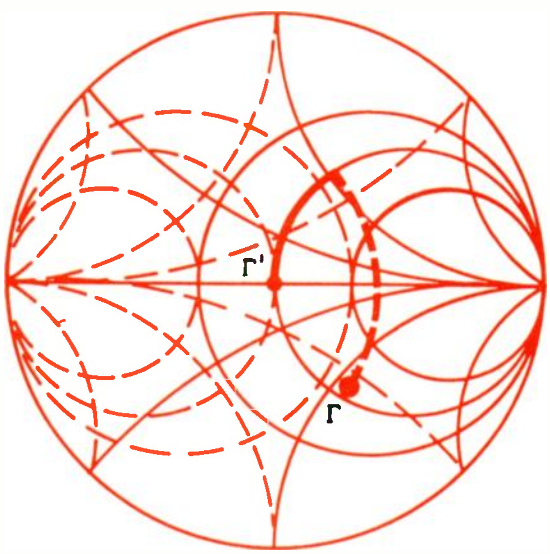



Figure 9B.

Capacitor: $X_C$ (ohms) = 50 ($\Delta$X)
Inductor: $X_L$ (ohms) = 50 ($\Delta$X)

## Impedance of Shunt Elements

Resistor: R (ohms) = 50/$\Delta$G
Capacitor: $X_C$ (ohms) = 50/$\Delta$B
Inductor: $X_L$ (ohms) = 50/$\Delta$B

## Element Values

Resistor: R (ohms) = R
Capacitor: C (farads) = $1/2\pi f X_C$
Inductor: L (Henries) = $X_L/2\pi f$

As an example, we will design a network to transform $\Gamma = 0.5 \angle -40°$ to $\Gamma' = 0$ (50 ohms). After plotting Γ on the combination chart, we find a path to the desired Γ′. As shown in Figure 9B, one possibility would be to travel upward along a circle of constant Re (y), then downward along a circle of constant Re (z). From our previously stated rules, these segments correspond to a shunt inductor and series capacitor. Along the circular segment representing the inductor, Im (y) changes from 0.32 to −0.48, for $\Delta$ B = 0.80.

For the capacitor segment, Im (z) changes from 1.30 to 0.00, for $\Delta$X = 1.30. Element values are then calculated as follows:

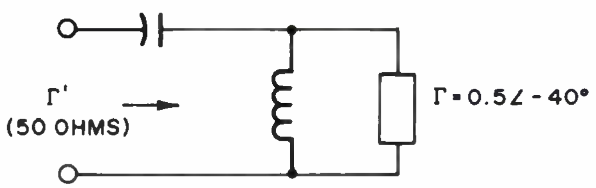



**Figure 9C.**

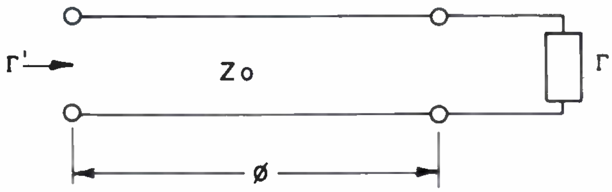



**Figure 10A. Series Transmission Line**

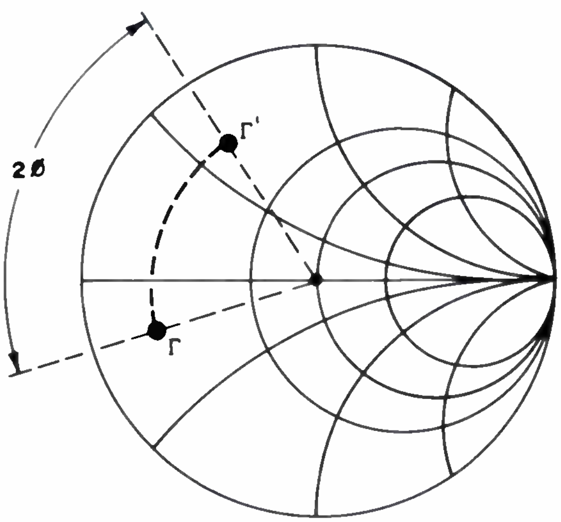



**Figure 10B.**

$$X_L = 50/0.80 = 62.5 \text{ ohms}$$

$$L = \frac{62.5}{2\pi \times 500 \times 10^6} = 19.89 \text{ nH}$$

$$X_c = 50 \times 1.30 = 65 \text{ ohms}$$

$$C = \frac{1}{65 \times 2\pi \times 500 \times 10^6} = 4.897 \text{ pFd}$$

The complete circuit is shown in Figure 9C.

## Distributed Elements

At upper RF and microwave frequencies, distributed elements are frequently used to synthesize circuit reactances. These elements are commonly realized as shunt or series transmission lines of arbitrary characteristic impedance and length. They may also be manipulated using the Smith chart.

## Series Transmission Lines:

Consider an arbitrary Γ. What is Γ′ if we cascade a transmission line of characteristic impedance $Z_o$ and electrical length Φ (see Figure 10A)? |Γ| will remain constant. A reflected signal, however, must now travel an additional electrical distance of 2Φ. The phase of Γ will therefore lag by an additional 2Φ, represented as a clockwise rotation on the chart (Figure 10B).

If the series transmission line impedance were an arbitrary $Z_A$, the following procedure would apply:

1. Plot Γ and note corresponding z
2. Re-normalize z to the characteristic impedance of the series transmission line.

$$z_1 = z \frac{Z_o}{Z_A}$$

3. Plot z, and rotate by 2Φ to $z_1'$.
4. Re-normalize $z_1'$ back to the characteristic impedance of $Z_o$.

$$z' = z_1' \frac{Z_A}{Z_o}$$

5. Impedance z′ locates the resultant Γ′

## Quarter-Wave Transformer:

An interesting special case of the arbitrary series transmission line occurs when Φ = 90° and $Z_A = \sqrt{Z_o R}$. (R is the impedance which corresponds to our initial Γ). Under these conditions, it may be shown that this one element will match R to the characteristic impedance of the system ($Z' = Z_o$) Analysis follows:

1. Normalize R to characteristic impedance $Z_A$.

$$R_1 = R/\sqrt{Z_o R}$$

2. Rotate $R_1$ by 2Φ. Recall that a Smith Chart rotation of 180° produces an impedance inversion.

$$R_1' = \frac{1}{R_1} = \sqrt{Z_o R/R}$$

3. Un-normalize $R_1'$

$$Z' = R' = (\sqrt{Z_o R})^2/R = Z_o$$

## Shunt Transmission Lines

A transmission line used as a shunt-connected circuit element is most easily handled by a two-step analysis. First, determine the normalized admittance (y) of the line at the point where it connects into the circuit. This is accomplished in the same manner as series line analysis, by plotting the initial Γ (which is usually a short or open circuit) and rotation by the correct 2Φ. Second, use the resultant B and/or G to modify the total circuit admittance at the point of connection, in the same manner as if the $\Delta B$ and/or $\Delta G$ were contributed by shunt-connected lumped elements.

## Designing With Transistors

The remainder of this article is devoted to the design of transistor amplifiers. First will be a brief qualitative discussion of the characteristics of RF transistors. We will then discuss and analyze several approaches to the design of amplifiers to meet performance goals. Although all formulas and equations necessary for design will be reproduced in this text, a detailed S-parameter review will not be included. For those who wish to read further on this subject, several excellent articles are listed in the bibliography. Finally, we will design in detail two amplifiers; a 440 MHz single-stage and a 1550 MHz two-stage. Working step-by-step through analytic and graphical design, computer-aided optimization, and prototype testing, I hope to illustrate several useful techniques for achieving desired amplifier performance.

## Transistor Characteristics

Small-signal performance of RF and microwave transistors is generally characterized by frequency-dependent two-port scattering parameters. Being defined under conditions of resistive input and output termination, the S parameters are much better suited to high frequency measurement on potentially unstable devices than parameters requiring short or open-circuit terminations.

The four S-parameters are $S_{11}$, $S_{22}$, $S_{21}$ and $S_{22}$. They are generally complex, having both magnitude and phase. $S_{11}$ and $S_{22}$ are the input and output voltage reflection coefficients (Γ). $S_{21}$ and $S_{12}$ are the forward and reverse voltage transfer coefficients, respectively.

Certain general characteristics are shared by most packaged small-signal RF and microwave transistors:

1. Input impedances which are capacitive at lower frequencies, gradually become inductive with increasing frequency. The crossover point is usually slightly below the center of the useful frequency range of the transistor.
2. Output impedances behave similar to the input impedances though generally not as rapidly changing with frequency (less dispersion).
3. Voltage gain which decreases with frequency at the rate of approximately 6 dB per octave throughout most of the useful frequency range.
4. Potential instability, particularly at low frequencies where gain is high.
5. Noise figure which increases with frequency. At a specific frequency, there will be a bias condition (usually low current) which potentially offers optimum noise performance.

Given the inherent characteristics and limitation of transistors, the designer's task is to develop methods for predicting and controlling the terminal behavior of completed amplifiers. Let us now discuss some commonly used approaches.

## Negative Feedback

In amplifiers, negative feedback is a method by which a portion of the output signal is subtracted from the signal present at the input. Intuitively we can see the effect this would have in minimizing gain variations. If the amplifier gain decreases, then so does that portion of output signal being subtracted from the input. The effective input signal level is therefore higher, so the overall gain does not decrease as much as it would have in a non-feedback configuration. For the following analysis of generalized negative feedback, refer to the diagram of Figure 11. "A" represents the gain of the uncompensated active device in the amplifier. $\beta$ controls the amount of $V_2$ (output signal) that is subtracted from $V_1$, (input signal). $V'$ is the actual signal presented to the input of "A". From inspection we can write two equations in the independent variables A and $\beta$, then solve by substitution for voltage gain, $V_2/V_1$.

1. $V_2 = AV'$
2. $V' = V_1 - \beta V_2$
3. $V_2 = A(V_1 - \beta V_2)$

$$AV_1 = V_2 + A\beta V_2$$
$$AV_1 = V_2(1 + A\beta)$$

4. $$\frac{V_2}{V_1} = \frac{A}{1 + A\beta}$$

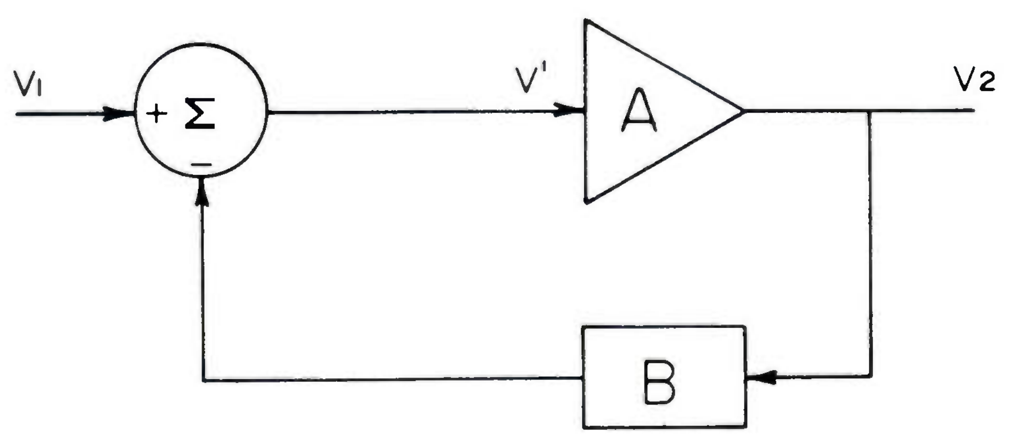



**Figure II. Feedback Amplifier**

The preceding is the exact expression for gain. However, if either A or $\beta$ is very large, then $A\beta >> 1$ and

$$\frac{V_2}{V_1} \approx \frac{A}{A\beta} = \frac{1}{\beta}$$

Examination of the exact expression for gain will illustrate the flattening effect of the feedback:

$$\frac{V_2}{V_1} = \frac{A}{1 + A\beta}$$

Let $A = 100$ (40 dB) and $\beta = 0.1$

Suppose "A" is the voltage gain of a single transistor. Over one octave bandwidth, "A" may be expected to decrease by 6 dB ($A' = 50$) amplifier gain becomes:

$$\frac{V_2}{V_1} = \frac{50}{1 + 50(0.1)} = \frac{50}{6} \approx 8.33\,(18.42\,dB)$$

Application of negative feedback has reduced a 6 dB active device gain variation to a 3/4 dB overall amplifier variation.

The previous example also illustrates one of the disadvantages of negative feedback. In order to achieve a reasonably flat amplifier with a voltage gain of about 10, we started with an active device gain of 100. We had to "throw away" most of our available gain. In practice, most designers reduce this problem by designing the feedback network such that $\beta$ is frequency dependent. By having $\beta$ decrease as frequency increases, more flattening can be realized with less reduction in overall gain.

As we approach upper RF and microwave frequencies, several difficulties arise in applying feedback techniques.

1. As devices become nonideal and phase characteristics must be considered, multiple signal paths make analysis extremely difficult.
2. Topology for coupling a signal from output to input becomes increasingly critical and difficult to implement at higher frequencies.
3. As "A" and "$\beta$" begin to acquire significant phase lag, the feedback signal may be added to, rather than subtracted from, the input. This net positive feedback may be a cause of potential instability or oscillation.
4. At higher frequencies, we become increasingly concerned with keeping all of our available device gain.

## Reactive Mismatch

At L Band and above, reactive mismatch is the most commonly used method of controlling amplifier gain. The technique here is to imbed our active devices between passive networks whose transfer characteristics over a frequency range of interest exactly cancel the frequency dependence of device gain, resulting in a flat amplifier. There are endless combinations which will accomplish this (see Figures 12A, B, C). One major advantage of the mismatch technique is that it results in a circuit consisting of cascaded elements, which lends itself easily to graphical analysis and design.

For graphical design, we make the *unilateral assumption* for active devices. If $S_{12}$ may be assumed negligible, then the device is said to be unilateral. When the unilateral device is terminated in a source and load having arbitrary reflection coefficients $\Gamma_S$ and $\Gamma_L$, the total gain is:

$$G_T = \frac{(1 - |\Gamma_s|^2)}{|1 - S_{11}\Gamma_s|^2} \times |S_{21}|^2 \times \frac{(1 - |\Gamma_L|^2)}{|1 - S_{22}\Gamma_L|^2}$$

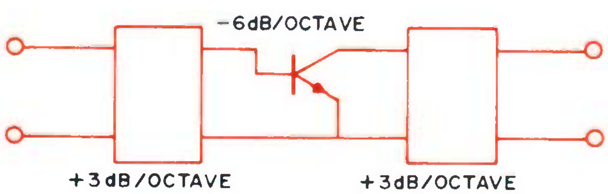


Figure 12A.

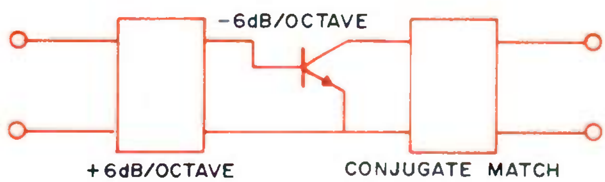


Figure 12B.

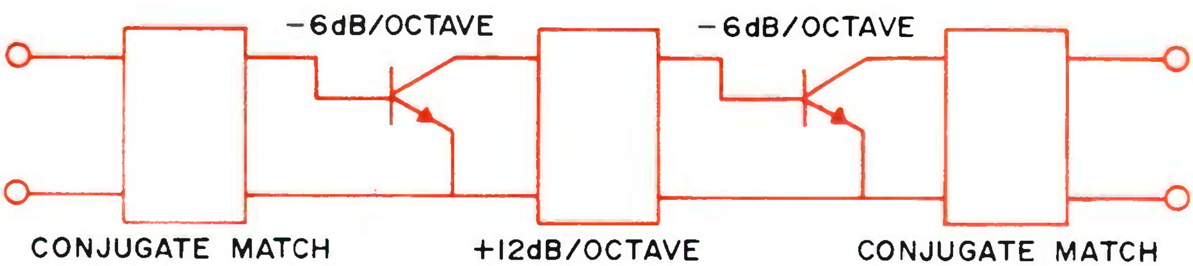


Figure 12C.

$$= G_s \times |S_{21}|^2 \times G_L$$

$G_s$ and $G_L$ are the "matching" gains (or losses) realized by terminating the device input and output in impedances other than $Z_0$. Notice that, for $\Gamma_s$ and $\Gamma_L = 0$, the expression for $G_T$ reduces to $|S_{21}|^2$. The maximum available gain occurs when $\Gamma_s$ and $\Gamma_L$ conjugately match $S_{11}$ and $S_{22}$. We then salvage all of the gain which was previously lost due to input and output VSWR.

In summary, our design procedure attempts to control the frequency characteristics of $G_s$ and $G_L$ such that they compensate for the variations in $|S_{21}|^2$. The input and output networks must be such that they transform our source and load to a $\Gamma_s$ and $\Gamma_L$ which will provide the desired matching gains (Figure 13A, B). □

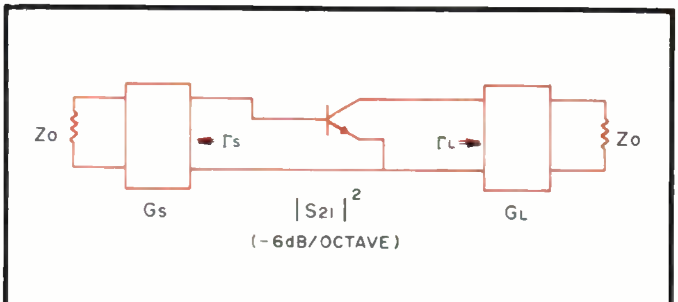



**Figure 13A.**

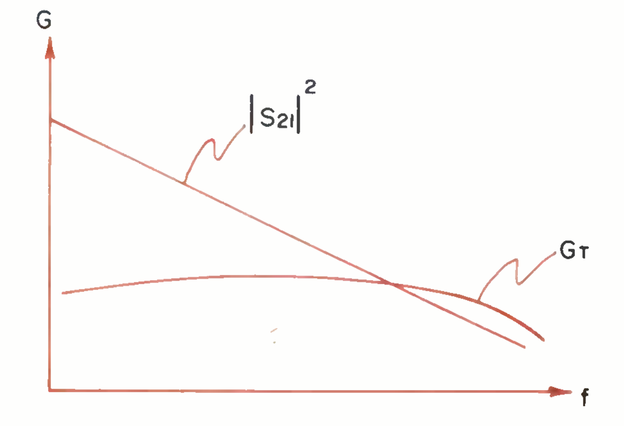



**Figure 13B.**

## References


1. White, Joseph F., *The Smith Chart: An Endangered Species?*, Microwave Journal, Nov. 1979, Pgs. 49-54.
2. Anderson, Richard W., *S-Parameter Techniques For Faster, More Accurate Network Design.*, Hewlett-Packard Application Note 95-1.
3. Frohner, William H., *Quick Amplifier Design With Scattering Parameters*, Electronics, Oct. 16, 1967.
4. Weinert, Fritz, *Scattering Parameter Speed Design of High-Frequency Transistor Circuits*, Electronics, Sept. 5, 1966.

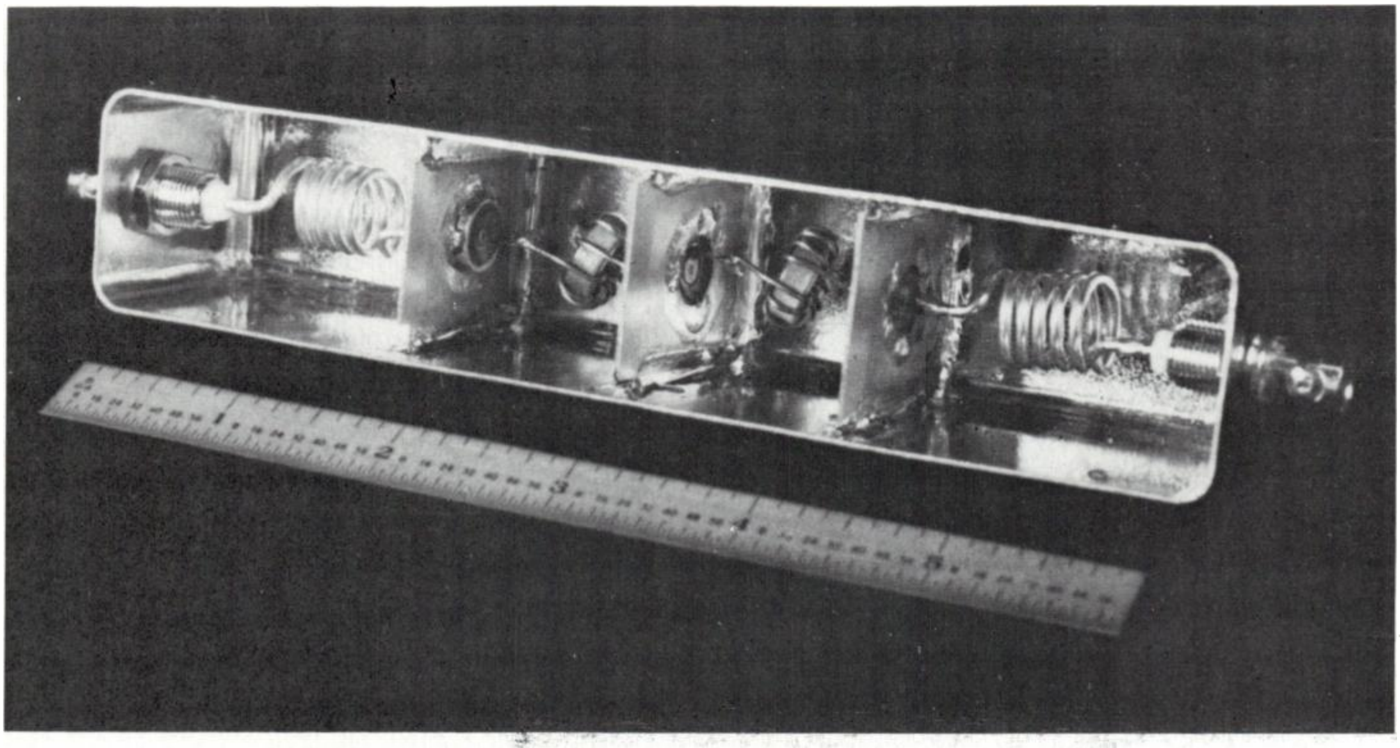


**Figure 2. Photograph of lowpass filter constructed from design #31.**

# lowpass filters

## Table of Precalculated Chebyshev Lowpass Filters with Inductive Input and Output, Part II

*Edward E. Wetherhold*
*Honeywell, Inc.*
*Annapolis, MD*

### Construction and Testing Of Filter Design #31

To further demonstrate the validity of the precalculated designs in Table 1, (see Part 1, July-August 1981) and to illustrate a suitable construction technique, design #31 was built and tested. Figure 2 shows a photograph of the completed filter. Normally, all inductors of a typical lowpass filter would be wound on iron powder toroidal cores because of the advantage of self-shielding provided by this form of construction. However, because of the relatively small inductance of L1 and L7 compared to L3 and L5 (0.205 $\mu$H compared to 0.758 $\mu$H), it was convenient to use an air-core construction for L1 and L7, and higher Q was possible with this type construction than with the toroidal core. For example, at 16 MHz, the Q of the air-core coils was 175 whereas an inductor wound on a Micrometals T68-10 core had a Q of only 140. Because inductors L3 and L5 had a much

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[13]D. Kochen, "Practical VHF and UHF Coil-winding Data," *ham radio,* April 1971.

higher value of inductance, it was more convenient to use iron-powder toroidal cores than air-core coils. Thirteen turns of #20 magnet wire on a Micrometals T50-6 core gave the required inductance for L3 and L5 with a Q of 260 at 16 MHz. For L1 and L7, six turns of #14 bus were wound on a 3/8-inch diameter form with a coil length of 11/16 inches.

The design data of Reference 13 was used as a guide in obtaining practical coil dimensions. The 3/8-inch diameter was selected so the winding would be separated from the sides of the case by more than the coil radius. This is necessary so that the coil Q and inductance will not be significantly reduced by the close proximity of the case.

Before installing the inductors they were adjusted to the design value while measuring them with a HP 4342A Q meter. The Erie Button-* * Mica feed-through capacitors were checked for correct capacity, and then installed on partitions cut from double-sided p.c. board. The partitions and capacitors were then soldered in place within the case, and coils L1 and L7 were installed. The circuits of L1, C2 and L7, C6 were checked to see if they resonated at the proper frequency** of 23.7 MHz. The measured resonant frequency of these two circuits was about 0.7 MHz too high, and this was probably due to a slight reduction in inductance of L1 and L7 after they were installed within the case. Because the measured resonance of these two circuits was close enough to the design frequency, no attempt was made to improve their tuning. The remaining inductors were installed, and the case was sealed with its cover. The case and cover were drawn steel with a hot-tin dip to facilitate the soldering of the partitions. The case and cover dimensions were 1.281 x

*ERIE registered trademark.

**$f_r = 1/(2\pi\sqrt{L1 \bullet C2})$

## Appendix A — Filter Equations

(1) $\rho$ = (R.C.)/100, where R.C. is the reflection coefficient in % and $\rho$ is the absolute value (magnitude) of the reflection coefficient.

(2a) $\rho = (1 - 0.1^{x})^{.5}$, for x = A/10 where A is the filter attenuation in dB at any frequency and $\rho$ is the magnitude of the reflection coefficient corresponding to A. For example, if A = 0.01 dB, $\rho$ = 4.7958%.

(2b) $A_{(dB)} = -10 \bullet \log(1 - \rho^2)$

(3a) $$\mathrm{VSWR} = \frac{1+\rho}{1-\rho}$$

(3b) $$\rho = \frac{\mathrm{VSWR} - 1}{\mathrm{VSWR} + 1}$$

(4) $\varepsilon = (10^{x} - 1)^{.5}$, for $x = A_p/10$ where $\varepsilon$ is the ripple factor, a parameter (less than one) related to the maximum ripple amplitude ($A_p$).

(5) $$\varepsilon = \frac{\rho}{\sqrt{(1-\rho^2)}}$$ where $\rho$ is the reflection coefficient at $F_{Ap}$

(6) $A_{s(\Omega)} = 10 \bullet \log[1 + (\varepsilon \bullet C_\Omega)^2]$ where $A_s$ is the filter stopband attenuation in dB and C is the value of the Chebyshev polynomial, both a function of $\Omega$.

(7) $C_\Omega = 64\Omega^7 - 112\Omega^5 + 56\Omega^3 - 7\Omega$, where C is the value of the Chebyshev polynomial for a 7-element filter as a function of $\Omega$.

(8) $\Omega$ = normalized frequency = $F_{As}/F_{Ap}$ where $F_{As}$ and $F_{Ap}$ are the stopband and cutoff frequencies depicted in Figure 1(B). For example, if $F_{Ap}$ = 0.82 MHz and $F_{As}$ = 1.64 MHz, then $\Omega$ = 2.0 and C = 5042. For a filter design with R.C. = 0.614%, $\varepsilon$ = 0.00614012 and $A_s$ = 29.8 dB at $F_{As}$ = 1.64 MHz.

(9) $$\Gamma_{(complex)} = \frac{[R^2 + X^2 - R_o^2] + j[2 \bullet X \bullet R_o]}{(R + R_o)^2 + X^2}$$ (For inductive input filter)

where: $\Gamma$ = complex reflection coefficient of the filter
R = resistive (real) component of filter input impedance (ohms)
X = reactive (imaginary) component of filter input impedance (ohms)
$R_o$ = filter termination resistance (real) in ohms

(10a) $$\Gamma_{(real)} = \Gamma_r = \frac{[R^2 + X^2 - R_o^2]}{(R + R_o)^2 + X^2}$$

(10b) $$\Gamma_{(imaginary)} = \Gamma_i = \frac{[2 \bullet X \bullet R_o]}{(R + R_o)^2 + X^2}$$

(10c) $\rho = \sqrt{\Gamma_r^2 + \Gamma_i^2}$

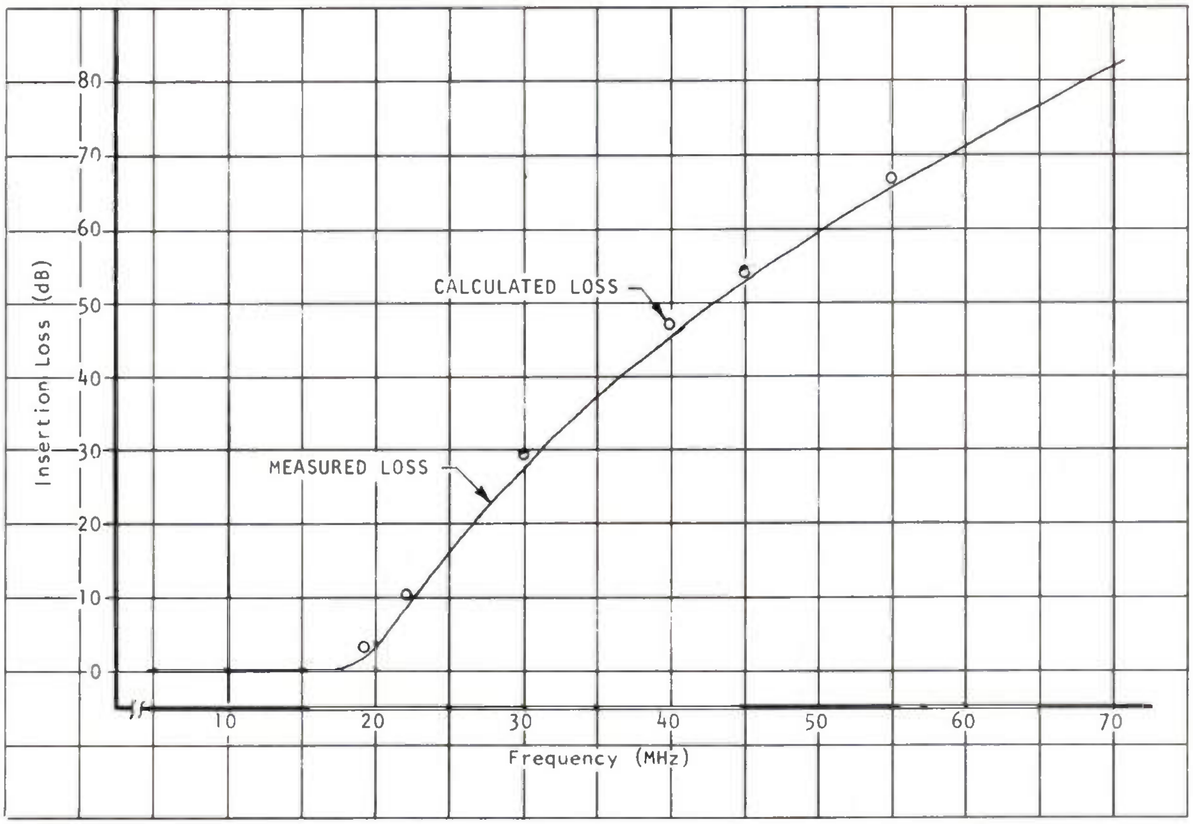



**Figure 3. Measured and Calculated Insertion Loss of Filter Design #31.**

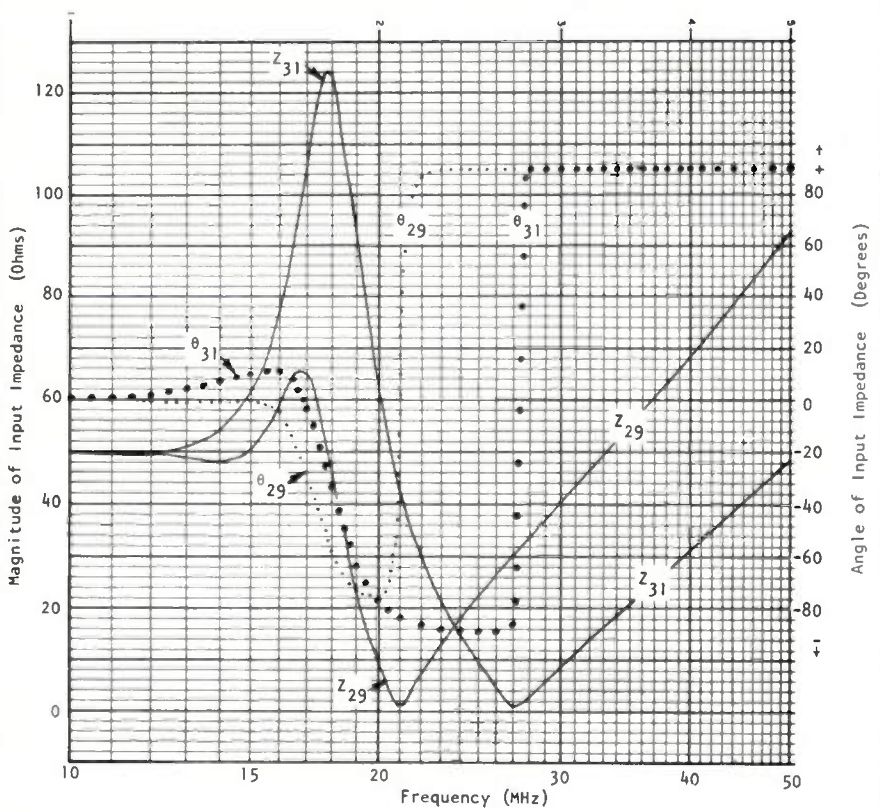



**Figure 4. Input Impedance of Designs 29 and 31 with 50-ohm Load.**

6.609 x 1.562 inches and 1.224 x 6.537 inches, respectively.

The completed filter was tested for insertion loss, and its measured and calculated responses are shown in Figure 3. The good agreement between the measured and calculated values indicate that the design was properly constructed. The measured insertion loss continues to rise above 60 MHz, indicating that the filter stopband performance is satisfactory. The superior stopband performance is due to the low impedance-to-ground of the Button-mica capacitors and to the good isolation between filter sections provided by the partitions.

Figure 4 shows the calculated magnitude ($Z_{31}$) and angle ($\theta_{31}$) of the input impedance of the constructed filter (#31), which has a relatively low reflection coefficient 6.57 • $10^{-3}$ percent. The input impedance of design #29, with a higher reflection coefficient of 2.21 percent, was also plotted for comparison. The input impedance of both designs varies considerably with frequency until the inductive reactance of L1 swamps out all shunt capacity of the filter. After this, the input impedance becomes purely inductive, and the impedance steadily increases with increasing frequency. These impedance curves indicate that a solid-state amplifier must be able to tolerate a varying load impedance above the $A_p$-cutoff frequency if the amplifier is to remain stable.

These curves also indicate the advisability of selecting the highest acceptable value of reflection coefficient and VSWR, and then making sure that the maximum operating frequency of the amplifier is placed in the filter passband so the maximum desired VSWR presented by the filter to the amplifier is not exceeded. In the two design examples of input impedance plotted in Figure 4, design #29 is more suitable for a 14 MHz lowpass filter application than design #31 because it provides more attenuation at the second harmonic frequency and it has a lower VSWR at 14 MHz.

The Butterworth or near-Butterworth designs must be used with discretion in RF filtering applications where it is important to minimize VSWR. Although the passband attenuation of these filters is relatively flat, the VSWR can become excessive sooner than anticipated. Consequently, the user should be aware of this problem and be prepared to compromise the attenuation response to assure that the filter VSWR is acceptable at the operating frequency.

## Summary

The occasional need for an inductive input lowpass filter to both attenuate harmonics and to maintain the stability of solid-state amplifiers was explained. A table of eighty-seven 50 ohm precalculated filter designs was given in which only standard-value capacitors were required to simplify construction. Procedures were given to confirm the validity of the designs, and to easily obtain designs for equal termination impedances other than 50 ohms. A filter design was constructed and its measured insertion loss was compared with its calculated insertion loss to confirm the validity of the design. The magnitude and phase of the input impedance of two terminated designs were calculated and plotted to illustrate the wide range of impedance variation the solid-state amplifier must tolerate without becoming unstable. The excessive input impedance variation of the Butterworth and near-Butterworth filter designs compared to the Chebyshev was discussed, and recommendations were given for the selection of suitable filters.

The data in this article and in references 3, 4, 5 and 6 provide the RF engineer with complete information on precalculated designs of the equally-terminated 7 element Chebyshev lowpass filter. Preference should be given to these designs whenever this type of filter is needed because only standard-value capacitors are required.


## Acknowledgements

The author gratefully acknowledges the assistance of:

Roosevelt Townsend of Honeywell for help in developing the computer programs used to calculate the filter table,

Joseph Green of Honeywell for deriving Equation (10) in Appendix A,

Charles Miller of Honeywell for plotting the filter insertion loss,

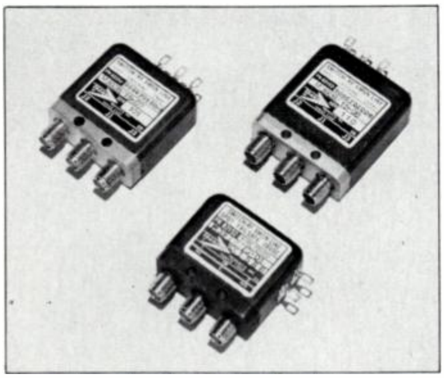
Philip Geffe of Scientific-Atlanta for providing the basic programs used to calculate the attenuation and input impedance of the filter designs,

and Joseph Gutowski of EWC Inc. and Rex Cox of Honeywell for their review of the manuscript.

## Transco Switch Special Mounting

In response to domestic and international market requirements, Transco announces a special outboard mounting for their single pole, dual throw coaxial switches 919C77000 and 919C77100.

According to Transco, these switches have established a worldwide reputation for reliability and long life. The new mounting permits their use in many additional applications including new hi rel.



For complete information on these models and Transco's extensive line of Qualified Products List (QPL) coaxial and waveguide switches, please call or write for your copy of the company's 92-page RF Switch Catalog.

Contact Transco Products, Inc., 4241 Glencoe Avenue, Marina Del Rey, CA 90291. INFO/CARD #140.

## 225-400 MHz Solid State Amplifier

For EW System developmental and on-line use, 225-400 MHz at power output levels of 35, 100, 200 and 400 watts CW class "AB" linear. Gain variation ± 1.5 dB, drive levels from 3 to 10 mW. These units feature a three range power meter that includes simultaneous forward, reflected and VSWR. May be used for precision broadband antenna and transmission

line evaluations with optional equipment (SL-1 leveler) to achieve ± 0.3 dB flatness over the entire frequency range. Other applications include RFI/EMI testing, NMR Spectroscopy, Gas Plasma and Laser Modulator Driver, OEM Modular Systems available.

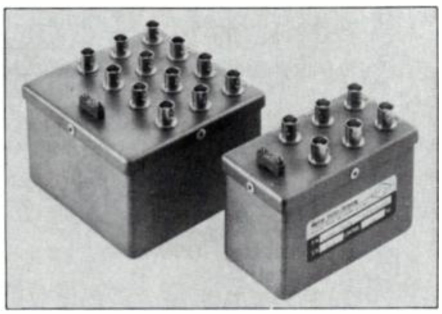
Contact RF Power Labs, Inc., 21820-87th S.E. Maltby Industrial Village, Woodinville, WA 98072. INFO/CARD #139.

## Multi-Pole, Multi-Throw Coax Switches

A family of multi-pole, multi-throw coax switches with high isolation, low insertion loss and excellent VSWR have been announced by MarLee Switch Co.

Covering a DC to 100 MHz frequency range, the coax switches have an operating speed of less than one millisecond, including bounce. They are available in one to ten poles and one to ten throws per pole, in a single housing.

Open contact isolation is as high as − 110 dB, insertion loss is as low as 0.2 dB. VSWR is typically 1.5:1. Characteristic impedence is 50 ohms. Other impedances or terminated versions are available to match system requirements.

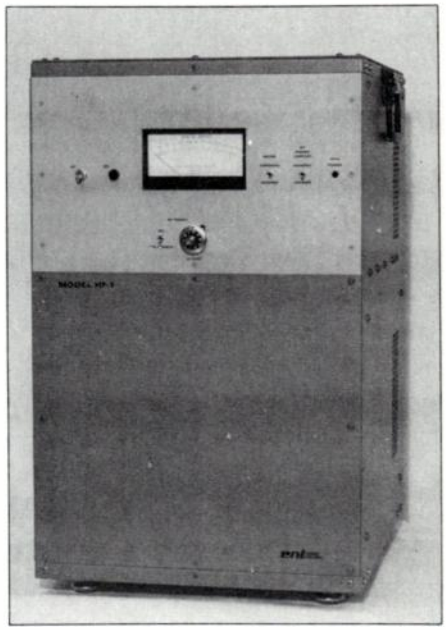


The MarLee switches are available with IEEE-488 or RS232C bus or telephone interfaces. They also are available for use with Triax or Twinax and for frequencies to 1.2 GHz.

Contact MarLee Switch Co., 933-D N. Central Ave., Upland, CA 91786. INFO/CARD #138.

## 1 KW Power Generator

ENI Power Systems, Inc. introduces the first 1 KW all solid state power generator — the HF-1. In the past the generation of AC power for plasma



systems, including reactive ion etching, vacuum deposition and sputtering has been accomplished by vacuum tubes. ENI Power Systems has pioneered the development of rugged, efficient, reliable, and versatile transistorized high power generators. The HF-1 is designed to operate at 13.56 MHz into any tuned load. However, this extremely rugged generator is warranteed by ENI Power Systems for a full year against failure to operate continuously into any load condition. It will operate without shut down, oscillation or failure into any mis-tuned load, including an open or short circuit.

The generator may be remotely programmed by any externally derived signal from either a computer or control device to produce any power profile.

Either manual or automatic power leveling on forward power is provided. Both forward and reflected power are monitored by a front panel wattmeter.

The HF-1 may be used anywhere, operating from any 208/240 VAC 50/60 Hz single phase line with a 20 ampere capacity. The air cooled unit has a well-regulated and filtered power supply. Extensive use of RF suppression and filtering insure that it more than meets FCC requirements for radio frequency for ISM equipment.

Contact ENI Power Systems, Inc., 3000 Winton Road South, Rochester, NY 14623. INFO/CARD #137.

## 1 To 1000 MHz Sweep Generator

Wavetek introduces the first in a series of versatile sweep generators capable of sweeping a 1000 MHz band without switching oscillators. The model 1080 covers a frequency range from 1 MHz to 1000 MHz and has three modes of operation: Full Sweep, ΔF and C.W. In Full Sweep mode the Start-Stop frequencies are fixed at 1 MHz and 1000 MHz respectively and a variable marker, controlled by a 10 turn potentiometer is displayed on a three plus one LED digital readout.

This control and frequency indicator becomes the center frequency of the sweep by simply pushing the function control button which switches to the ΔF mode. The sweep may then be expanded by the Sweep Width controls from 1000 MHz to 200 kHz for maximum resolution.

The 1080 has a variable power output from + 13 to − 70 dBm and is displayed on a three digit LED readout to a resolution of 0.1 dB. A unique "auto-zero" circuit automatically corrects for long-term drift by periodically resetting the zero frequency of the oscillator. A wide variety of sweep rates are selectable along with adjustable marker amplitude and width.

Harmonic crystal markers are included at 1, 10, and 1000 MHz and may be used to calibrate the 1080 frequency output to ± 1 MHz. (Accuracy without calibration is ± 10 MHz to 1000 MHz.)

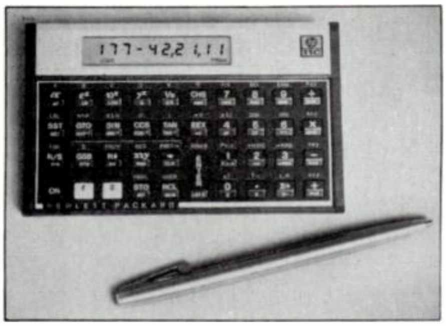
Contact Wavetek Indiana, Inc., 5808 Churchman, P.O. Box 190, Beech Grove, IN 46107. INFO/CARD #136.

## 'Slim Line' Scientific Calculator from HP

A powerful new programmable scientific calculator is the first of its kind in a "slim line" design that fits in a shirt pocket from Hewlett-Packard.

The new HP-11C features a horizontal keyboard design with liquid-crystal display, continuous memory. Many tools for easy programming, and a full scientific function set with advanced functions such as a random-number generator, are also featured.

Suggested retail price for the HP-11C



is $135. HP also has introduced a slim line programmable financial calculator, the HP-12C, at a suggested price of $150.

Programming the HP-11C to solve complex and/or repetitive scientific and engineering problems is easier thanks to innovative programming tools. Fifteen program labels make it easy to call a program or subroutine with the touch of one key. Other programming tools include indirect addressing, conditional tests, flags, four levels of nested subroutines, and a 200-line memory.

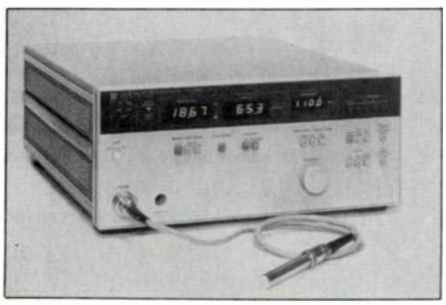
The HP-11C Slim Line Advanced Programmable Scientific Calculator with Continuous Memory, and the HP-12C Slim Line Advanced Financial Programmable Calculator with Continuous Memory and Special Functions, are both available now from dealers of HP calculators and from HP.

Contact Inquiries Manager, Hewlett-Packard Co., 1820 Embarcadero Road, Palo Alto, CA 94303. INFO/CARD #135.

## Vector Impedance Meter, 400 kHz To 110.0 MHz

This new HP model 4193A provides a constant current test signal to a circuit or device and then measures impedance magnitude from 10 milliohm to 120.0 kilohms and phase from − 180.0 to + 180 degrees. Frequency range is 400 kHz to 110.0 MHz.

Reliable in-circuit testing can be done using a furnished, grounded probe. Devices of various size and shapes can be tested using available test futures. No preadjustment of the probe is necessary. Typical stability of measured impedance magnitude and zero values is 2 milliohm per degree C, and 0.02 degree per degree C.



The HP 4193A features 3-1/2 digit impedance magnitude readings with accuracy as good as ± 3 percent of reading. Phase readings feature resolution 0.1 degrees and accuracy as good as ± 3.2 degrees.

Test frequency is variable from 400 kHz to 110.0 MHz with 4 digit resolution. Frequency can be tuned to individual "spot" frequencies or swept either linearly or logarithmically. Frequency accuracy is 0.01 percent of setting.

Applications which require high frequency resolution and accuracy such as ceramic resonator testing are handled by connecting a precise external signal source to the HP 4193A external oscillator input.

Included as standard features are HP-IB (IEEE-488) and analog outputs. These capabilities make the HP 4193A a good choice for automatic system applications in R & D, production, product assurance and incoming inspection.

Contact Inquiries Manager, Hewlett-Packard Company, 1820 Embarcadero Road, Palo Alto, CA 94303. INFO/CARD #134.

## Microprocessor-Based Sweep Generator Brochure

A new family of microprocessor-based Sweep Generators covering the

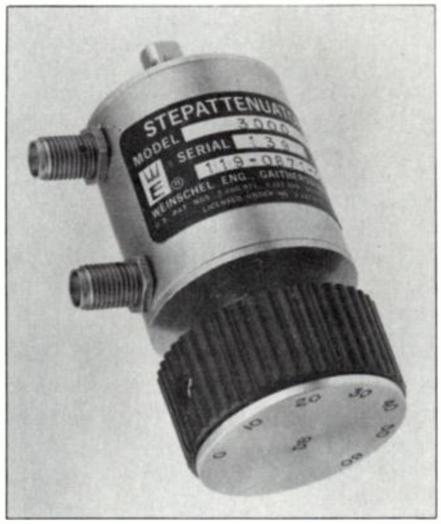

10 MHz to 40 GHz range is described in this 24-page 6600 Series brochure. Also included is the 560 Scalar Network Analyzer which when used with the 6600 Sweep Generator forms a network analyzer system with 40 dB directivity over the 10 MHz to 18 GHz continuous-sweep range. Graphics illustrate controls, test setups, applications, accuracy characteristics, specifications, and test results. For a free copy, write:

Walt Baxter, Wiltron Company, 805 East Middlefield Road, Mountain View, CA 94043. INFO/CARD #133.

## DC - 2.5 GHz Step Attenuators

The Weinschel Engineering 3000 Series Step Attenuators offer small size, broadband performance and a unique, dependable design.

Ideal for OEM applications in the DC-2.5 GHz Frequency Range, the Weinschel 3000 Series emphasizes mass production while retaining



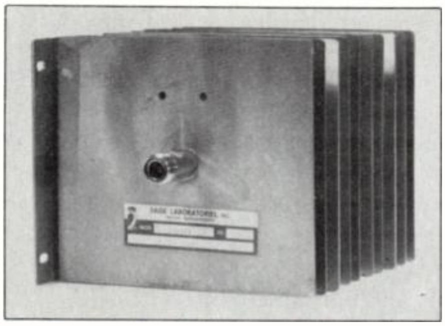
proven design-tested performance and reliability together with competitive pricing. A Weinschel patented detent mechanism tested to 1,000,000 operations at + 70°C and − 40°C adds to dependability. Use of patented resistive elements on ceramic substrates provide uniquely flat frequency response and consistent performance. Offered in single and concentric-shaft, dual drum configurations, the 3000 Series is available in 31 standard attenuation ranges and step increments. Incremental Attenuation Accuracy is: 0 − 60 dB: ± 0.3 dB or 1 percent, 60-110 dB: ± 2 percent, 110-140 dB: ± 3 percent. Radiated RF Leakage is greater than 85 dB below input. Frequency Sensitivity is typically 0.1 to 0.2 dB up to 2.0 GHz.

Contact Weinschel Engineering, One Weinschel Lane, Gaithersburg, MD 20877. INFO/CARD #131.

## Air Cooled Medium Power Dry Load

Sage Laboratories, Inc. introduces an 800 to 8000 MHz Dry Load optimized over the TACAN Band.

The model FT2974 is designed for low cost volume production. The unit offers a VSWR of less than 1.15:1 from 960 to 1215 MHz and a VSWR of less than 1.35:1 from 800 to 8000 MHz. The unit has a power rating of 5800 watts peak and 210 watts average. The size of the load is 6.95 X 5.18 X 6.00 inches.



Contact Ken Paradiso, Sage Laboratories, Inc., 3 Huron Drive, Natick, MA 01760. INFO/CARD #132.

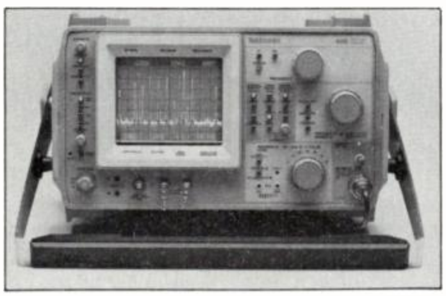
## SAW Literature

A 6-page brochure is available discussing new surface acoustic wave techniques. Featured are second-generation SAW filters exhibiting very low insertion loss, (4-6 dB) and SAW resonators for use as stable oscillator frequency control elements, all in the VHF-UHF range.


Contact RF Monolithics, Inc., 4441 Sigma Road, Dallas, TX 75234. Circle INFO/CARD #129.

## Tektronix 496 and 496P Spectrum Analyzers

The 496 and programmable GPIB compatible 496P Spectrum Analyzers



now available from Tektronix, Inc., are similar to the popular 492/492P Spectrum Analyzers in appearance, operation, and rugged environmental specifications. While the 492/492P Spectrum Analyzers cover the RF spectrum from 50 kHz to 220 GHz; the new 496/496P Spectrum Analyzers cover frequencies from 1 kHz to 1.8 GHz.

Important product features like frequency span, resolution, and stability have been optimized for the baseband, HF, VHF, and UHF high performance markets. The 492/492P and 496/496P, form a family of rugged high quality portable spectrum analyzers. Product emphasis has been placed on portability, ease of use, and high performance standards.

The 496/496 provide portability plus high performance spectrum analysis and measurements in the 1 kHz to 1.8 GHz range. Their high stability and 80 dB dynamic range meet user demands for proof-of-performance measurements, on site or on the bench. With the 496/496P, resolution bandwidth can be varied from 1 MHz to 30 Hz over the entire frequency range. Automatic phase lock stabilization reduces incidental FM to 10 Hz p-p; phase noise sidebands are at least −75 dBc at 30 times the resolution offset. Frequency drift is no greater than 330 Hz in 10 minutes; and there is 30 Hz resolution bandwidth with −127 dBm sensitivity.

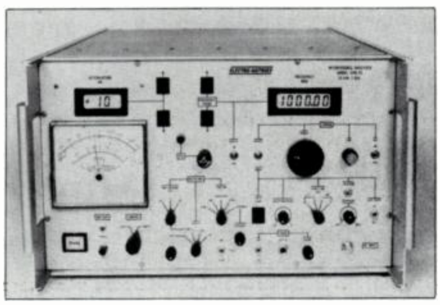
The 496P is the fully programmable and GPIB compatible version of the 496 Spectrum Analyzer. When used as a manual instrument, it incorporates all of the lab quality performance and ease-of-use features of the 496. Full programmability allows the user to operate the 496P under program control, change the front panel settings, read data from the CRT display, and send spectral waveforms from the internal digital source memory to other GPIB devices.

Contact Marketing Communications Department, D/S 76-260, P.O. Box 1700, Beaverton, OR 97075. INFO/CARD #126.

## EMI Testing Interference Analyzer

Electro-Metrics, a Penril Company announces the addition of model CPR-25 Interference Analyzer. The new CPR-25 is designed to satisfy the rapidly-growing need for instrumentation capable of performing compliance testing to the newly-proposed FCC Mandates.

In addition to the FCC testing application, the CPR-25 also includes all instrumentation characteristics for EMI measurements in conformance with CISPR, VDE, and ANSI requirements.

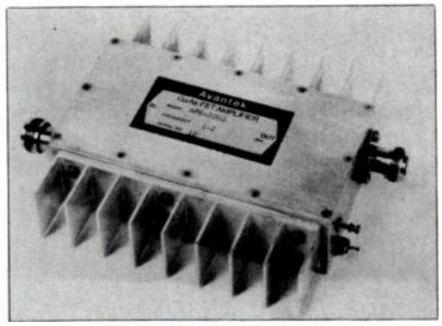


Included in the CPR-25 Interference Analyzer are a number of features which represent substantial advancements over instrumentation previously available. LCD digital readouts display attenuation setting and tuned frequency to an accuracy of ±0.1 percent over the entire 10 kHz to 1000 MHz range of the instrument. In addition to CISPR-mandated quasi-peak detection, the CPR-25 also incorporates circuits for detecting peak, average and true RMS levels. This information is either displayed on the 80 dB front panel meter or on a single or dual-pen X-Y recorder. The CPR-25 also includes video response over each octave-range band which can be presented on any conventional or storage-type oscilloscope, to create a tuned-front-end spectrum analyzer capability.

Contact Electro-Metrics, 100 Church St., Amsterdam, NY 12010. INFO/CARD #130.

## One-Watt, 1-2 GHz GaAs FET Amplifiers

Avantek, Santa Clara, Calif. has developed a series of GaAs FET amplifiers that offer +30 dBm minimum output power (1 dB gain compression) over the 1 to 2 GHz frequency range. Designated the APG-2000 Series,



these amplifiers feature a choice of 10, 20 or 30 dB gain with as low as ±0.5 dB full-band gain flatness, 4.5 to 5.0 dB noise figures and wide dynamic range performance, indicated by a +40 dBm third-order intercept point for intermodulation products. All APG-2000 Series amplifiers are packaged in compact aluminum cases with O-ring lid seals for protection from humidity and heat-dissipating fins to permit free-air operation without special cooling provisions. These amplifiers may be used as intermediate power (IPA) drivers for TWTs in electronic defense equipment or in the laboratory to increase the available output from sweep generators.

All versions in the series operate from a single +15 VDC supply, requiring only 875 to 975 mA (typ.) for operation.

Contact Avantek, Inc., 3175 Bowers Avenue, Santa Clara, CA 95051. INFO/CARD #127.

## Driver Switches

Model 207 driver switches in nanoseconds, has rates up to 10 MHz. Fully TTL-compatible, device provides extra current-spiking for quick recovery, plus inverting programmable output currents in both positive and negative directions. Model uses +5, −12 V supplies and is packaged in 3/8 x 3/8 flatpack for operation up to +125°C. Screening for military requirements available. Delivery: Stock to two weeks.

(Continued on page 45.)


Contact New England Microwave Corp., 26 Hampshire Drive, Hudson, NH 03051. INFO/CARD #128.

## Voice Analyzer and Controller

A revolutionary self-contained analyzer and voice controller by Covox Company extracts the most important speech cues in the manner of a real human listener. Tolerant of noise and distortion, it accurately identifies voicing existance, voice fundamental pitch, and voicing duration. Vowel cues are cross correlated with voicing so as to improve accuracy. Conventional radio and telephone channels suffice for this speaker independent system, suggesting a variety of remote control applications. Common kinds of noise and clicks are suppressed, even when considerably more intense than the message signal.


Priced under $400, 16 separate and distinct commands are recognized with unlimited expansion capability when used with a microcomputer. Suitable for battery power, anyone with a working larynx can use the system, including the handicapped. Anything that can be switched can be controlled by voice, such as pumps, motors, conveyors, lights, etc. Accurate extraction of fundamental pitch offers an additional data element suitable for proportional control. When operated with a host microprocessor, a high degree of security can be achieved for identification, accessing privileged files, etc. Related applications include continuous speech recognition, speech bandwidth compression, and "speech to touch" aids for the deaf and deaf/blind.

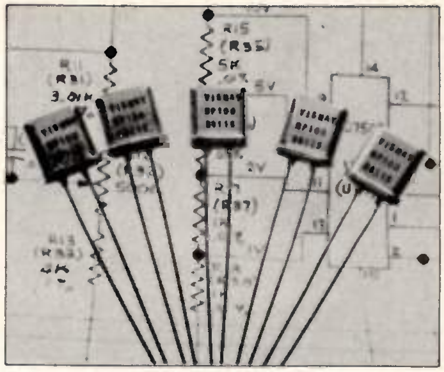
Contact Covox Company, P.O. Box 2342, Santa Maria, CA 93455. Circle INFO/CARD #125.

## Thermotropic Precision Resistor

Vishay Resistive Systems Group of Vishay Intertechnology, Inc. has announced a technical breakthrough in resistor design. . . a resistor with self-correcting resistance versus temperature performance. Named the Model HP100, the new Vishay thermotropic precision resistor is characterized by several unusual features, the most significant being the component's essentially zero TCR.

In practice, as temperature changes cause resistance to move away from the desired ohmic value, corrective factors within the HP100 reverse the direction of resistance change to restore the resistance to its initial ohmic value. In other words, changing temperatures, which normally increase error, actually provide thermal stability in the Vishay Model HP100.

In addition to the "zero" TCR, the HP100 is hermetically sealed against moisture, exhibits extremely small thermal EMF, offers excellent stability, and is essentially non-inductive in performance. HP100 specifications include:



- Resistance Tolerance Constant Within 0.003 percent (30ppm) regardless of temperature (0 to + 125°C)
- TCR: Less than 1 ppm/°C (0°C to + 125°C)
- Tolerance: to + 0.005 percent
- Shelf life stability: 0.0005 percent (5ppm) max delta R (1 year); 0.001 percent (10ppm) max delta R (3 years)
- Thermal EMF: 0.05 $\mu$ V/°C

Contact Vishay Resistive Systems Group, 63 Lincoln Highway, Malvern, PA 19355. INFO/CARD #124.

## TRW Introduces Low Noise Communication Transistor

A high-performance, small-signal, low-noise transistor for front-end receivers, the LT-4700, has been intro-

duced by TRW RF Semiconductors.

TRW said the device features forward insertion gains of 15 dB at 1 GHz and 21 dB at 500 MHz, and typical noise figures of 1.6 dB at 1 GHz and 1.1 dB at 500 MHz. According to TRW, these figures are superior to comparable data for existing devices.

The LT-4700 is housed in a hermetic, low-parasitic, high-frequency 100-square-mil package, has a maximum storage temperature rating of 200°C, and is suitable for both industrial and military use.

Contact Gene Brannock, product engineer, TRW RF Semiconductors, 14520 Aviation Blvd., Lawndale, CA 90260. INFO/CARD #123.

## Frequency Synthesizers

Programmed Test Sources has significantly expanded the capabilities of its VHF synthesizer line, which covers 0.1-200 MHz.

A new — (programmable) — very fast step attenuator is now available for the PTS 160 and PTS 200. With this option the output level can be

controlled over a 99 dB range with 1 dB resolution both manually and remotely through the parallel or the GPIB interface. This additional capability extends the range of potential applications into the signal generator domain. The superior switching speed of 5-20 $\mu$s for the synthesizer and 50 $\mu$s for the attenuator portion provides a very fast, agile, yet highly stable frequency source for automated test systems.

The GPIB (IEEE-488) interface may be specified in place of the parallel interface. It handles the frequency control functions at a rate of up to 20 $\mu$s per step, not limited by any other aspect of the interface circuitry except the controller itself. It also provides level control functions over a 10 dB range in 1 dB increments and optionally with the new attenuator a 90 dB range in 10 dB increments. The 488 interface is highly immune to invalid or redundant data, preventing unwanted states. GOTO LOCAL commands are implemented both as addressed and as a device-dependent (type) command.

Contact Programmed Test Sources, Inc., 9 Beaverbrook Road, P.O. Box 617, Littleton, MA 01460. INFO/CARD #122.

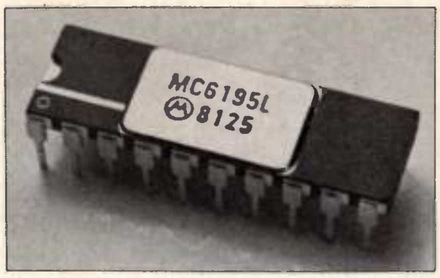
## Frequency Synthesizer TV Tuning

A phase-locked loop (PLL) frequency synthesizer device, the MC6195, is now available from Motorola. An NMOS silicon gate device, it is the nucleus of a digital tuning system for CATV converters and broadcast TV receivers and interfaces with a linear control chip and an Emitter Coupled Logic (ECL) prescaler to form the tuning system.

The phase-locked loop section of the MC6195 consists of a 100x 15-bit channel-conversion ROM which converts the channel number into the preset code for a 12-bit programmable divider. The prescaled local oscillator frequency is divided and compared to a divided-down reference frequency by a phase detector. The reference frequency is provided by an on-chip oscillator that uses either a 4 MHz or a standard color burst 3.58 MHz external crystal.

The device features are: remote-control capability for on/off control and for up/down channel scanning; Automatic Fine Tuning (AFT) circuitry which supplies the tuning voltage to the external linear amplifier; causing the change in frequency to the desired channel; and Binary Code Decimal (BCD) channel information for external LED display drivers.

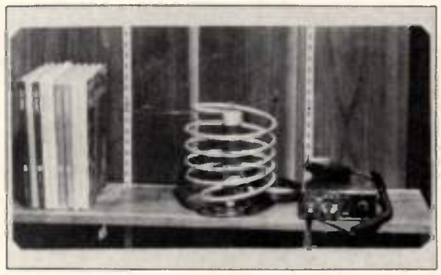
The MC6195 is the cost effective addition to Motorola's new M6190 family of phase-locked loop frequency synthesizers. The MC6190 through MC6195 are designed for use in a variety of TV and CATV applications. Another addition to the family, the MC6196, is in pre-production stages. It will be the cost effective version of the MC6192 system for broadcast TV frequencies. The price will be $2.50 in quantities of 100,000 and up, $4.00 in quantities of 1,000 and up, and they are available now.




Contact Motorola Inc., MOS Integrated Circuits Division, 3501 Ed Bluestein Blvd., Austin, TX 78721. Circle INFO/CARD #121.

## Compact, Indoor 10-Meter Antenna

The W1HGZ "Helican-10", just six inches high, is designed for those 10-meter enthusiasts who are restricted to indoor use. This convenient 10-meter antenna can handle up to 1,000 watts (PEP). With a range of 26 to 36 MHz, the "Helican-10" is easily tuned by adjusting the clip on the helix coil. The VSWR is about 1:1. This weather



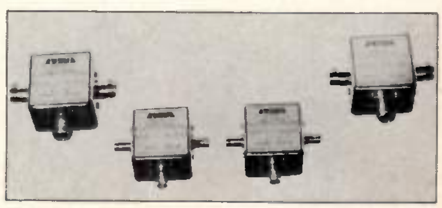
resistant antenna also has threaded sleeves to accept standard magnetic mounts. Price is $58.25 per unit; delivery time is one week.

Contact Emily Bostick, Microwave Filter Co., Inc., 6743 Kinne Street, East Syracuse, NY 13057. INFO/CARD #120.

## Return Loss Bridge

Eagle is announcing a family of new return loss bridges to augment their present line of time saving test and measurement accessories. These proven bridges assist in making easy and precise SWR measurements in two ranges: 0.04 MHz to 150 MHz and 5 MHz to 500 MHz.

These bridges are enclosed in a heavy duty nickel plated brass enclosure using all stainless steel hard-



ware. Each unit is proven with a series of over 300 mechanical and electrical tests in a fully automatic test set; this insures compliance with guaranteed specifications as well as insuring maximum field reliability. Power limits are an amazing 5 watts peak with better than – 40 dB of directivity over an operating temperature range of – 25°C to 50°C.

Eagle return loss bridges are useful in the following applications: VSWR measurements of filters, mixers, antennas or amplifiers; drive level measurements of mixers or amplifiers; power combining when source isolation is important. There are no nonlinear devices in these bridges thus virtually eliminating intermodulation products from the bridge.

Contact Eagle, 300 N. Main, Fallbrook, CA 92028. INFO/CARD #119.

### Handheld DMM's

The John Fluke Mfg. Co., Inc., has just introduced four new handheld DMM's. Dubbed the 8020B Series, these instruments are designed to replace the existing "A" series line which the company began manufacturing in 1977 with the model 8020A.

In keeping with the company's tradition of constantly making updates and revisions to improve the performance and usefulness of its products, the 8020B Series incorporates many changes suggested by owners of Fluke DMM's. While retaining all the popular functions and features of their predecessors, these models now feature:

- new, easier to read front-panel nomenclature
- a heavy-duty 600V dual fuse system on the current input to protect against accidental high energy inputs
- non-skid rubber feet and a locking tilt bail to keep the instrument firmly in place
- high-speed continuity beepers on three models; with a specified response time of 50 microseconds (typically 2000 times faster than other DMM's), these beepers will detect even the shortest mechanical contracts.
- two-year parts and labor warranties
- specifications guaranteed for a full two years — a significant benefit when one considers the costs associated with calibration and downtime

The four new models which make up the family are:

- The 8022B Handheld DMM. A six function instrument (ac and dc voltage and current, resistance and diode test) with 0.25 percent basic dc accuracy priced at $139 U.S.
- The 8021B Handheld DMM. Identical to the 8022B with the addition of the high-speed audible continuity beeper. Price $149 U.S.
- The 8020B Handheld DMM, the improved version of the most popular DMM ever made. The model retains all the features of the 8020A including conductance and adds the high-speed continuity indicator. The instrument has eight functions in all, 0.1 percent basic dc accuracy and lists for $189 U.S.
- The 8024B Handheld DMM is the updated version of Fluke's state-of-the-art 8024A. With 11 functions, the instrument will measure temperature with K-type thermocouples, retain

peak voltage and current readings and detect logic and continuity with audible and visual indicators, 0.1 percent basic dc accuracy and a price of $239 U.S.

Contact John Fluke Mfg. Co., Inc., P.O. Box 43210, Mountlake Terrace, WA 98043. INFO/CARD #118.

## Small Signal, Low Noise Microwave Transistor Data

A four-page data sheet describes TRW Semiconductors' LT4700 small signal, low noise microwave transistor that features a 6 GHz gain-bandwidth product ($F_t$) and a minimum noise figure ($NF_{min}$) of 1.6 dB a 1 GHz.

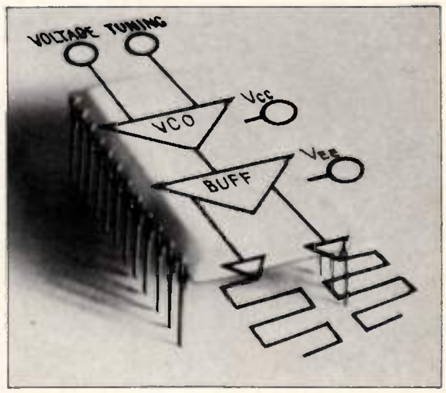
The data sheet provides the following charts: typical noise figure and associated power gain vs. frequency, noise measure vs. frequency, noise figure vs. collector current, gain-bandwidth product vs. collector current, insertion gain vs. collector current and output power vs. collector current. Also included in the data sheet is a tabulation of "s" parameters vs. frequencies from 100 MHz to 2 GHz. In addition, noise parameters are tabulated for 0.5, 1.0, 1.5 and 2.0 GHz.

Among the significant features of the LT4700 are low noise, high gain, wide dynamic range, high output capability, metal-ceramic hermetic package and MIL-S-19500 qualification. Its major applications are: satellite down conversion links, microwave relay communication links, ECM receivers, oscillators, mixers and multipliers.

Contact TRW Semiconductors, 14520 Aviation Blvd., Lawndale, CA 90260. INFO/CARD #116.

## 21-55 MHz VCO's

Frequency Sources Semiconductor Division has announced a new family of ECL Voltage Controlled Oscillators tailored for phase lock loop and clock applications in the 21-55 MHz range. Tailored for digital, computer, communications and instrumentation applications, the KJ1000 series gives high performance in disk memory data separation, system/slave clock generation, digital information processing, digital telecommunications, including any digital and analog circuits requiring reproducible and reliable performance. Low phase jitter is achieved through stable L/C tuning, giving digital designers superior performance in high speed circuits which can not tolerate noisy voltage controlled multivibrators.




The KJ1000 series features dual buffered complementary outputs for 10K, 10 KH and 100K levels and can be shifted for TTL. Floating tuning voltage inputs of 1-5 volts produce a 20 percent frequency range. One to 20 volts gives wide 50 percent bandwidth all in a 24 pin DIP package operating from a +5 or −5.2 volt single supply and requiring no external components.

Contact: Frequency Sources, Semiconductor Division, 16 Maple Road, Chelmsford, MA 01824. INFO/CARD #115.

## Video Measurement Capability Added to HP Oscilloscopes

Video sync and display capabilities are now optionally available in Hewlett-Packard's high frequency Models 1715A (200 MHz) and 1725A (275 MHz) oscillo-

scopes. Video and display systems containing high speed logic require these high frequency measurements for examination of fast transitions and precise timing relationships. This optional TV/Video Sync provides the ability to conveniently make separate high frequency and video measurements with one instrument.

This Option 005 consists of a module mounted on the instrument top covers and does not require internal modifications to the oscilloscope.

A 75 ohm input for impedance matching most video signal sources is provided. To insure a stable trace, a switchable TV clamp combines ac coupling and negative clamping. It compensates for vertical position shifting caused by varying levels of video information. Two trigger outputs, for main and delayed inputs, are provided to trigger the oscilloscope on composite video.

A vertical video output provides a method for connecting the signal to the instrument vertical input channel. With the proper trigger signals from the TV/Video Sync Option, and using the oscilloscope controls, specific portions of composite video can be selected for viewing. A single line scan control on the option allows precise examination of line segments.

Contact Inquiries Manager, Hewlett-Packard Company, 1507 Page Mill Rd., Palo Alto, CA 94304. INFO/CARD #113.

## Solid State Noise Sources

Micronetics, Inc., now offers a full range of solid state coaxial and waveguide noise sources. Included is the NSI Series for use with noise figure meters; frequency ranges are from 0.2 GHz to 26.5 GHz. Also, Waveguide NSI Series covers the broadband @ 18 GHz - 40 GHz. Available for use in systems is the RFN Series @ 10 kHz to 18 GHz, with operating temperatures at $-55°C$ to $+85°C$, temperature sensitivity: $-0.1$ DB/C° max.

Contact Micronetics, Inc., 36 Oak St., Norwood, NJ 07648. INFO/CARD #117.

# Classifieds

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**Prefer 2+ yrs. experience in MATV, CATV, or μW TV Communications.**

Pioneer of TV antennas and MATV receiving equipment has openings for RF Engineers to design broadband receiving equipment to 12 GHz.

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The successful candidate must be thoroughly familiar with up to date power supply technology and with all aspects of "RF" oscillators including oscillator matching to various load impedances. A BSEE from an accredited college is required.
The company has an excellent engineering staff providing an atmosphere conducive to professional advancement.
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Send resume in confidence to: Box RF-9-81-1, R.F. Design Magazine, 3900 S. Wadsworth Blvd., Denver, CO 80235



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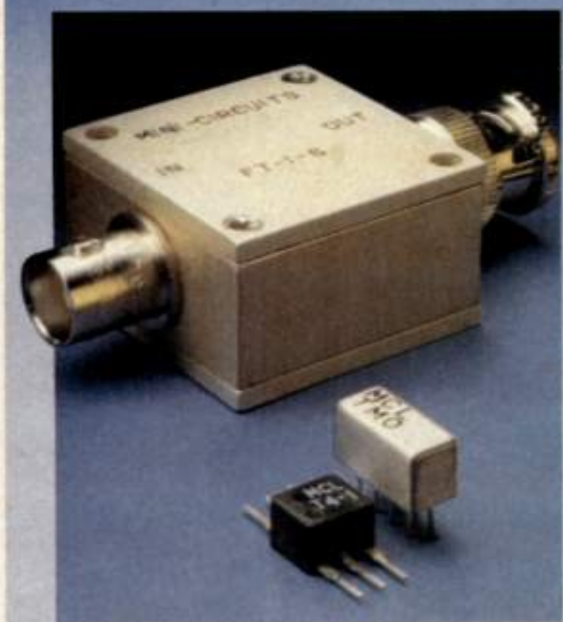


Select from the economical, microminiature T–series (plastic case) or TMO series (hermetically-sealed metal case) covering 10 KHz to 800 MHz. These models operate from 12.5 to 800 ohms with insertion loss typically less than 0.5 dB.

For large dynamic range applications, specify the T–H series which can handle up to 100 mA primary current without saturation or distortion.

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|---|---|---|---|---|---|---|---|---|---|
| Imped. Ratio | 1 | 1 | 1.5 | 2.5 | 4 | 9 | 9 | 16 | 16 |
| Freq. (MHz) | 15-400 | 8-300 | 1-300 | 01-100 | 02-200 | 15-200 | 2-90 | 3-120 | 7-85 |
| T Model (10-49) | $2.95 | $4.95 | $3.95 | $3.95 | $3.95 | $3.45 | $5.45 | $3.95 | $5.95 |
| TMO model (10-49) | $4.95 | | $6.75 | $6.45 | $6.45 | $6.45 | | $6.45 | |

**CENTER-TAPPED DC ISOLATED PRIMARY & SECONDARY**

| Model No. | T1-1T TMO1-1T | T2-1T TMO2-1T | T2.5-6T TMO2.5-6T | T3-1T TMO3-1T | T4-1 TMO4-1 | T4-1H | T5-1T TMO5-1T | T13-1T TMO13-1T |
|---|---|---|---|---|---|---|---|---|
| Imped. Ratio | 1 | 2 | 2.5 | 3 | 4 | 4 | 5 | 13 |
| Freq. (MHz) | 05-200 | 07-200 | 01-100 | 05-250 | 2-350 | 8-350 | 3-300 | 3-120 |
| T Model (10-49) | $3.95 | $4.25 | $4.25 | $3.95 | $2.95 | $4.95 | $4.25 | $4.25 |
| TMO model (10-49) | $6.45 | $6.75 | $6.75 | $6.45 | $4.95 | | $6.75 | $6.75 |

**UNBALANCED PRIMARY & SECONDARY**

| Model No. | T2-1 TMO2-1 | T3-1 TMO3-1 | T4-2 TMO4-2 | T8-1 TMO8-1 | T14-1 TMO14-1 |
|---|---|---|---|---|---|
| Imped. Ratio | 2 | 3 | 4 | 8 | 14 |
| Freq. (MHz) | 025-600 | 5-800 | 2-600 | 15-250 | 2-150 |
| T model (10-49) | $3.45 | $4.25 | $3.45 | $3.45 | $4.25 |
| TMO Model (10-49) | $5.95 | $6.95 | $5.95 | $5.95 | $6.75 |

**FT FTB**

| Model No. | FT1.5-1 | FTB1-1 | FTB1-6 | FTB1-1-75 |
|---|---|---|---|---|
| Imped. Ratio | 1.5 | 1 | 1 | 1 |
| Freq. (MHz) | 1-400 | 2-500 | 01-200 | 5-500 |
| (1-4) | $29.95 | $29.95 | $29.95 | $29.95 |

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INFO/CARD 27

November/December 1982

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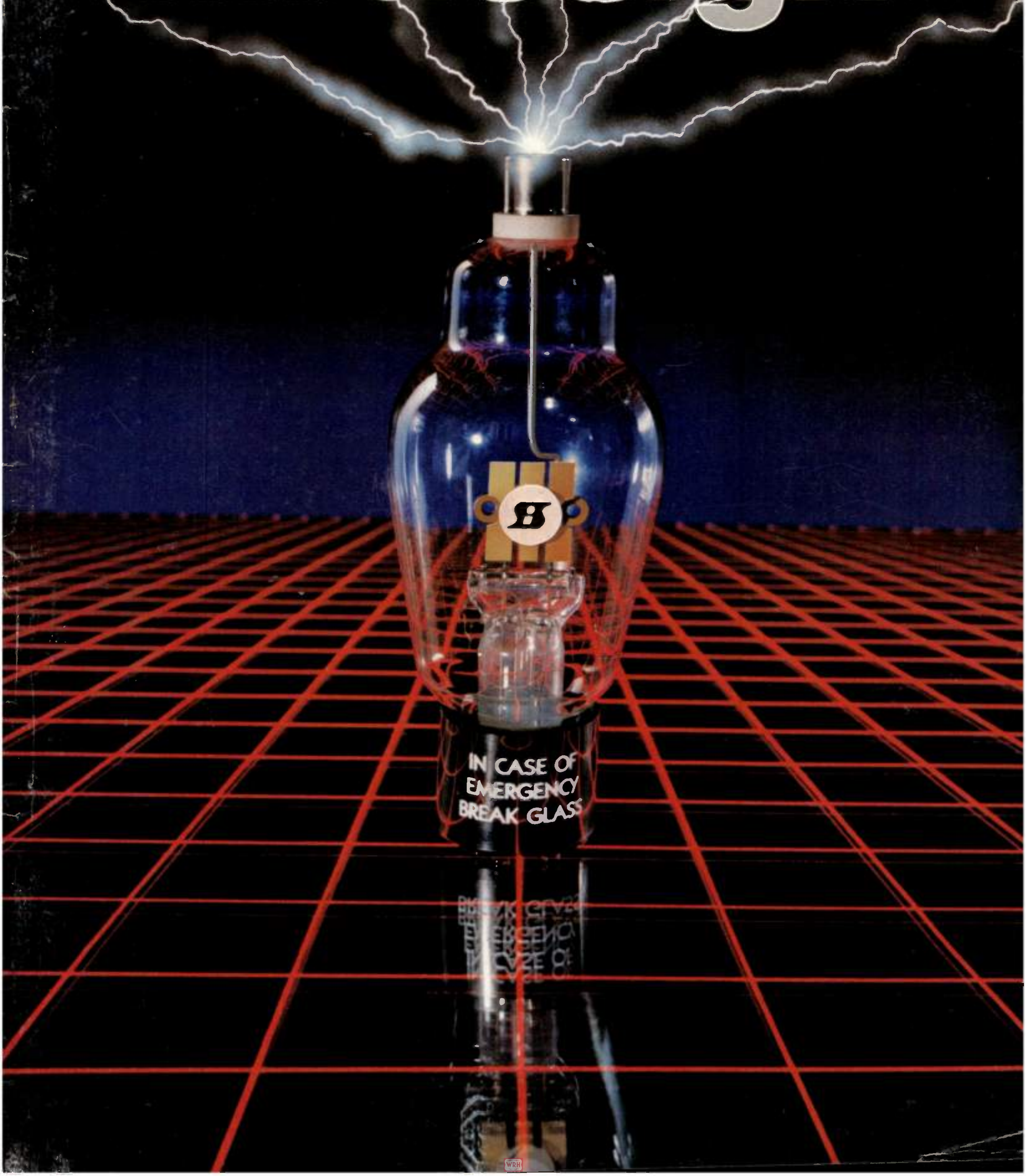
# r.f. design



# The Winning Combinations


These are just a few of the many possible winning combinations for low cost LNAs that can be designed with NEC microwave transistors.

The NE700, NE710 and NE720 GaAs FETs will provide excellent performance at substantially lower costs than ever before. The NE70083 and the NE71083 are designed for low noise applications from 4-18 GHz. The NE70083 has a 0.7 dB noise figure at 4 GHz and 1.9 dB at 12 GHz. The NE71083 has a 0.6 dB noise figure at 4 GHz and 1.6 dB at 12 GHz. These devices are excellent choices for 3.7-4.2 and 11.7-12.2 GHz TVRO applications. While the NE72089 is designed as a low cost general purpose FET, it has a 4 GHz noise figure of 1.0 dB, and makes an excellent local oscillator which produces 15 mW at 10 GHz.

The NE64535 and NE21935 are two of our most popular bipolar transistors. They are designed for low noise amplifier and oscillator applications from 500 MHz to 4 GHz.

All of the devices are available in chip form or in hermetically sealed packages. Grade CX screening is also available for military applications.

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# TM8 autoranging millivoltmeter

The Texscan TM8 is an autoranging, true r.m.s. millivoltmeter with a specified operating range of 10kHz to 1GHz with useful indications to 1500MHz. It measures r.f. voltages from 1mV to 3V full scale in eight ranges and may be operated automatically (in the autoranging mode), manually or remotely. The maximum voltage measured can be increased to 300V r.m.s. using the 100:1 divider provided and readings on any range can be held for at least three minutes by momentary depression of the probe switch.

Measurements are true r.m.s. but average responding can also be selected with a front panel switch. The meter reads 100$\mu$V–3V and also features a corresponding dB scale from −60 to +23dBm. A single logarithmic range is also provided and covers four decades scaled −30 to +10dBm. Fast or slow meter response may be selected to permit the observation or removal of rapid signal amplitude changes.

Remote programming is a standard feature and the socket is located on the rear panel. An output to drive a pen recorder is also provided from this socket and this can be used with an interface module, OB2, to connect to IEEE-488 bus systems.

The TM8 is supplied complete with probe (integral with input lead), probe to BNC adapter, 'T' connector, and a 100:1 high impedance divider.



**Texscan Instruments**
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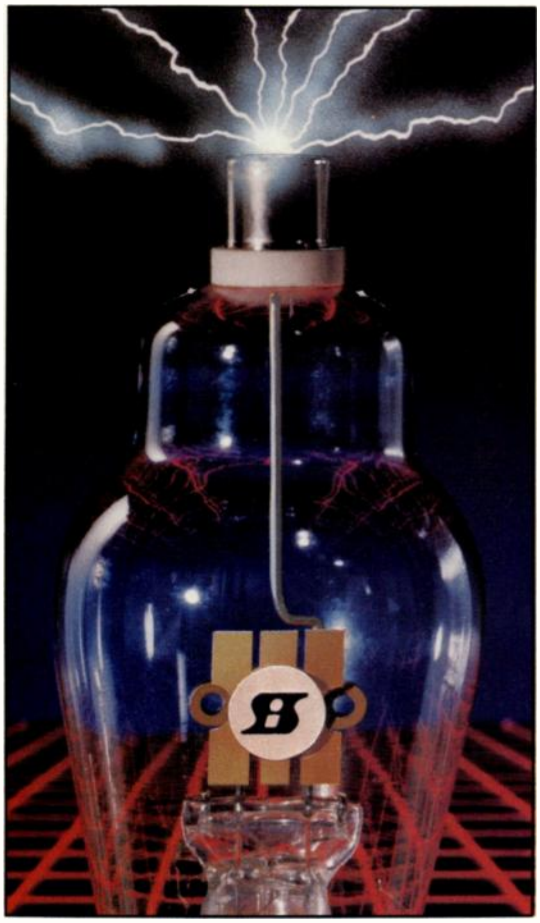
INFO/CARD 2

r.f. design

November/December 1982



**November/December Cover**—The cover, courtesy of Siliconix, Inc. amplifies the theme of their contributed article that high-voltage, active semiconductor devices now give the performance benefits of high-voltage RF tubes without compromises. **1**

**Loop Antenna Design**—This article presents an analysis of loop antenna characteristics to allow optimizing a design. **12**

**Various Power Gains and Their Meanings**—This article discusses the various choices of input and output power quantities, the various power gains thus formulated, and the physical meanings of the various power gains. **22**

**Power FETs for RF Amplifiers**—The second part of a two-part article demonstrates with a practical example how power FET characteristics can simplify circuit design. **30**

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November/December 1982, Volume 5, No. 6, r.f. design (ISSN 0163-321X) is published bi-monthly by Cardiff Publishing Company, a subsidiary of Cardiff Communications, Inc. 6430 S. Yosemite St., Englewood, Colo. 80111 (303) 694-1522.  Second Class postage paid at Englewood, Colorado, and additional mailing offices. 

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| | | | | DC-1000 | 1000-1500 | DC-1000 | 1000-1500 |
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| —AT—6 | 6 | ±0.3dB | DC-1500 | 0.6dB | 0.8dB | 1.3:1 | 1.5:1 |
| —AT—10 | 10 | ±0.3dB | DC-1500 | 0.6dB | 0.8dB | 1.3:1 | 1.5:1 |
| —AT—20 | 20 | ±0.3dB | DC-1500 | 0.6dB | 0.8dB | 1.3:1 | 1.5:1 |

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Subsidiary of Cardiff Communications, Inc.
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# letters

## Transistor Article Corrections

Peter Ledger of M/A-Com Silicon Products, Inc. points out that a couple of problems crept into his article "Slicing S-Parameters" which was presented in the Sept./Oct. 1982 issue. Specifically Table 3, line 16 of the COMPACT™ program should read:

CON BB T2 3 5

Also a line of the program was left out between lines 16 & 17 of the same table. Insert the following line at this point:

CON CC T2 3 4

A correct copy of the program is printed below for your convenience. —Ed.

**COMPACT COMPUTER PROGRAM FOR CALCULATION OF S-PARAMETERS**

| PROGRAM STEP | COMMENTS |
|---|---|
| RES AA SE 15 | RESISTANCE $R_b$, SERIES CONNECTED 15 ohms |
| CAP BB SE 4.0 | CAPACITANCE $C_{TE}$, SERIES CONNECTED 4 pF |
| RES CC SE 5.0 | RESISTANCE $r_e$, SERIES CONNECTED 5 ohms |
| RES DD SE 1.0 | RESISTANCE $R_e$, SERIES CONNECTED 1 ohm |
| CAP EE SE 0.250 | CAPACITANCE $C_{ce}$, SERIES CONNECTED 0.25 pF |
| CAP FF SE 0.016 | CAPACITANCE $C_c$, SERIES CONNECTED 0.016 pF |
| CAP GG SE 0.417 | CAPACITANCE $c_{bd}$, SERIES CONNECTED 0.417 pF |
| GEN HH CC .01 IE5 .990 10000 .00001E-6 | DEFINES THE GENERATOR |
| CAP II SE 0.380 | CAPACITANCE $C_{cep}$, SERIES CONNECTED 0.38 pF |
| SRL JJ SE 5.0 0.65 | SERIES CONNECTED $R_c$, 5.0 ohms. $L_c$, 0.65 nH |
| CAP KK SE 0.20 | CAPACITANCE $C_{bep}$, SERIES CONNECTED 0.2 pF |
| IND LL SE 1.0 | INDUCTANCE $L_b$, SERIES CONNECTED 1.0 nH |
| IND MM SE 0.225 | INDUCTANCE $L_e$, SERIES CONNECTED 0.225 nH |
| CAP NN SE 0.03 | CAPACITANCE $C_{bcp}$, SERIES CONNECTED 0.03 pF |
| CON AA T2 2 3 | CONNECTS AA BETWEEN NODES 2, 3 |
| CON BB T2 3 5 | CONNECTS BB BETWEEN NODES 3, 5 |
| CON CC T2 3 4 | CONNECTS CC BETWEEN NODES 3, 4 |
| CON DD T2 5 6 | CONNECTS DD BETWEEN NODES 5, 6 |
| CON EE T2 6 7 | CONNECTS EE BETWEEN NODES 6, 7 |
| CON FF T2 3 7 | CONNECTS FF BETWEEN NODES 3, 7 |
| CON GG T2 2 7 | CONNECTS GG BETWEEN NODES 2, 7 |
| CON HH T4 4 5 3 7 | CONNECTS HH BETWEEN NODES 4, 5 and 3, 7 |
| CON II T2 8 0 | CONNECTS II BETWEEN NODES 8, 0 |
| CON JJ T2 7 8 | CONNECTS JJ BETWEEN NODES 7, 8 |
| CON KK T2 1 0 | CONNECTS KK BETWEEN NODES 1, 0 |
| CON LL T2 1 2 | CONNECTS LL BETWEEN NODES 1, 2 |
| CON MM T2 6 0 | CONNECTS MM BETWEEN NODES 6, 0 |
| CON NN T2 1 8 | CONNECTS NN BETWEEN NODES 1, 8 |
| DEF AA T2 1 8 | DEFINES AA AS NODE 1 TO NODE 8 |
| PRI AA S1 50 | PRINTS S-PARAMETERS FOR NODES 1,8 |
| END | |
| 400 2000 100 | FREQUENCY RANGE 0.4 GHz to 2 GHz in 100 mHz STEPS |
| END | |

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INFO/CARD 4

# r.f. design

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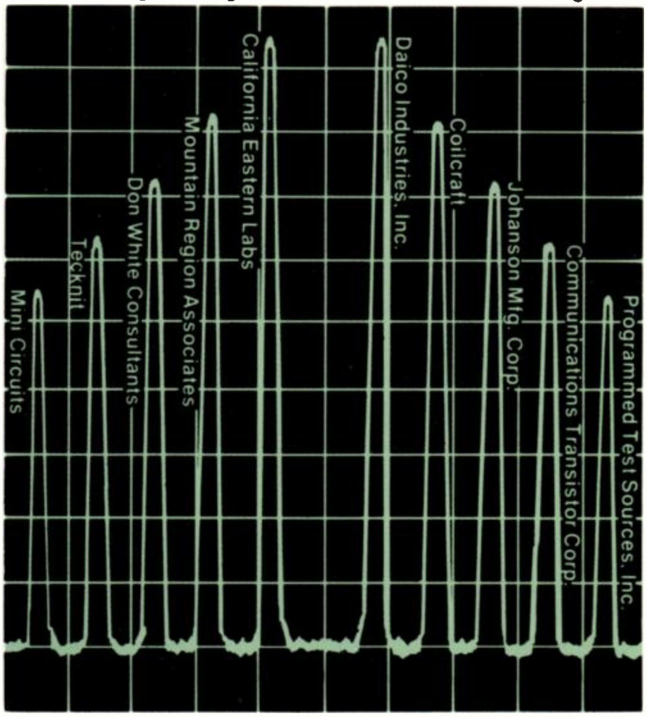


## The Winning Spectrum

***for July/August issue***

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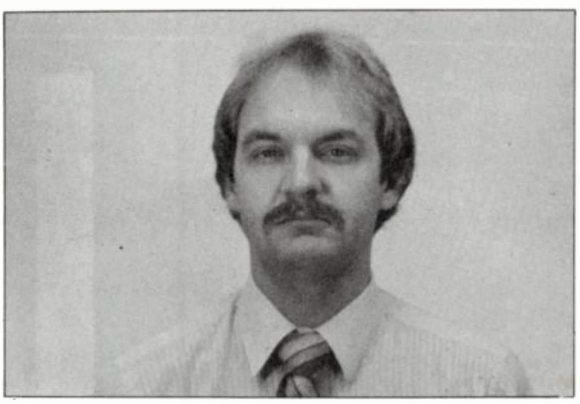

Tim Holmes is employed by RCA Consumer Electronics. He designs automatic test equipment for television product. Tim attended Purdue University and received his BSEE in 1979 and his MSEE in 1981. He is recently married and lives in Indianapolis, Indiana. His hobbies include photography, snow and water skiing, and watching automobile races.

# LOOP ANTENNA DESIGN

This article presents an analysis of loop antenna characteristics to allow optimizing a design.

*By Andrzej B. Przedpelski*
*A.R.F. Products, Inc.*
*R & D Laboratory*
*Boulder, Colorado*

There has been comparatively little activity in loop antenna design lately. However, particularly at low frequencies, the loop antenna has several advantages and may be the only viable solution. An example is the near magnetic field used in the early garage door openers[1]. A method will be described suitable for optimizing the design for maximum signal recovery over a specified frequency range or for maximizing the signal to noise ratio. It will be shown that the two designs are not necessarily compatible.

In the actual antenna design three factors, beside the mechanical configuration, have to be considered:

- signal recovery,
- bandwidth,
- signal to noise.

The relative importance of these factors has a pronounced effect on the overall design. Let's consider these factors individually. A ferrite rod loop antenna will be used in the analysis (Figure 1); however, the same analysis applies to air loops.

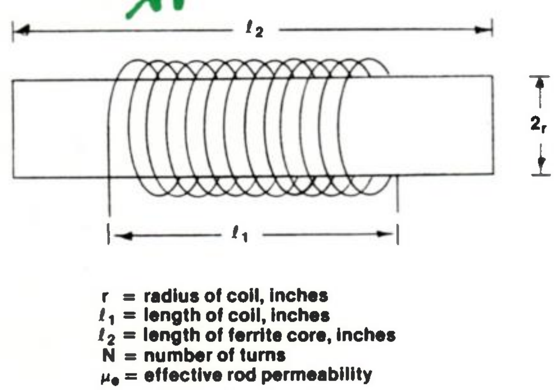


**Figure 1. Typical Ferrite Rod Loop Antenna**

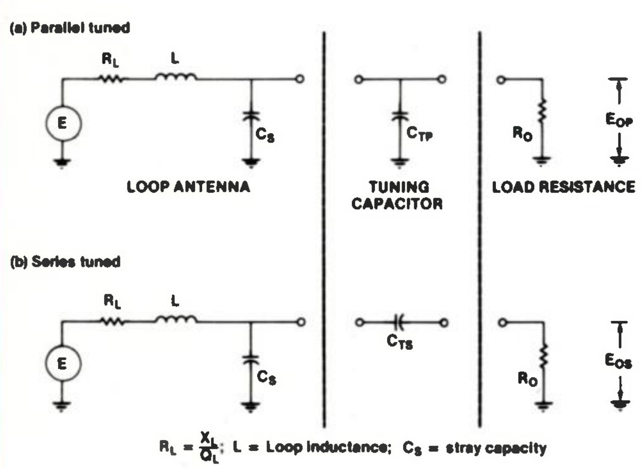


**Figure 2. Loop Antenna Equivalent Circuit**

## Signal Recovery

The maximum voltage induced in the loop, E, is:

$$E = 13.55 \times 10^{-12} \mu_e e N A f F_A \text{ volts} \quad (1)$$

where: e = field strength, volts/meter
$\mu_e$ = effective rod permeability
N = number of turns
A = area of loop, $\pi r^2$, in$^2$
f = frequency, Hz
$F_A$ = averaging factor

The averaging factor, $F_A$, is a function of loop configuration. Snelling[2] gives curves of $F_A$ for some typical cases. However, since $F_A$ varies between 1.0 and 0.83 in most instances, a close approximation is:

$$F_A = 1 - 0.17(\ell_1/\ell_2) \quad (2)$$

The effective rod permeability, $\mu_e$, is a factor of loop dimensions and is often overestimated. Curves of $\mu_e$ versus material permeability and form factor are available in the literature[3], but the approximate formula for even very high (>5000) material permeability,

$$\mu_e \approx 2.934\,(\ell_2/2r)^{1.394} \tag{3}$$

shows that the actual rod permeability is quite small. For instance, for $\ell_2/2r = 10$, maximum possible rod permeability, $\mu_e$, is only about 70.

It is necessary to tune the antenna to maximize the voltage developed across the load (Figure 2). Two possiblities exist: parallel or series tuning.

## Parallel Tuning (Figure 2a)

From Figure 3 it can be seen that the circuit gain can be calculated using:

$$\frac{E_{OP}}{E} = \frac{I(R_1 + jX_1)}{E} = \frac{R_1 + jX_1}{R_2 + jX_2} \tag{4}$$

For circuits with $Q_L > 5$ and $R_0 > 5\,X_c$, this expression is maximized when $X_c = X_L$ (within about 1%). Table I gives a program for the exact calculation of $E_{op}/E$. (It also calculates the value of $C_{TP}$ in register RO3.) At resonance, for values of $Q_L$ and $R_o$ within the above limits, equation (4) can be simplified to:

$$\left(\frac{E_{OP}}{E}\right)_{max} = \frac{R_O Q_L}{X_L Q_L + R_O} \tag{5}$$

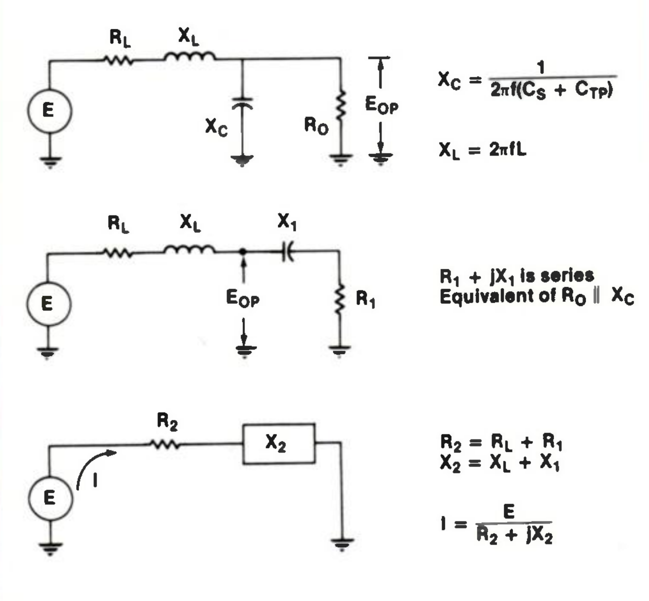


**Figure 3. Parallel Tuned Equivalent Circuits**

This approximation is within about 5% of the actual gain value, which is adequate for most applications. If the complete circuit Q, $Q_O$, can be measured, the following resonant gain equation also applies:

$$\left(\frac{E_{OP}}{E}\right)_{max} = Q_O \tag{6}$$

To illustrate the application of the preceeding equations assume the following conditions:

$R_o$ = 10,000 ohm
$f$ = 20,000 Hz
$L$ = 0.037 H
$Q_L$ = 50
$C_s = 100 \times 10^{-12}$ F
$e = 10 \times 10^{-6}$ V/M

$\ell_1$ = 7.5 in.
$\ell_2$ = 8.0 in.
$r$ = 0.625 in.
$N$ = 640 turns
$\mu_e$ = 17
$F_A$(calculated = 0.85 using Eq. 2)

From equation (1) we obtain the loop induced voltage:

$$E = 0.0308 \times 10^{-6}\ \text{volts}$$

Using program of Table I the tuning capacity, $C_{TP}$, can be calculated:

$$C_{TP} = 1612 \times 10^{-12}\ \text{F}$$

Using equation (4) the exact circuit gain can be calculated:

$$\left(\frac{E_{OP}}{E}\right)_{max} = 2.0617$$

```
01•LBL "LAP 1"
02 RCL 06
03 2
04 *
05 PI
06 *
07 STO 07
08 X↑2
09 RCL 04
10 *
11 1/X
12 RCL 02
13 -
14 STO 03
15 RCL 02
16 +
17 RCL 07
18 *
19 RCL 01
20 1/X
21 R-P
22 1/X
23 P-R
24 STO 08
25 X<>Y
26 CHS
27 STO 09
28 X<>Y
29 RCL 04
30 RCL 07
31 *
32 ST+ 09
33 RCL 05
34 /
35 RCL 08
36 +
37 RCL 09
38 X<>Y
39 R-P
40 1/X
41 X<>Y
42 CHS
43 RDN
44 RDN
45 R-P
46 R↑
47 *
48 RDN
49 /
50 R↑
51 END
```

| | |
|---|---|
| $R_{00}$ | |
| $R_{01}$ | $R_O$ |
| $R_{02}$ | $C_S$ |
| $R_{03}$ | $[C_{TP}]$ |
| $R_{04}$ | L |
| $R_{05}$ | $Q_L$ |
| $R_{06}$ | f |
| $R_{07}$ | $[2\pi f]$ |
| $R_{08}$ | [ ] |
| $R_{09}$ | [ ] |
| $R_{10}$ | |
| $R_{11}$ | |
| $R_{12}$ | |
| $R_{13}$ | |
| $R_{14}$ | |
| $R_{15}$ | |
| $R_{16}$ | |
| $R_{17}$ | |
| $R_{18}$ | |
| $R_{19}$ | |
| $R_{20}$ | |
| $R_{21}$ | |
| $R_{22}$ | |
| $R_{23}$ | |
| $R_{24}$ | |
| $R_{25}$ | |
| $R_{26}$ | |
| $R_{27}$ | |
| $R_{28}$ | |
| $R_{29}$ | |
| $R_{30}$ | |

**TABLE I.**

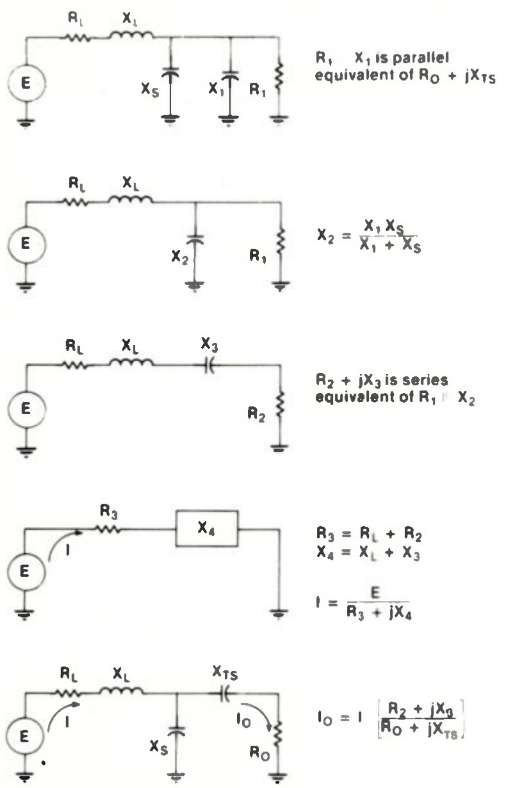


**Figure 4. Series Tuned Equivalent Circuits**

The approximate gain value using the simplified equation (5) is:

$$\left(\frac{E_{OP}}{E}\right)_{max} = 2.0620$$

or close enough for most applications.

Thus, the recovered voltage across the load, at resonance, is:

```
01*LBL "LAS 1"          28 STO 10
02 RCL 06               29 X<>Y
03 ST+ X                30 STO 11
04 PI                   31 RCL 04
05 *                    32 RCL 07
06 STO 07               33 *
07 RCL 03               34 STO 08
08 *                    35 +
09 1/X                  36 X<>Y
10 CHS                  37 RCL 08
11 STO 09               38 RCL 05
12 RCL 01               39 /
13 R-P                  40 +
14 1/X                  41 R-P
15 P-R                  42 STO 08
16 1/X                  43 RCL 09
17 X<>Y                 44 RCL 01
18 RCL 02               45 R-P
19 RCL 07               46 ST* 08
20 *                    47 RCL 11
21 CHS                  48 RCL 10
22 +                    49 R-P
23 X<>Y                 50 RCL 08
24 1/X                  51 /
25 R-P                  52 RCL 01
26 1/X                  53 *
27 P-R                  54 .END.
```

1/X
X→Y
RCL 02

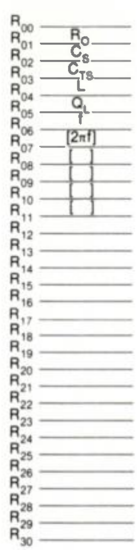


**TABLE II.**

$$\left(\frac{E_{OP}}{E}\right)_{max} E = E_{OP} = 0.0635 \times 10^{-6} \text{ volts}$$

## Series Tuning (Figure 2b)

The series tuned circuit can be similarly analyzed. The stray capacity, $C_s$, complicates calculations somewhat, as shown in Figure 4.

The gain of the circuit is:

$$\frac{E_{OS}}{E} = R_O \left[\frac{R_2 + jX_3}{(R_3 + jX_4)(R_O + jX_{TS})}\right] \qquad (7)$$

While it is possible to solve for the value of $C_{TS}$ to maximize the above expression, it is easier to use trial-and-error and the program of Table II. Selected values for $C_{TS}$ are stored in register 03 and the calculation performed until a maximum value of circuit gain is obtained. It can be seen from Figure 2b that with no stray capacity ($C_S = 0$) the maximum circuit gain occurs when the reactance of $C_{TS}$ is equal to the reactance of L. The maximum circuit gain obtainable under these conditions is always less than unity (unity for infinite $Q_L$).

For the same example used to illustrate the parallel tuned case, using equation (7) maximum gain is obtained for $C_{TS}$ = 1612 PF or the same as for the parallel tuned circuit. However, the circuit gain is only 1.0510, or considerably less. For large ratios of $C_{TS}/Cs$, the value of $C_{TS}$ will be the same as $C_{TP}$ (for the same circuit conditions) and gain will be less. Thus, it would seem that the series tuned circuit is not as desirable as the parallel tuned. That is not necessarily the case, however. It will be shown later that under some conditions the S/N ratio may be higher for the series tuned circuit even though the recovered signal is lower. In some cases, as the value of $C_s$ is artificially increased (and the value of $C_{TS}$ is lowered) a condition of high gain may exist. In the previous example, for instance, when $C_s$ is increased to 1500 PF and $C_{TS}$ is decreased to 215 PF the circuit gain becomes 5.0165, which is considerably more than was obtained in the parallel tuned case.

In the limiting case when $C_s$ is 1612 PF and $C_{TS}$ is infinity we obtain the equivalent parallel tuned circuit and a circuit gain of 2.0617.

## Bandwidth

While the signal recovery (effective height) of a loop antenna is important, its bandwidth may be another critical parameter. In some applications tuning of the antenna is either not desirable or possible, and the antenna has to be designed to cover the entire frequency range of interest. For maximum sensitivity the load resistance, $R_O$, should determine the bandwidth (Q) of the circuit, since additional loading to increase bandwidth would introduce unnecessary losses. The common method is to select a loop antenna and then increase or lower its impedance, by means of a transformer, until the desired performance is obtained. While usually satisfactory, this procedure introduces another critical component, the transformer, which has to be designed with the proper bandwidth. In addition, the transformer itself may introduce stray pick-up, since the operating signal levels are very low.

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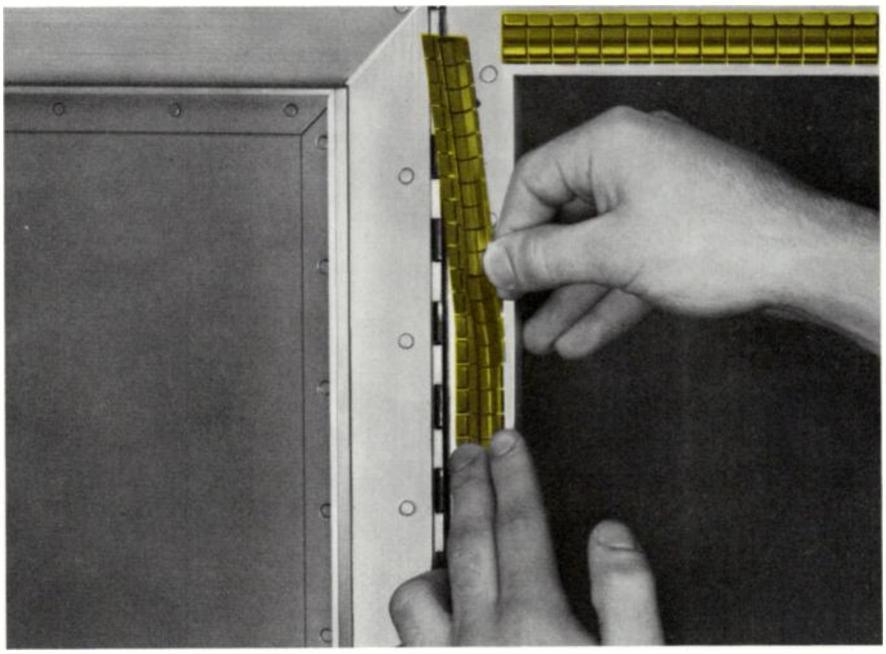

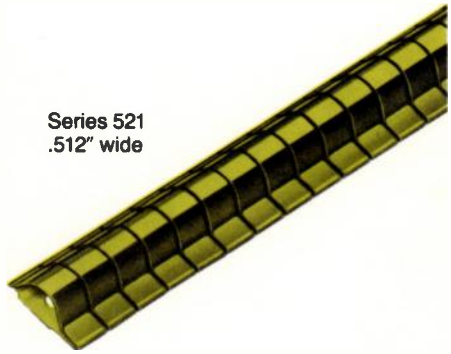

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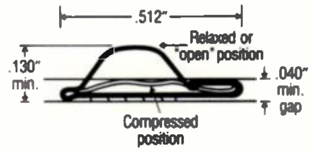


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```
01*LBL "LAF 2"
02 RCL 03
03 RCL 04
04 *
05 SQRT
06 STO 06
07 RCL 03
08 RCL 04
09 -
10 /
11 STO 02
12 1/X
13 RCL 05
14 1/X
15 -
16 RCL 01
17 *
18 STO 13
19 STO 19
20 RCL 06
21 PI
22 *
23 ST+ X
24 ST/ 19
25 RCL 13
26 *
27 1/X
28 STO 18
29 RCL 07
30 9
31 *
32 RCL 09
33 10
34 *
35 +
36 RCL 19
37 *
38 RCL 09
39 RCL 10
40 /
41 CHS
42 0.17
43 *
44 1
45 +
46 STO 12
47 *
48 RCL 06
49 /
50 SQRT
51 RCL 07
52 /
53 1 E3
54 *
55 STO 11
56*LBL 01
57 "F=?"
58 PROMPT
59 VIEW X
60 STO 13
61 RCL 07
62 X↑2
63 *
64 PI
65 *
66 RCL 11
67 *
68 RCL 08
69 *
70 13.55 E-12
71 *
72 RCL 12
73 *
74 STO 14
75 2
76 ST* 13
77 PI
78 ST* 13
79 RCL 13
80 RCL 18
81 *
82 RCL 01
83 1/X
84 R-P
85 1/X
86 P-R
87 STO 17
88 X<>Y
89 CHS
90 STO 16
91 X<>Y
92 RCL 19
93 RCL 13
94 *
95 ST+ 16
96 RCL 05
97 /
98 RCL 17
99 +
100 RCL 16
101 X<>Y
102 R-P
103 1/X
104 X<>Y
105 CHS
106 RDN
107 RDN
108 R-P
109 R↑
110 *
111 RDN
112 /
113 R↑
114 STO 15
115 RCL 14
116 *
117 LOG
118 20
119 *
120 VIEW X
121 STOP
122 GTO 01
123 .END.
```

| | |
|---|---|
| $R_{00}$ | |
| $R_{01}$ | $R_o$ |
| $R_{02}$ | $[Q_o]$ |
| $R_{03}$ | $f_2$ |
| $R_{04}$ | $f_1$ |
| $R_{05}$ | $Q_L$ |
| $R_{06}$ | $[f_o]$ |
| $R_{07}$ | $r$ |
| $R_{08}$ | $\mu_e$ |
| $R_{09}$ | $\ell_1$ |
| $R_{10}$ | $\ell_2$ |
| $R_{11}$ | $[N]$ |
| $R_{12}$ | $[F_A]$ |
| $R_{13}$ | $[2\pi f]$ |
| $R_{14}$ | $[E/e]$ |
| $R_{15}$ | $[E_o/E]$ |
| $R_{16}$ | $[\ ]$ |
| $R_{17}$ | $[\ ]$ |
| $R_{18}$ | $[C]$ |
| $R_{19}$ | $[L]$ |
| $R_{20}$ | |
| $R_{21}$ | |
| $R_{22}$ | |
| $R_{23}$ | |
| $R_{24}$ | |
| $R_{25}$ | |
| $R_{26}$ | |
| $R_{27}$ | |
| $R_{28}$ | |
| $R_{29}$ | |
| $R_{30}$ | |

**TABLE III.**

## Parallel Tuning

While the voltage induced in the loop increases with frequency (Equation (1)), the parallel tuned loop antenna equivalent circuit has basically a lowpass filter configuration (Figure (2a)). Thus these two effects can be made to produce an overall bandpass effect.

While it is possible to use a trial-and-error procedure and the program of Table I, it is more convenient to obtain all the design data from the program of Table III. Using as inputs the physical characteristics of the loop, the center frequency, the required circuit Q (a function of bandwidth), the expected coil Q, and the load resistance, the program gives the number of turns, the coil inductance and the total tuning capacity ($C_s + C_{TP}$). First the center frequency is calculated. As with most filters, the geometric mean gives the best results:

$$f_o = \sqrt{f_2 f_1} \tag{8}$$

Then circuit Q is calculated

$$Q_o = \frac{f_o}{f_2 - f_1} \tag{9}$$

Using (8) and (9) and

$$X_L = R_o\left(\frac{1}{Q_o} - \frac{1}{Q_L}\right) \tag{10}$$

and

$$L = \frac{X_L}{2\pi f_o} \tag{11}$$

the program calculates the inductance of the coil, L, and stores it in register 04. The required number of turns is then calculated using:

$$L = \mu_e F_A \frac{(rN)^2}{9r + 10\ell_1} \times 10^{-6}\ \mathrm{H} \tag{12}$$

and is stored in register 11. The total tuning capacity is also calculated and stored in register 18. LBL 01 of the program will then calculate the voltage induced in the loop for a given field strength (E/e), store it in register 14, calculate circuit gain ($E_o/E$), store it in register 15 and finally calculate the overall "gain" ($E_o/e$) in dB. (While it is not strictly correct to express $E_o/e$ in dB, since the units for $E_o$ and e are not the same, it is convenient for overall system calculations.) By repeating LBL 01, for different frequencies, a complete frequency response can be obtained.

Let us consider the following example:

$f_2$ = 30 KHz
$f_1$ = 10 KHz
$R_o$ = 5000 ohm
$\ell_1$ = 7.0 inches
$\ell_2$ = 8.0 inches
r = 0.375 inches
$\mu_e$ = 60
$Q_L$ = 10

Using program of Table III we obtain:

$f_o$ = 17.32 KHz (REG 06)
$Q_o$ = 0.866 (REG 02)
L = 0.0485 H (REG 19)
C = $1742 \times 10^{-12}$ F (REG 18)
N = 599 turns (REG 11)

Using LBL 01 of Table III, the frequency response can be then calculated, and is shown in Table IV. The calculated frequency response is within < 0.2 dB of that required, which is quite adequate for this large a bandwidth.

## Series Tuning

Series tuned antenna does not provide a suitable bandpass response. The E/e response is the same as for the parallel tuned case, but $E_o/E$ has a bandpass characteristic; thus, the overall response has a highpass rather than bandpass shape. This is shown in Figure 5. The same loop antenna as in the above example was used. The total capacity of 1742 PF was divided into $C_s$ = 100 PF and $C_{TS}$ = 1642 PF. The calculated response using equations (1) and (7) is shown in Table V.

| Frequency | E/e | $E_o/E$ | $E_o/e$(dB) | Freq. Resp. (dB) |
|---|---|---|---|---|
| 5000 Hz | $9.16 \times 10^{-4}$ | 1.00 | −60.74 | −9.48 |
| 10,000 | $1.83 \times 10^{-3}$ | 1.03 | −54.49 | −3.23 |
| 15,000 | $2.75 \times 10^{-3}$ | 0.96 | −51.61 | −0.35 |
| 17,320 | $3.17 \times 10^{-3}$ | 0.86 | −51.26 | 0 (ref) |
| 20,000 | $3.66 \times 10^{-3}$ | 0.73 | −51.44 | −0.18 |
| 25,000 | $4.58 \times 10^{-3}$ | 0.51 | −52.65 | −1.39 |
| 30,000 | $5.49 \times 10^{-3}$ | 0.36 | −54.14 | −2.88 |
| 35,000 | $6.41 \times 10^{-3}$ | 0.26 | −55.53 | −4.27 |

**Table IV. Parallel Tuned Antenna Frequency Response.**

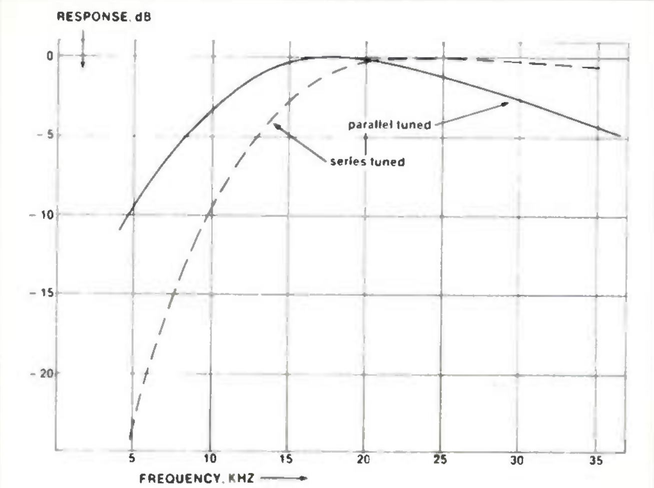


**Figure 5. Comparison of Parallel and Series Tuned Antenna Response**

## Signal to Noise Ratio

So far only the signal recovery characteristics of a loop antenna were considered. In most cases, however, the signal to noise ratio of the recovered signal is of more importance. To simplify the analysis, a noiseless amplifier is assumed, the amplifier input thermal noise being generated in the input resistance $R_o$. To permit a complete analysis "noise" has to be also defined. Dr. Hubert[4] points out that noise in RLC circuits is not in general white and has to be integrated over the bandwidth of interest:

$$\varrho_n = \sqrt{\int_{\omega_1}^{\omega_2} \overline{i_n}^2 |Z_T(\omega)| d\omega} \qquad (13)$$

where $\overline{i_n}^2$ = total noise current density, $A^2/Hz$

$Z_T(\omega)$ = total impedance (function of frequency)

Since the bandwidth of interest is not necessarily the noise bandwidth of the antenna circuit, noise density rather than total noise will be used in signal to noise calculations. This is reasonable, since these antennas are often used to receive several narrow band signals (later separated by filters) and the total noise over the narrow bandwidth can be closely approximated by multiplying the noise density by $\sqrt{\text{bandwidtn}}$. Again, the parallel tuned and the series

| Frequency | $E_o/E$ | $E_o/e$(dB) | Freq. Resp. (dB) |
|---|---|---|---|
| 5000 Hz | 0.27 | – 72.12 | – 22.75 |
| 10,000 | 0.60 | – 59.16 | – 9.79 |
| 15,000 | 0.91 | – 52.05 | – 2.68 |
| 17,320 | 0.95 | – 50.44 | – 1.07 |
| 20,000 | 0.91 | – 49.59 | – 0.22 |
| 25,000 | 0.74 | – 49.37 | 0 (ref) |
| 30,000 | 0.60 | – 49.62 | – 0.25 |
| 35,000 | 0.50 | – 49.87 | – 0.50 |

**Table V. Series Tuned Antenna Frequency Response.**

```
01•LBL "LAP 3"
02 RCL 06
03 ST+ X
04 PI
05 *
06 STO 07
07 RCL 04
08 *
09 STO Y
10 RCL 05
11 /
12 R-P
13 1/X
14 P-R
15 STO 08
16 1/X
17 X<>Y
18 RCL 02
19 RCL 03
20 +
21 RCL 07
22 *
23 CHS
24 +
25 RCL 01
26 1/X
27 RCL 08
28 +
29 R-P
30 1/X
31 RCL 01
32 1/X
33 RCL 08
34 +
35 4
36 *
37 1.38 E-23
38 *
39 300
40 *
41 SQRT
42 *
43 .END.
```

| | |
|---|---|
| $R_{00}$ | |
| $R_{01}$ | $R_O$ |
| $R_{02}$ | $C_S$ |
| $R_{03}$ | $C_{TP}$ |
| $R_{04}$ | L |
| $R_{05}$ | $Q_L$ |
| $R_{06}$ | f |
| $R_{07}$ | [2πf] |
| $R_{08}$ | [ ] |
| $R_{09}$–$R_{30}$ | |

**TABLE VI.**

```
01•LBL "LAS 2"
02 RCL 06
03 ST+ X
04 PI
05 *
06 STO 07
07 RCL 04
08 *
09 STO Y
10 RCL 05
11 /
12 R-P
13 1/X
14 P-R
15 X<>Y
16 RCL 02
17 RCL 07
18 *
19 -
20 X<>Y
21 R-P
22 1/X
23 P-R
24 X<>Y
25 RCL 03
26 RCL 07
27 *
28 1/X
29 -
30 X<>Y
31 R-P
32 1/X
33 P-R
34 STO 08
35 RDN
36 RCL 01
37 1/X
38 RCL 08
39 +
40 R-P
41 1/X
42 RCL 01
43 1/X
44 RCL 08
45 +
46 4
47 *
48 1.38 E-23
49 *
50 300
51 *
52 SQRT
53 *
54 .END.
```

| | |
|---|---|
| $R_{00}$ | |
| $R_{01}$ | $R_O$ |
| $R_{02}$ | $C_S$ |
| $R_{03}$ | $C_{TP}$ |
| $R_{04}$ | L |
| $R_{05}$ | $Q_L$ |
| $R_{06}$ | f |
| $R_{07}$ | [2πf] |
| $R_{08}$ | [ ] |
| $R_{09}$–$R_{30}$ | |

**TABLE VII.**

tuned antenna will be considered separately. As an example let's assume the following conditions:

$R_o$ = 10,000 ohm  r = 0.375
$C_s = 100 \times 10^{-12}$ F  $\mu_e = 60$
$C_{TP} = C_{TS} = 1768 \times 10^{-12}$ F  $\ell_1 = 7$
$L = 15.5 \times 10^{-3}$ H  $\ell_2 = 8$
$Q_L = 20$  N = 339

(These figures were obtained using program of Table III and $f_1$ = 25,000 Hz and $f_2$ = 35,000 Hz to obtain a representative antenna design). The requirement is to determine the signal to noise ratio at several frequencies over the antenna passband with a field strength of 100 microvolts per meter, and a noise bandwidth of 100 Hz. Since the antenna bandwidth is rather large and the noise bandwidth small, the proposed



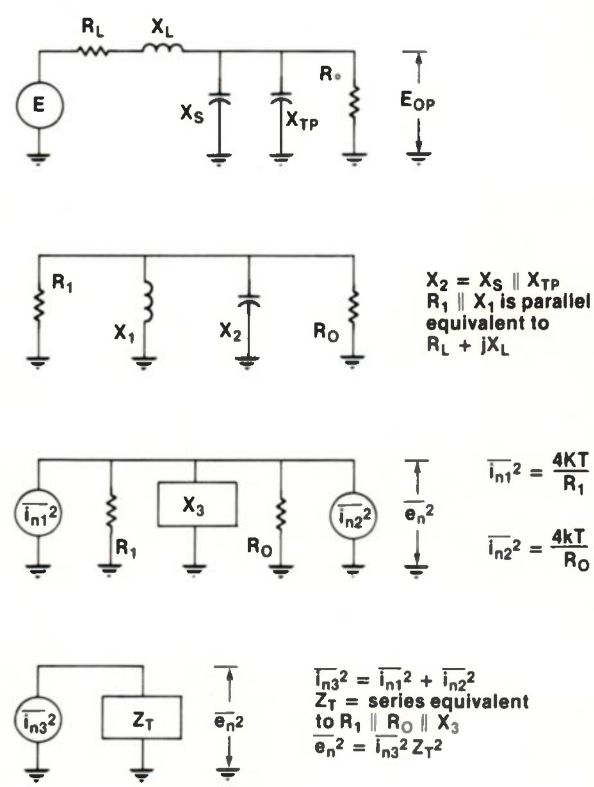


**Figure 6. Parallel Tuned Antenna Equivalent Noise Circuit**

| Frequency | $E_O/e$ | Signal | Noise | S/N |
|---|---|---|---|---|
| 25,000 Hz | −43.96 dB | $6.35 \times 10^{-7}$ | $8.44 \times 10^{-9}$ | 37.53 dB |
| 29,600 | −40.86 | $9.05 \times 10^{-7}$ | $1.19 \times 10^{-8}$ | 37.62 |
| 30,000 | −40.88 | $9.04 \times 10^{-7}$ | $1.18 \times 10^{-8}$ | 37.69 |
| 35,000 | −43.80 | $6.45 \times 10^{-7}$ | $8.37 \times 10^{-9}$ | 37.74 |
| 20,000 | −49.15 | $3.49 \times 10^{-7}$ | $4.74 \times 10^{-9}$ | 37.34 |
| 40,000 | −47.13 | $4.40 \times 10^{-7}$ | $5.66 \times 10^{-9}$ | 37.81 |

**Table VIII. Parallel Tuned Antenna Signal and Noise Characteristics.**

approximation of calculating signal to noise density ratio and multiplying by $\sqrt{100}$ should be valid.

## Parallel Tuning

The method of calculating the recovered signal has been already described. To calculate the noise voltage density, the total noise current density and total impedance method[4] will be used. The method of arriving at the parallel tuned loop antenna equivalent noise circuit is shown in Figure 6. Series-parallel and parallel-series transformations are used to allow simplification of the circuit. The noise voltage density, at any given frequency, is

$$\overline{e_n} = \overline{i_n}|Z_T| \tag{14}$$

The program of Table VI simplifies the calculations. The results are shown in Table VIII. (The signal levels were calculated using previously shown methods). From these

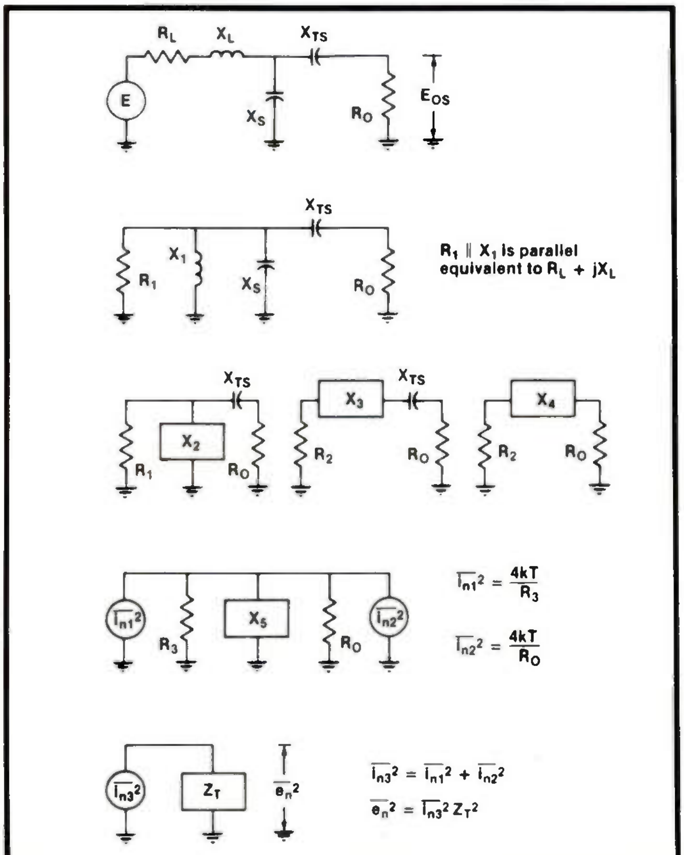


**Figure 7. Series Tuned Antenna Equivalent Noise Circuit**

| Frequency | E | $E_O/E$ | Signal | Noise | S/N |
|---|---|---|---|---|---|
| 25,000 Hz | $2.59 \times 10^{-7}$ | 1.02 | $2.64 \times 10^{-7}$ | $1.99 \times 10^{-9}$ | 42.46 dB |
| 29,600 | $3.07 \times 10^{-7}$ | 1.04 | $3.19 \times 10^{-7}$ | $1.62 \times 10^{-9}$ | 45.89 |
| 30,000 | $3.11 \times 10^{-7}$ | 1.04 | $3.24 \times 10^{-7}$ | $1.64 \times 10^{-9}$ | 45.91 |
| 35,000 | $3.62 \times 10^{-7}$ | 1.05 | $3.81 \times 10^{-7}$ | $2.27 \times 10^{-9}$ | 44.50 |

**Table IX. Series Tuned Antenna Signal and Noise Characteristics.**

| Frequency | $E_O/E$ | Signal | Noise | S/N |
|---|---|---|---|---|
| 25,000 Hz | 1.00 | $2.59 \times 10^{-7}$ | $3.43 \times 10^{-9}$ | 37.56 dB |
| 29,600 | 1.00 | $3.07 \times 10^{-7}$ | $4.01 \times 10^{-9}$ | 37.68 |
| 30,000 | 1.00 | $3.11 \times 10^{-7}$ | $4.06 \times 10^{-9}$ | 37.68 |
| 35,000 | 1.00 | $3.62 \times 10^{-7}$ | $4.69 \times 10^{-9}$ | 37.75 |

**Table X. Untuned Antenna Signal and Noise Characteristics.**

results it can be seen that the expected S/N ratio for a 100 microvolt/meter signal and 100 Hz bandwidth will be in the 17-18 dB range. Note that while the signal peaks in the middle of the antenna bandwidth, so does the noise, and the resultant S/N remains quite constant.

## Series Tuning

The calculation of noise voltage density for the series tuned antenna is more complicated because of the stray capacity, $C_s$. However, the same method of circuit simplification can be used, as shown in Figure 7. The program of Table VII performs the required calculations. Using our example the figures shown in Table IX can be obtained. Using the series tuned antenna, the noise dips in the middle of the passband and is lower than in the parallel tuned case resulting in higher S/N even though the recovered signal is lower. The expected S/N for the 100 Hz bandwidth of the example will be in the 22-26 dB range over the 25-35 KHz frequency range.

## Untuned Antenna

In view of the somewhat unexpected results, it may be interesting to see how an untuned antenna would perform. The calculations can be performed using the series tuned antenna programs and a large value for $C_{TS}$ (1 F, for instance). The results are shown in Table X. The resultant S/N characteristics are very similar to the parallel tuned case; however, both signal and noise levels are lower.

## Summary

The presented analysis of loop antenna characteristics indicate that, in general, one should not generalize in loop antenna design, but each case should be individually examined and all pertinent factors considered.

Some trends are apparent, however, and are shown in Table XI. These should be used with caution, confirming the characteristics with actual calculations. In addition, the following characteristics should also be considered in the circuit design, but also only as starting points:

- If a well controlled broadband 3dB passband signal "gain" (output voltage/field strength) is required consider the parallel tuned design.
- If maximum S/N is required for narrow band channels give the series tuned design a try.
- If a constant S/N for several narrowband channels is needed, the parallel tuned or untuned designs may be best.
- While the bandwidth of a parallel tuned antenna (in terms of signal recovery) is well defined, its S/N bandwidth (constant S/N for narrowband channels) is much larger.
- Because of the above characteristic, the parallel tuned loop antenna bandwidth can be made quite large by shaping frequency response after first low level amplification, since this would have no effect of S/N.
- To increase gain-bandwidth maximize E/L ratio. □

## References


1. J.J. Hupert, "Remote Control by 'Near' Magnetic Field," Electronic Industries, July 1959.
2. E.C. Snelling, "Soft Ferrites, Properties and Applications," C.R.C. Press, Cleveland, OH, 1969.
3. "Ferrite Cores and Assemblies for the Electronic Industry," Fair-Rite Products Corp., Wallkill, NY 12589, March 1980.
4. Julius J. Hupert, "Noise in Electronic Amplifiers", Electro-Technology, March 1966.


| | UNTUNED | PARALLEL TUNED | SERIES TUNED |
|---|---|---|---|
| SIGNAL RECOVERY | Lowest (increases with frequency) | High (peaks in middle of passband) | Low (increases with frequency) |
| NOISE DENSITY | Low (increases with frequency) | High (peaks in middle of passband) | Lowest (dips in middle of passband) |
| S/N | Low (constant with frequency) | Low (constant with frequency) | High (varies with frequency) |

**Table XI. Trends in Loop Antenna Characteristics.**


Errata Jan/Feb 83

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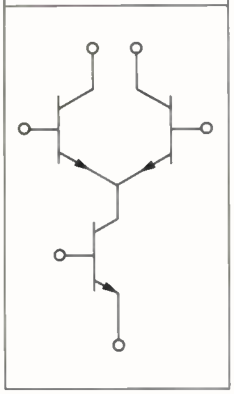

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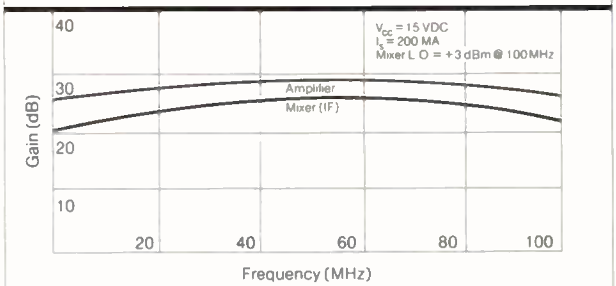
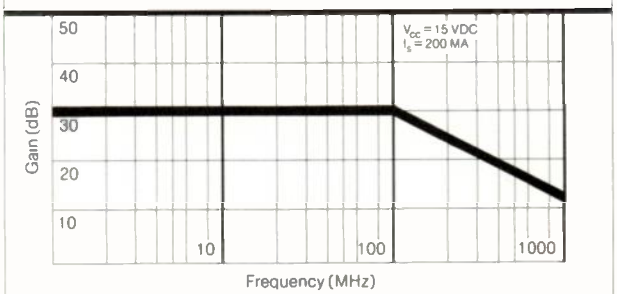



**Maximum Device Gain vs. Frequency (Typical)**

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# VARIOUS POWER GAINS & THEIR MEANINGS

Power gain is defined as the ratio of the output power to the input power. This article discusses the various choices of input and output power quantities, the various power gains thus formulated, and the physical meanings of the various power gains.

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*School of Electrical Engineering*
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*Atlanta, Georgia*

Any power gain can be defined as the ratio of a power quantity associated with the output of an amplifier or network to a power quantity associated with the input of that amplifier or network. The input and output power quantities forming the power gains are chosen to best represent some particular power-related aspect of amplifier performance; however, it is not necessary that either the input or output quantity correspond to an actual power in the circuit.

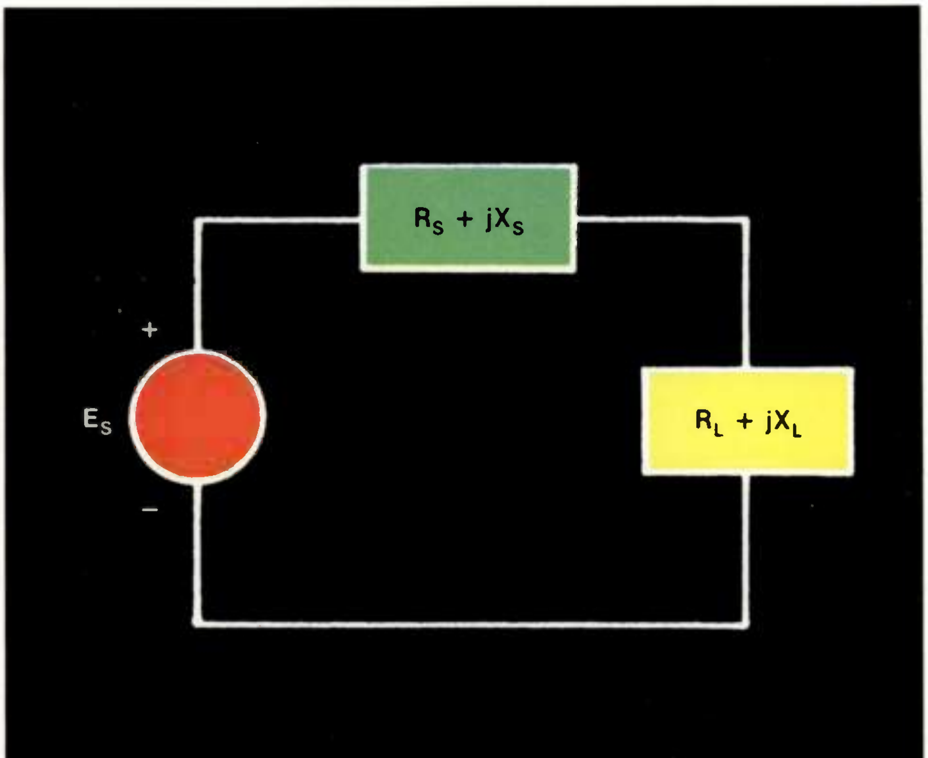


Figure 1. The circuit used for calculating $P_{AVS}$.

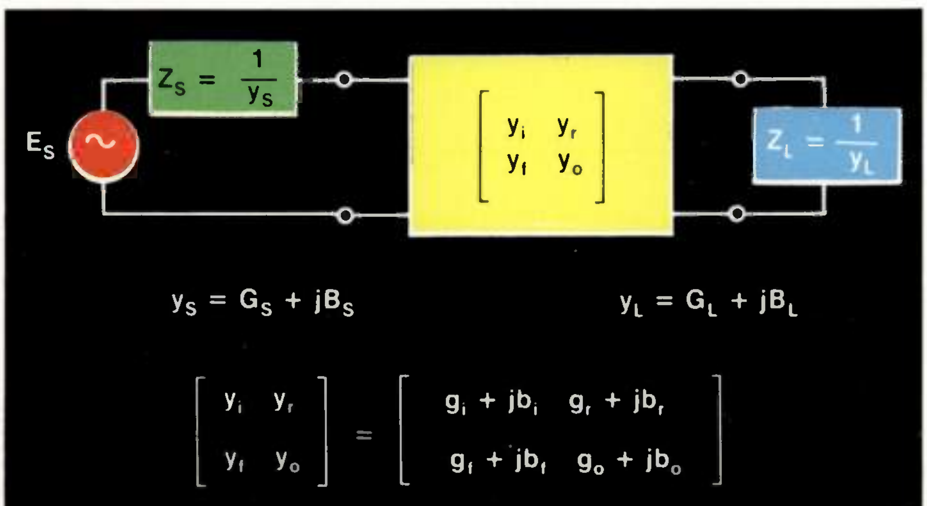


Figure 2. Definitions of quantities used in the power gain discussions.

## Available Power

All rf sources have associated with them a non-zero source resistance. The value of the resistance may be adjusted by matching networks, but it can never be eliminated. In order to be compatible with transmission line systems, most low power rf sources can be assumed to have a resistance equal to the system characteristic impedance, typically 50 or 75 ohms. Since the source resistance cannot be eliminated, its value determines the maximum power that can be obtained from the source. To examine the effect of this resistance, consider the circuit in Figure 1. The power delivered to the load is

$$P_L = \frac{|E_S|^2 R_L}{(R_S + R_L)^2 + (X_S + X_L)^2} \tag{1}$$

The maximum power occurs when $Z_L = Z_S^*$

$$P_L(\max) = P_{avs} = \frac{|E_S|^2}{4R_S} \tag{2}$$

The maximum power that can be delivered by a source is called the power available from the source, $P_{avs}$. Note that the voltage source and source impedance in Figure 1 could represent the Thevenin equivalent circuit of the output of an amplifier or other device. In such a case, the available power would represent the power available at the output of the amplifier.

In actual operation, a circuit seldom absorbs or delivers an available power. However, the power available from the source and the power available at the output represent maximum power values that could be absorbed or delivered and thus are valuable as reference quantities for power gain specifications.

## The Several Power Gains

There are four power gains: the transducer power gain, $G_T$; the operating power gain, $G_P$; the available power gain, $G_A$; and the insertion power gain, $G_I$. Each of these gains is particularly useful in certain circumstances. Depending upon the state of stability of the amplifier, some or all of the power gains may not exist. The question of stability is examined later in this article. For the following discussion of the definitions of the power gains, it is assumed that the amplifier or network is inherently or unconditionally stable and thus all of the power gains exist. The various power gains will be defined with reference to Figure 2. Admittance parameters will be used in the following discussion, but the observations and conclusions are applicable to any other parameter set, including S-parameters.

The transducer power gain is defined as

$$G_T = \frac{P_L}{P_{avs}} \tag{3}$$

Where $P_L$ is the actual power delivered to the load and $P_{avs}$ is the power available from the source. The transducer power gain is often the most useful gain for design engineers because it includes the effects of both the input and output terminations of the amplifier. If the output circuit is mismatched, the power delivered to the load will decrease resulting in reduced transducer gain. If a mismatch occurs at the input, the power delivered to the input will be less than the power available from the source which will result in an output that is less than when the available power is present at the input. The transducer gain will measure this decreased power output.

This symmetry of treatment is seen in the equation for the transducer power gain in terms of the y-parameters.

$$G_T = \frac{4|y_f|^2 G_S G_L}{|(y_i + Y_S)(y_o + Y_L) - y_r y_f|^2} \tag{4}$$

Note that the equation is unaffected if the input and output quantities are interchanged. This symmetrical treatment of input and output ports is the most useful property of the transducer power gain.

## Operating Power Gain

The operating power gain is defined as

$$G_P = \frac{P_L}{P_{IN}} \tag{5}$$

Where $P_{IN}$ is the power delivered to the input port of the amplifier. The operating power gain is often thought of as the most "physical" or most easily understood of the several power gains. This gain function is useful in several design situations, but in all it is probably not as valuable as the transducer power gain. It is easily shown that the operating power gain can be interpreted as a special case of the transducer power gain. If the input circuit is matched, then the actual power delivered to the input is all of the power that is available from the source or, $P_{IN} = P_{avs}$. Applying this condition to the equation for $G_P$,

$$G_P = \frac{P_L}{P_{IN}} = \left.\frac{P_L}{P_{avs}}\right|_{\substack{P_{avs} = P_{IN} \\ \text{or} \\ \text{input matched}}} = G_T\Big|_{\text{input matched}} \tag{6}$$

Thus, the operating power gain is equal to the transducer power gain when a conjugate match is enforced at the input. This means that the operating power gain and transducer power gain can be equal only when the input is conjugately matched. It is also correct to state that the operating power gain is affected by only what occurs at the output port and ignores the input. This is evident upon examination of the y-parameter equations for the operating power gain.

$$G_P = \frac{|y_f|^2 G_L}{|y_o + Y_L|^2 \operatorname{Re}\left\{y_i - \dfrac{y_r y_f}{y_o + Y_L}\right\}} \tag{7}$$

Note that only output circuit quantities appear in the equation.

## Available Power Gain

While the operating power gain tells us nothing about the input circuit, its dual quantity, the available power gain, gives no information about the output circuit. The available power gain is defined as,

$$G_A = \frac{P_{avo}}{P_{avs}} \tag{8}$$

Where $P_{avo}$ is the power available at the output. This power is the maximum that could be delivered by the output port under any conditions of termination. The available power gain can also be considered to be a special case of the transducer power gain by noting that when the output port is matched it receives the power available from the output.

$$G_A = \frac{P_{avo}}{P_{avs}} = \left.\frac{P_L}{P_{avs}}\right|_{\substack{P_L = P_{avo} \\ \text{or} \\ \text{input matched}}} = G_T\Big|_{\text{output matched}} \tag{9}$$

Hence the available power gain can be interpreted as the transducer power gain when a conjugate match is enforced at the output. Consequently, the available power gain can be equal to the transducer power gain only when the output port is conjugately matched. The available power gain can be written in terms of the y-parameters as,

$$G_A = \frac{|y_f|^2 G_S}{|y_i + Y_S|^2 \operatorname{Re}\left\{y_o - \dfrac{y_r y_f}{y_i + Y_S}\right\}} \tag{10}$$

This equation is seen to be identical in form to that for the operating power gain with the input and output quantities interchanged. Hence, the available power gain is the dual of the operating power gain. The operating power gain describes only what occurs at the output port and the available power gain describes only what occurs at the input port. Since the noise behavior of an amplifier is determined by its input termination, the available power gain is very useful when optimizing the overall noise figure of an amplifier.

## Insertion Power Gain

The last gain of interest is the insertion power gain. This gain, which is probably less useful than the other three, is defined as

$$G_I = \frac{P_L}{P_{DL}} \qquad (11)$$

Where $P_{DL}$ is the power delivered to the load when it is directly connected to the source with no attempt made to match the source and load. This gain would be useful in situations in which matching networks could not be used. The equation for the insertion power gain can be written in terms of the y-parameters as,

$$G_I = \frac{|y_f|^2|Y_S + Y_L|^2}{|(y_i + Y_S)(y_o + Y_L) - y_r y_f|^2} \qquad (12)$$

Note that if the load and source admittances are complex conjugates, the insertion power gain reduces to the transducer power gain.

## Effect of Stability on Gain

The existence of the several power gains is dependent upon the condition of stability of the amplifier. A power gain has meaning only when the output power quantity of the amplifier vanishes when the input power quantity is zero. An unstable amplifier is one that produces an output without the necessity of an input, i.e., it functions as a power source or oscillator. The origin of this instability in an amplifier is the presence of negative conductance or resistance at the input or output of the amplifier. If enough negative conductance is present to cancel the positive conductance of the input or output circuit, then the amplifier will produce a response with no input. Thus a particular gain function does not exist if the conditions under which it is defined results in a net zero or negative conductance.

The transducer power gain, which considers both ports simultaneously, exists if the net conductance at each port is positive or,

$$G_S + G_{IN} = G_S + R_e(Y_{IN}) > 0 \qquad (13)$$

where

$$Y_{IN} = y_i - \frac{y_r y_f}{y_o + Y_L} \qquad (14)$$

and

$$G_L + G_{out} = G_L + R_e(Y_{out}) > 0 \qquad (15)$$

where,

$$Y_{OUT} = y_o - \frac{y_r y_f}{y_i + Y_S} \qquad (16)$$

If these conditions are satisfied, the circuit is stable and the transducer power gain has meaning. A more restrictive, but more useful condition for the existence of the transducer power gain is that

$$g_i > 0 \qquad (17)$$

$$g_o > 0 \qquad (18)$$

and that Stern's stability factor k

$$k = \frac{2(g_i + G_S)(g_o + G_L)}{Re(y_r y_f) + |y_r y_f|} \qquad (19)$$

is greater than unity. Stern's factor assumes that the imaginary parts of $Y_{IN}$ and $Y_{OUT}$ are conjugately matched to $Y_S$ and $Y_L$, respectively, thus giving a worst-case condition and making the stability dependent upon only $G_L$ and $G_S$. Equations 17-19 are sufficient to ensure stability, but are not necessary since they do not consider the effect of specific source and load susceptances. Note that if Stern's factor is greater than unity, only the transducer power gain is guaranteed to exist. The available and operating power gains may or may not exist.

Recall that the operating power gain can be interpreted as the transducer power gains with a conjugate match enforced at the input port. The conjugate match makes $G_S = G_{IN}$ and $B_S = -B_{IN}$. Thus if $G_{IN}$ is less than or equal to zero the operating power gain does not exist. Mathematically, the operating power gain exists whenever,

$$G_{IN} = Re\left\{ y_i - \frac{y_r y_f}{y_o + Y_L} \right\} > 0 \qquad (20)$$

Notice that the existence of the operating power gain depends only upon the device and its output termination.

By analogy with the operating power gain, the available power gain is defined whenever the output conductance is positive or,

$$G_{OUT} = \mathrm{Re}\left\{ y_o - \frac{y_r y_f}{y_i + Y_S} \right\} > 0 \qquad (21)$$

The existence of the available power gain depends only upon the device and its input termination.

The insertion power gain is a function of both $Y_S$ and $Y_L$ and is thus similar to the transducer power gain. The insertion power gain exists whenever the transducer power gain exists.

All of the power gains exist under all conditions of termination when the transistor or device is inherently stable. Inherent stability is a property of the transistor or other device (described by its small-signal parameters) only and does not depend upon the terminations $Y_S$ and $Y_L$. A device or network is inherently stable if for all passive terminations $Y_L$ and $Y_S$, $G_{IN}$ and $G_{OUT}$ are always positive. A network is inherently stable whenever

$$g_i > 0 \qquad (22)$$

$$g_o > 0 \qquad (23)$$

and

$$0 \leqslant C = \frac{|y_r y_f|}{2 g_i g_o - \mathrm{Re}(y_r y_f)} < 1 \qquad (24)$$

The parameter C is called the Linvill stability factor. Its reciprocal known as Rollett's stability factor is also used as an indicator of stability, particularly with S-parameters.

## Example

The following numerical example illustrates some of the points discussed above. Pure-real y-parameters are used to simplify the calculations. Consider,

$$y_r = 10 + j0 \qquad y_r = 1.5 + j0$$
$$\text{(mmho)}$$
$$y_f = 150 + j0 \qquad y_o = 10 + j0$$

Here and throughout the example, all admittances will be in mmho.

The first step in any design procedure is usually the evaluation of the Linvill stability factor. Using Eq. (24) we get,

$$C = \frac{|(1.5)(150)|}{2(10)(10) - (1.5)(150)} = -9.0 < 0$$

Thus, the device is potentially unstable and some of the power gains may not exist depending upon the in-

*(Continued on page 28.)*

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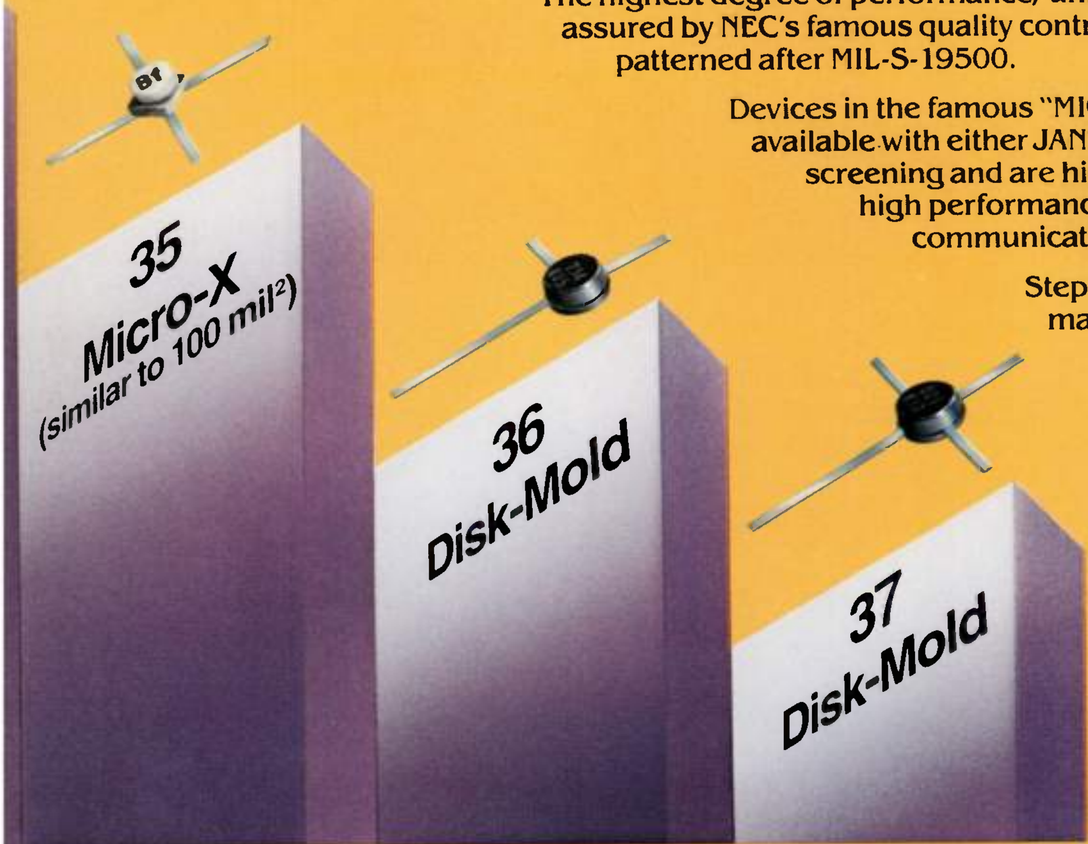


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3005 Democracy Way, Santa Clara, California 95050 • Tel. (408)988-3500 • Tlx. 346393 or 171197 put and output terminations. For this example, assume that the terminations are:

$$Y_S = 20 + j0 \qquad \text{(mmho)}$$
$$Y_L = 10 + j0$$

Since $g_i$ and $g_o$ are positive, Stern's stability factor may be used to examine the worst-case stability of the amplifier with the specified load and source terminations. Using Eq. (19) the result is:

$$k = \frac{2(10 + 20)(10 + 10)}{\mathrm{Re}((1.5)(150)) + |(1.5)(150)|} = 2.67 > 1$$

Hence, Sterns' stability factor shows that the circuit is stable for all values of $B_S$ and $B_L$ as long as $G_S$ and $G_L$ are maintained at the specified values. The Stern's factor tells nothing about the existence of the other power gains. A confirmation of the existence of the transducer gain and knowledge of the other gains is obtained by using Eqs. (11-13) to obtain the net input and output conductances for the specified terminations. These results are:

$$G_{IN} = \mathrm{Re}(Y_{IN}) = R_e \left\{ 10 - \frac{(1.5)(10)}{10 + 10} \right\}$$
$$= -1.25 < 0$$

$$G_S + G_{IN} = 20 - 1.25 = 18.75 > 0$$

$$G_{OUT} = \mathrm{Re}(Y_{OUT}) = \mathrm{Re} \left\{ 10 - \frac{(1.5)(10)}{10 + 20} \right\}$$
$$= 2.5 > 0$$

$$G_L + G_{OUT} = 10 + 2.5 = 12.5 > 0$$

Since $Y_{IN}$ has a negative real part, the operating power gain is not defined. However, the net input circuit conductance is positive and the output conductance is positive so that the transducer power gain exists. This confirms the result obtained from the Stern stability factor. Since the output conductance is positive, the available power gain has meaning. These latter two gains calculated from Eqs. (4) and (10), respectively. Such calculations are summarized below.

$$G_P = \text{not defined}$$

$$G_T = \frac{4|150|^2(20)(10)}{|(10 + 20)(10 + 10) - (1.5)(150)|^2}$$
$$= 128.0 = 21.07 \text{ dB}$$

$$G_A = \frac{|150|^2(20)}{|10 + 20|^2 \,\mathrm{Re} \left\{ 10 - \dfrac{(1.5)(150)}{10 + 20} \right\}}$$
$$= 200 = 23.01 \text{ dB}$$

Note that the transducer power gain is less than the available power gain. This is a consequence of the available power gain as the special case of the transducer power gain where conjugate match is enforced at the output port. In all cases, the transducer power gain will be less than or equal to the smaller of the two gains; operating and available. Only if an inherently stable transistor is operated in the condition of maximum gain, will all three gains be equal.

## Summary

Each of the several power gains has attributes that makes it useful in a particular design situation. The transducer power gain, which includes the effects of the input and output terminations is most useful in general design circumstances. Specialized design procedures such as low-noise design or stability studies may be best accomplished with another power gain. A good understanding of the meanings of the various power gains can help the rf design engineer produce a better design in less time with less laboratory testing. □


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INPUT ATTEN
10 dB ATTEN
LOG SCALE
10 dB/
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# Powers FETs for RF Amplifiers

## Part II: Design the Amplifier

New-generation RF power MOSFETs bring a host of advantages to high-power high-frequency circuit design. These benefits, outlined in the first article of this two-part series, include thermal stability, low noise generation, reduced feedback, and higher system reliability. Part two demonstrates with a practical example how these power FET characteristics can simplify circuit design.

*Gary Appel, Applications Engineer, and*
*Jim Gong, Product Manager*
*Siliconix, Inc.*
*Santa Clara, California*

### Amplifier Design

A simplified schematic for a broadband push-pull amplifier is shown in Figure 15. Small-signal transconductance for this amplifier is simply that transconductance existing at the chosen bias conditions. Symmetry requires that no signal currents flow through the output transformer, T1. Output voltage and power can then be expressed as

$$V_{out} = \frac{gm\, R_L\, V_{in}}{2} \tag{2}$$

$$P_{out} = V^2_{out}/R_L = \frac{gm^2\, R_L (V_{in})^2}{4} \tag{3}$$

Because transistors Q1 and Q2 represent current sources in series, the impressed voltage is the same as with a single device driving $R_L$. Input voltage is

$$V_{in} = \sqrt{2R_g\, P_{in}} \tag{4}$$

The power gain of the amplifier can now be expressed as

$$P_{out}/P_{in} = gm^2 R_L R_g/2 \tag{5}$$

As signal level increases, gate voltage moves along the transconductance vs $V_{GS}$ curve. Linear amplification is possible here due to the square-law behavior of the power FET. The decrease in transconductance as gate voltage decreases is offset by the increase in transconductance as gate voltage rises. Second-harmonic distortion created by this process is cancelled by the push-pull operation.

Eventually, the flattening out of the transconductance at high gate voltages is compensated for by the zero transconductance below the gate threshold voltage. This nonlinear transfer characteristic results in linear Class AB amplification because it represents a near optimum nonlinearity.

While one FET is cut off, RF current from the other device must be split equally between the load and T1. This implies that T1 is now acting as a 4:1 transformer, and the load impedance has become $R_L/4$. Transconductance, however, should have doubled by this time, thereby maintaining the original gain.

The output capacitances, originally in series due to the push-pull operation, are now in parallel because of the center-tapped transformer action of T1. In short, only the analysis changes; gain, bandwidth, and output power calculations are unchanged. Although T1 is a 4:1 transformer and must be designed appropriately, the 100V supply simplifies the task.

The broadband capability and linearity of the DVD150T RF power FET can be demonstrated by applying the device to an amplifier circuit design. The circuit used is a 2 to 30 MHz linear amplifier with the FETs operating in push-pull and under Class AB bias conditions. Required power output is 250W PEP.

For the circuit in Figure 15, required load resistance is

$$R_L = \frac{2\,(V_{DS} - V_{sat})^2}{P_{out}} \tag{6}$$

Using respective values for $V_{DS}$, $V_{sat}$ and $P_{out}$ of 100V, 20V, and 250W, this equation yields an $R_L$ of 51.2 ohms. Using a standard load resistance of 50 ohms, T1 becomes a 12.5 to 50-ohm transformer, and each output capacitance is shunted by 25 ohms of load resistance. Due to the RC time constant, the 3dB bandwidth of the output circuit is 63 MHz. The output design, consisting of a 50-ohm balun B2 and a 4:1 transformer, is complete.

If necessary, the output network could be improved by optimizing the series inductance (including the leakage inductance from B2), and providing a final shunt capacitor at the output to form a three-section low-pass filter. This refinement will not be considered here, however.

In order to get the same bandwidth at the input, which has four times as much capacitance as the output, the source resistance must be decreased by a factor of four. The required value for $R_g$ is 6.25 ohms, and the input network consists of a 4:1 transformer plus a 12.5-ohm balun and two 6.25-ohm resistors. Again, no additional attempt is made here to get an optimum low-pass filter for the required bandwidth.

Using the appropriate bias to get the recommended small-signal transconductance of 0.5 mhos, the corresponding power gain using Equation 5 is

$$\text{Gain} = \tfrac{1}{2}(0.5)^2(50)(6.25) = 39.1$$

or 15.9dB. The schematic for the complete amplifier is shown in Figure 16. The simplicity of this circuit is obvious. No feedback is employed, and the input has no frequency compensation. Bias is supplied through two 10k-ohm resistors, rather than from another power semiconductor bolted onto a heat sink. The impedance levels are reasonable, and the bypass and blocking capacitors are not critical.

Available output power for a given intermodulation ratio at 30 MHz was increased by increasing the value of the input termination resistors from the calculated 6.25 ohms to 12 ohms. This adjustment increases gain and improves the intermodulation performance while limiting the input VSWR to approximately 2:1. The 12-ohm resistors each consist of two 24-ohm 2W resistors connected in parallel.

With this adjustment, the new calculated gain becomes

$$\text{Gain} = \tfrac{1}{2}(.5)^2(12)(50) = 75$$

or 18.7dB. This figure is modified by a 0.5dB mismatch loss due to the 2:1 input VSWR. Thus predicted amplifier gain is 18.2dB.

## Test Results

Figure 17 shows the gain and input VSWR measured for this amplifier. As predicted, the input VSWR is slightly less than 2:1 over the entire operating range, while the gain is very close to the value of 18.2dB. (Prior to changing the value of $R_g$, a maximum VSWR of 1.2 was reached at a power gain of 15.5dB.)

The intermodulation distortion levels at 2 and 30 MHz are shown in Figure 18. The distortion at 2 MHz is entirely due to amplitude nonlinearity. Crossover distortion, causing the peak around 150 W, could be reduced by increasing the quiescent current somewhat. ($I_{DQ}$, chosen as 250 mA per transistor to compromise between distortion and no signal dissipation, is somewhat lower than the optimum value described earlier in terms of the quiescent transconductance value.) Phase distortion contributes to the intermodulation distortion at 30 MHz, resulting in a reduced output power level for a given intermodulation ratio.

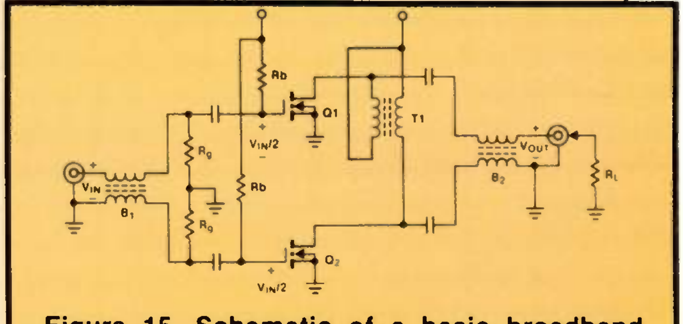

Figure 15. Schematic of a basic broadband, push-pull amplifier.

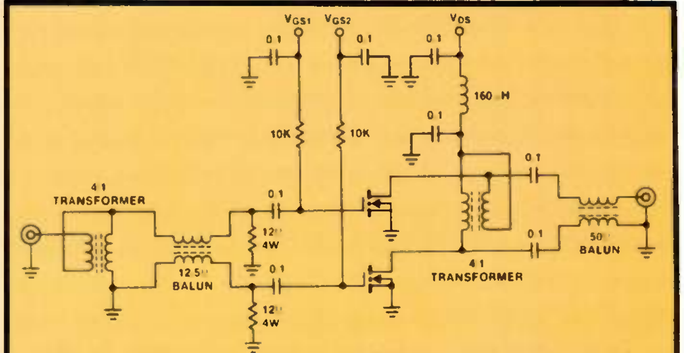

Figure 16. Complete 250W broadband, push-pull RF amplifier requiring neither feedback nor input frequency compensation.

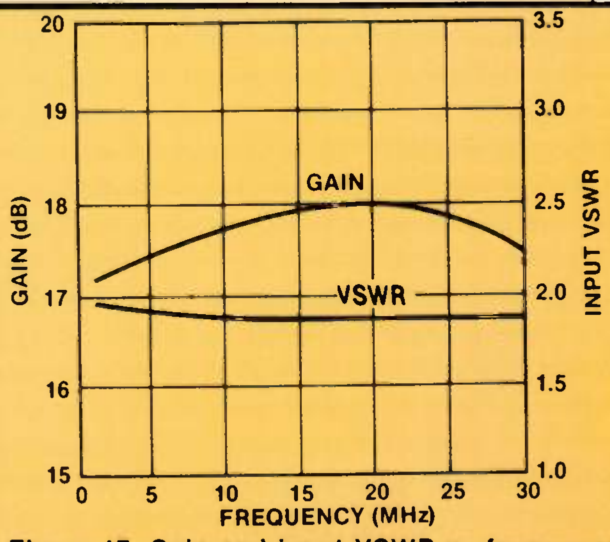

Figure 17. Gain and input VSWR vs frequency for the circuit in Figure 16.

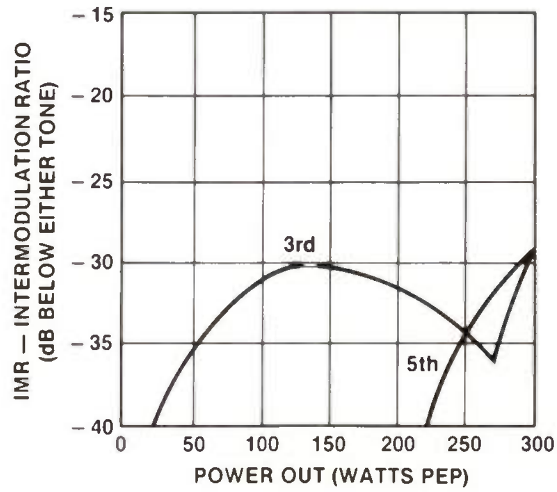


**Figure 18a. Intermodulation ratio vs. output power for the circuit in Figure 16 at 2 MHz.**

Finally, the amplifier has a noise figure of only 7.6dB, which is unattainable with a bipolar design and is quiet enough for use at the front end of a high-frequency communications receiver.

## Conclusion

The design procedures and test results for the 2-30 MHz linear amplifier described in this article illustrate the broadband capability and linearity available from present FET technology. Simple circuit design techniques yield excellent correlation between predicted and measured performance. Other application possibilities for exploiting the advantages of RF power MOSFETs include low-noise power amplifiers, distributed amplifiers, high-efficiency amplifiers, gate-modulated amplifiers and ultralinear amplifiers using feed-forward techniques. □

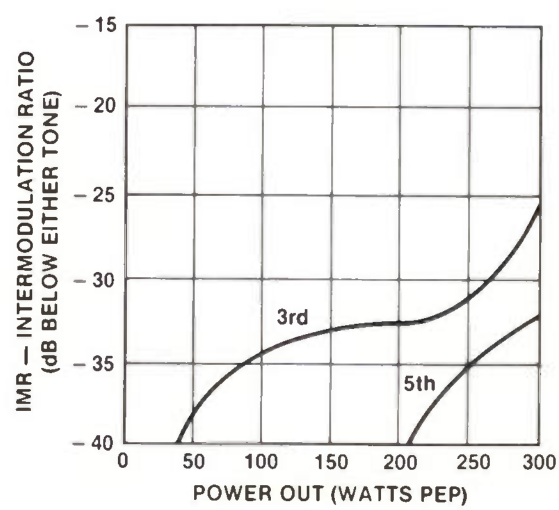


**Figure 18b. Intermodulation ratio vs. output power for the circuit in Figure 16 at 30 MHz.**

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2. Date of Filing: September 30, 1982.
3. Frequency of Issue: Bi-Monthly.
4. Location of Known Office of Publication: 6430 S. Yosemite St., Englewood, Colorado 80111.
5. Location of the Headquarters or General Business Offices of the Publishers: 6430 S. Yosemite St., Englewood, Colorado 80111.
6. Names and Addresses of Publisher and Editor. Publisher: Bill W. Childs, 6430 S. Yosemite St., Englewood, Colorado 80111. Editor: Larry Brewster, 6430 S. Yosemite St., Englewood, Colorado 80111.
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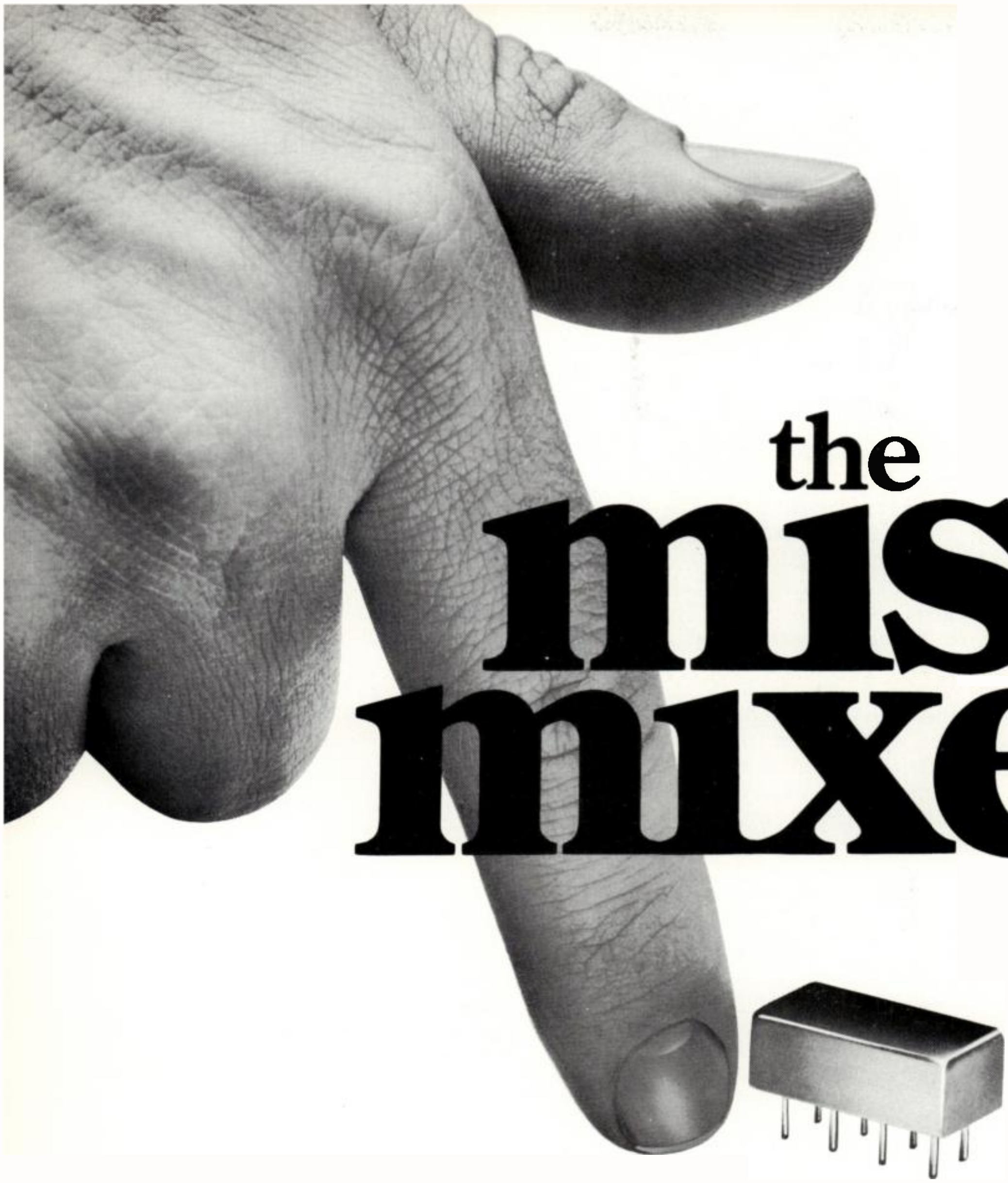

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| IF | | DC-500 | | 0.5-500 | |
| *CONVERSION LOSS, dB* | | *Typ.* | *Max.* | *Typ.* | *Max.* |
| One Octave from Band Edge | | 5.5 | 7.5 | 6.0 | 7.0 |
| Total Range | | 6.5 | 8.5 | 7.0 | 9.0 |
| *ISOLATION, dB* | | *Typ.* | *Min.* | *Typ.* | *Min.* |
| Lower Band Edge to | LO-RF | 50 | 40 | 50 | 35 |
| One Decade Higher | LO-IF | 45 | 35 | 45 | 30 |
| Mid Range | LO-RF | 45 | 30 | 40 | 25 |
| | LO-IF | 40 | 25 | 35 | 20 |
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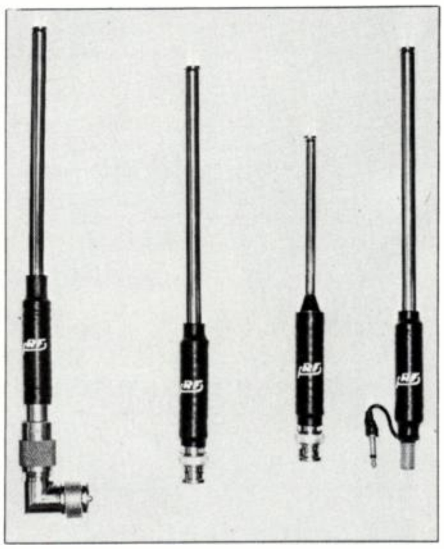

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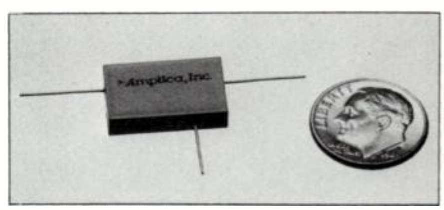

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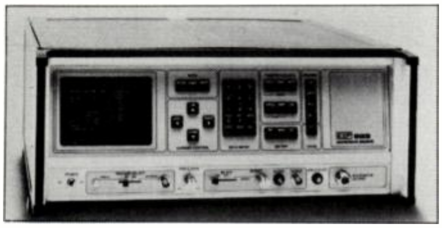
### Microwave Sweep Oscillators

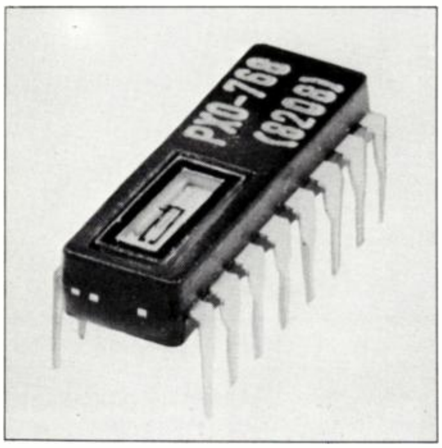
EIP Microwave, Inc. announces a



new line of six microwave sweep oscillators covering the 1 to 18.6 GHz range which are equipped with interactive CRT display. **EIP Microwave, Inc., 2731 North First Street, San Jose, CA 95134, (408) 946-5700, or circle INFO/CARD #136.**

## Programmable Crystal Oscillator

A crystal oscillator that can be programmed to produce 57 discrete frequencies has been introduced by Statek Corp. Output frequencies range from 0.002 Hz to 1.25 MHz and possess the high accuracy and stability of the crystal oscillator base frequency. The PXO Series of programmable crystal




oscillators is packaged in a standard 16-pin DIP that contains both the crystal and a CMOS IC. The device is TTL compatible. Price is less than $10 each in 1,000-pc. qty. Delivery is stock to 8 weeks. **Statek Corp., 512 N. Main, Orange CA 92668, (714) 639-7810 or INFO/CARD #135.**

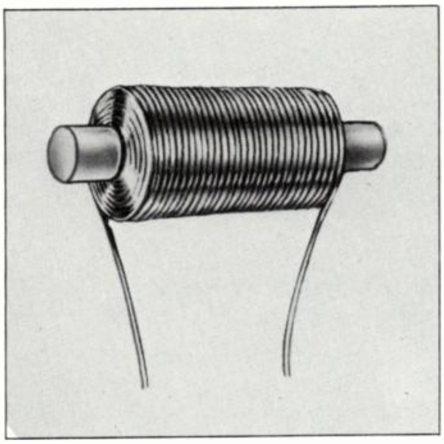
## Bi-Phase Modulator

The Anzac Division of Adams-Russell has introduced a bi-phase modulator covering the 10-750 MHz frequency range. The PM-102 is a TO-8 device that provides 0/180° phase modulation with phase deviation of typically less than 1°. The PM-102 has an insertion loss of less than 3 dB, VSWR of 1.25:1 and can handle RF inputs of + 17 dBm max. Carrier suppression is typically 35 dB. The unit includes an ECL Series 10000 driver for easy system interfacing. The PM-102 is designed to meet and is screen-


able to MIL-STD-883 and all specifications are guaranteed over the – 55 to + 85° C range. The PM-102 delivery is stock at $125 each in small quantity. **Adams Russell, Anzac Division, 80 Cambridge St., Burlington, MA 01803, (617) 273-3333 or INFO/CARD #134.**

## Microminiature Rod Core Inductors

A series of self-leaded, rod core coils suitable for high volume, high-reliability applications is available from PICONICS, INC. of Tyngsboro,



MA Piconics JA Series provide inductance values from .045uH to 1MH, operating frequencies from .25MHz to 50MHz, and self resonating frequencies from 1.1 to 1200 MHz. Piconics JA Series coils are priced from $.70 for quantities of 100-299, to $.26 for 100,000 pcs. and up. **Piconics, Inc., Irving Kadesh, VP, Sales/Marketing, 26 Cummings Road, Tyngsboro, MA 01879, (617) 649-7501, or please circle INFO/CARD #133.**


## TO-8 Packaged Amplifier

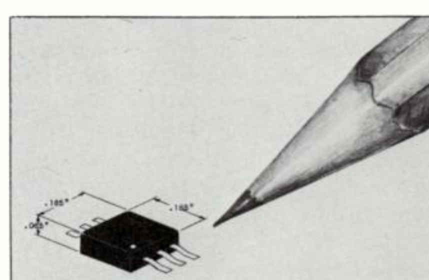
New to Aydin Vector's line of hybrid RF amplifiers, Model MHT-255 operates from 5 to 500 MHz with a noise figure of 2.2 dB. Amplifier gain is 14 dB with +10 dBm power out and +22 dBm third order intercept. Maximum input/output VSWR is 2.0:1 at 50 ohms. Operating voltage is +15 V at 26 ma. Operating temperature range is −54° to 100°C. **Aydin Vector, P.O. Box 328, Newtown, PA 18940. (215) 968-4271, or circle INFO/CARD #132.**

## 500 MHz A/D Converter

The Microelectronics Division of Alpha Industries, Inc. model A-232 thick film hybrid A/D Converter is an ultra fast 3 bit Analog to Digital Converter subset. The encoder is a Flash (parallel) Quantizer with 2 binary encoders providing 3 bit binary data. Each unit has an overflow bit. Encoding overflow bits enable cascading units for higher resolution A/D's. The A-232 is manufactured in a hermetically sealed 24 pin double dip package. **Alpha Industries, Microelectronics Division, 3015 Advance Lane, Colmar PA 18915, (215) 822-1311, or circle INFO/CARD #131.**

## Wideband Transformers

Operational in the 1-300 MHz range with typical insertion loss of less than .5dB characterize the low profile molded ultra miniature wideband tranformer family. Designed to meet the most stringent military and space requirements, the LP series is capable of operating over a temperature range



of −55°C to 125°C. Average price (qty. 1000) $5.00 each, delivery eight weeks ARO. **Vanguard Electronics Company, 1480 West 178th Street, Gardena, California 90248, (213) 323-4100, or circle INFO/CARD #130.**

## Crystal Oscillator

Techtrol introduces model XO-270 5" high crystal oscillator available at any frequency in the 100 MHz – 1 GHz range. Features include compact 2.0 × 2.0 × .5" size, ± 10 PPM stability −20 to +60°C, +10dBm output into 50 ohms (ECL compatible), low power consumption 50 ma @ + 15VDC (F0 = 1 GHz), low phase noise – 135 dBc/Hz @ 1 KHz removed @ FO = 100 MHz, PC board mounting. **Techtrol Cyclonetics, Inc., 815 Market Street, New Cumberland, Penn. 17070, (717) 774-2746, or INFO/CARD #129.**

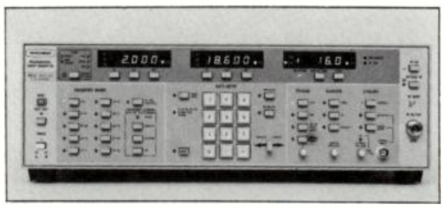
## RF Fuse Holders

Coaxial package for subminiature fuse. DC-1000 MHz. Delivery Stock-4 weeks, 50RF-010—N Male-N Female $30.00, 50RF-011—BNC Male-BNC Female $28.00, 5ORF-012—BNC Female-BNC Female Panel Mount $60.00. **JFW Industries, 2719 E. Troy Ave.,**

**Indianapolis, IN 46203, (317) 783-9875, or INFO/CARD #128.**

## Programmable Sweep Generator

Leveled-power output of 40 mW from



2 to 18.6 GHz is available from the new WILTRON 6637 A-40 Programmable Sweep Generator. The instrument is especially useful to electronic warfare and countermeasures engineers who, until now, have had to use amplifiers and elaborate switching arrangements to make high-power measurements over a broad frequency range. Now tests can be made at typically greater than 50 mW over much of the 2 to 18.6 GHz range, either under computer or front-panel control. Price: Model 6637A-40, $28,750, Option 2, 70 dB Attenuator, $1,500. Delivery: 90 days. **Contact Walt Baxter, Wiltron Company, 805 East Middlefield Road, P.O. Box 7290, Mountain View, CA 94042-7290, Tel: (415) 969-6500 or circle INFO/CARD #125.**

## Digital LCR Meter

Leader Instruments Corporation of Hauppauge, New York, has announced the availability of a new automatic ranging digital bridge meter, the LCR-745. The CPU-controlled LCR meter provides direct resistance, capacitance, and inductance measurements of components and equivalent series or parallel circuits. The unit's wide automatic measurement ranges, from 0.001 ohm to 19.99 Mohm for resist-

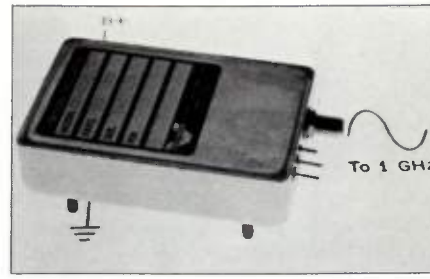


INFO/CARD 26

ance, 0.1 uH to 199.9 H for inductance, and 0.1 pF to 1999 uF for capacitance, greatly reduce the operation time associated with manual LCR instruments. **Leader Instruments Corporation, 380 Oser Avenue, Hauppauge, NY 11788, (516) 231-6900, or INFO/CARD #127.**

## Crystal Oscillator to 1 GHz

Model CO-283W Crystal Oscillator provides a stable output at any fixed frequency in the 500-1000 MHz frequency range at a level exceeding 0.5 vrms into 50 ohms (+7 dBm), with +13 dBm output optional. It provides stability better than ±25 ppm over 0°C to 70°C. Stability options include ±50 ppm over −55°C to +85°C and improved stability of ±3 ppm over 0-50°C. While the oscillator is factory




set to within 10 ppm of the specified frequency, a frequency adjustment for setting to within 1 ppm is optionally available, as is electronic tuning (VCXO) for phase locking. **Vectron Laboratories, Inc., 166 Glover Ave., Norwalk, CT 06850, (203) 853-4433 or INFO/CARD #126.**



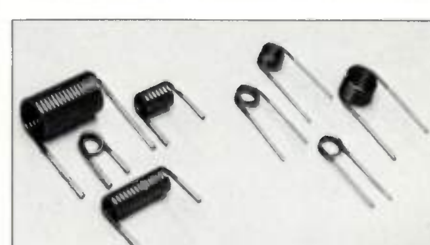
## Microwave Tuning Elements

Millimeter Wave Tuning Elements are now available from Johanson Manufacturing Corporation for operating frequencies of up to 100 GHz. Included are both metallic and dielectric tuning element types; efficient and cost effective for applications requiring precision, low loss, high resolution tuning. Designed for use in combline and interdigital filters, attenuators, Impatt and Gunn oscillators, impedance transformers, and waveguide circuitry. Prices are from $2.50

to $6.00 each in 1,000 piece order quantities. Delivery: 6-10 weeks. **Johanson Manufacturing Corp., Rockaway Valley Rd., Booton, N.J. 07005, (201) 334-2676 or INFO/CARD #105.**

## Air Core Inductors

Dale Electronics, Inc. has announced the expansion of its magnetic components production to include air core inductors. Dale is producing these low-cost, high-volume inductors for a wide variety of electronic applications,



including television, radio, statellite communication systems, microwave and other high-frequency uses. Dale's new air core inductor line includes models with 2 to 32 turns #18 to #32 guage wire, fabricated with variable pitch clockwise and counter-clockwise windings. Typical pricing for volume quantities of the new inductors range from \$.04 to \$.15, depending on exact specifications and requirements. Estimated delivery time is 4 to 6 weeks. **Dale Electronics, Inc., East Highway 50, Yankton, South Dakota 57078, (605) 665-9301, or INFO/CARD #104.**

## Rotary Joint: DC to 2.5 GHz

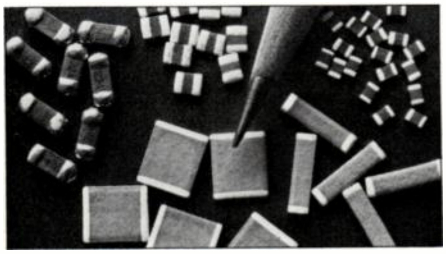
Model 30-2050 inline style rotary joint operates over the frequency range of DC to 2500 MHz. Maximum VSWR is 1.10 max. to 800 MHz, 1.25 max. to 2500 MHz. Insertion loss is 0.1 dB maximum, WOW .1 dB max/360° rotation. Power handling capability is 1 megawatt peak, and 10 Kw @ 200 MHz, 4.5 Kw @ 1500 MHz. Coaxial connectors—EIA, 1-5/8" with bullet and bead. **Diamond Antenna & Microwave Corp., 35 River Street, Winchester, MA. 01890. (617) 729-5500. TWX 710-348-1066 or INFO/CARD #100.**

## 100W, 40 dB Attenuator

KDI Pyrofilm has introduced a new 100 watt, 40 dB attenuator for Stripline and Microstrip Applications. The PPA-100D-40 is a dual 20 dB attenuator on a single mounting flange measuring only 1.90 inches by 1.040 inches. The VSWR is 1.25:1 from DC to 750 MHz. The PPA-100D-40 is rated at 100 watts maximum at a heat sink temperature of 85°C. The attenuation is 40 dB ± 2 dB. **KDI Pyrofilm Corporation, 60 South Jefferson Road, Whippany, N.J., 07981, (201) 887-8100 or INFO/CARD #99.**

## Ceramic Chip Capacitors

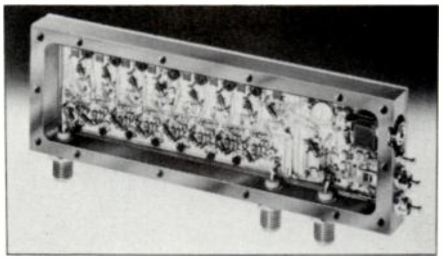
An expanded product line of monolithic ceramic chip capacitors is offered by San Fernando Electric Division of SFE Technologies. NPO/COG and X7R chips are available in 25, 50, 100, 200 and 500 V versions; Z5U chips are provided in 25, 50 and 100 V versions. The capacitance spectrum has been increased to a range of 1 pF



to 3.30 uF. Chip terminations are available with either palladium silver, Sn 62 solder, or SFE's exclusive "NIBAR" nickel barrier process. **San Fernando Elec. Div., SFE Technologies, 1501 First St., San Fernando, CA 91341 or INFO/CARD #103.**

## Log Amplifiers

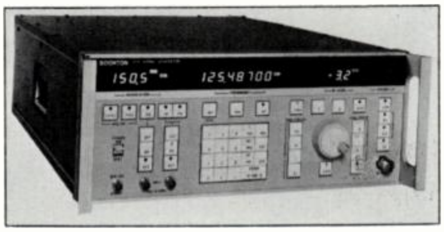
A new series of ultrabroadband log amplifiers covering the 100 to 500 MHz, 200 to 600 MHz and 500 to 1000 MHz frequency ranges is available from RHG Electronics Laboratory, Inc. Designated the ICLW series, these new multioctave log amps feature: rise times as low as 3 nsec, dynamic ranges to 80 dB, built-in voltage regulators, temperature-compensation and direct-coupled video. **RHG Electronics Laboratory, Inc., 161 E. Industry Court,**



**Deer Park, NY 11729. (516) 242-1100. INFO/CARD #102.**

### Signal Generator

Boonton Electronics announces new versions of its programmable



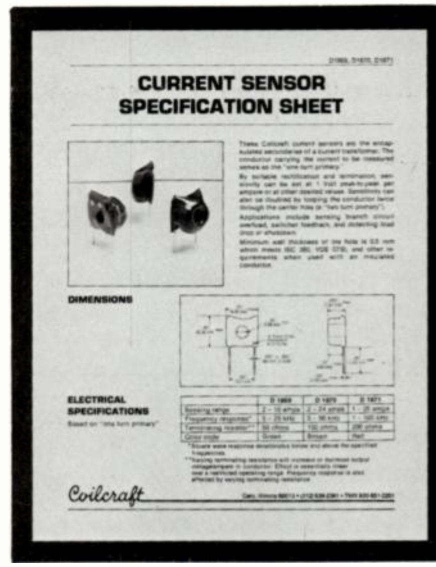
signal generators designed to test the wideband i.f. amplifiers and demodulators of satellite communication links. Designated the 1020/1021-S/1, these generators provide low-distortion f.m. to 600 kHz peak deviation at internal modulation rates to 50 kHz. External f.m. sensitivity is 1 dB down at 100 kHz and 3 dB down at 200 kHz. Phase modulation capability is to 6 radians peak, and a 75-ohm r.f. output to 200 MHz is available, calibrated in both dBm and voltage. **Boonton Electronics, P.O. Box 122, Parsippany, NJ 07054, (201) 887-5110, or INFO/CARD #101.**



## New Literature

### Sensor Data Sheet

Coilcraft has published a two page data sheet on its new line of low cost current sensors designed for switch-




ing power supply applications. Three models are described, with sensing ranges up to 10, 24, or 35 amps. Frequency response extends to 25, 50, and 100 kHz respectively. **Contact Mary Nendze, Coilcraft, Cary, IL 60013, (312) 639-2361, or INFO/CARD #124.**

### Amplifier Handbook

The LOCUS Amplifier Handbook is now available, and includes tables and graphs to aid design engineers, as well as listing the technical specifications of LOCUS amplifiers.

LOCUS produces an ever increasing number of high performance RF amplifiers which cover frequency ranges from under 100 kHz to over 12 GHz. **Locus, Inc. P.O. Box 740, State College, PA 16801, (814) 466-6275, or INFO/CARD #123.**

### Signal Processing Catalog

Tele-Tech Corporation announces a new catalog of signal processing components. The 32 page catalog features complete product specifications, mixer application notes and a specification guide. Mixers, RF switches, transformers, frequency doublers, hybrids and couplers are introduced. In addition, a description of the custom amplifiers, oscillators, VCO's and phase-locked loop assemblies are described. **Tele Tech,**

*(Continued on page 45.)*

P.O. Box 1827, Bozeman, MT 59715, (406) 586-0291, INFO CARD #122.

## Ferrite Selection Brochure

A 20-page brochure from Siemens Components Group covering the company's ferrite products and ferrite selection guides. The brochure provides full criteria for selection of ferrite materials and core types as well as required hardware for configuration of ferrites for specialized applications. It also contains general technical data and applications information. **Siemens Corporation, Box 1000, Iselin, New Jersey 08830 or INFO/CARD #121.**

## RF Filter Catalog

Catalog RF/82 is 28 pages of diplexers, bandpass, band reject and low pass/high pass filters for use in

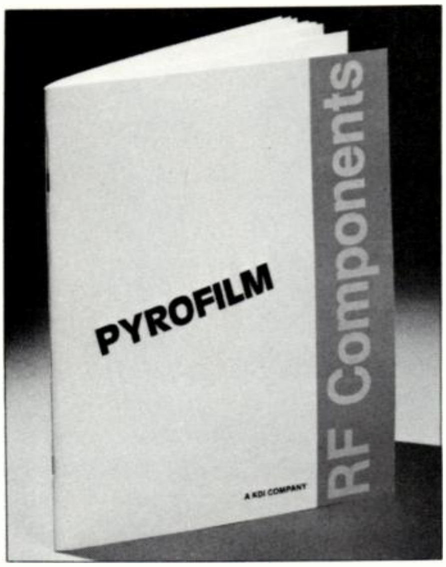

communications and other equipment operating 5-1000 MHz. **Microwave Filter Co., Inc., 6743 Kinne St., East Syracuse, NY 13057, 1-800-448-1666 (toll-free) or INFO/CARD #120.**

## Test Equipment Catalog

MobCat-82 is a new 20-page catalog of Mobile Communications Support Test Equipment by RF power instrument manufacturer Bird Electronic Corporation. **Bird Electronic Corporation, Cleveland (Solon) Ohio 44139, (216) 248-1200 or INFO/CARD #118.**

## RF Components Catalog

A 40 page catalog details KDI Pyrofilm's line of RF and Microwave Resistive Components for military and




commercial applications in the frequency range of DC to 18 GHz. **KDI Pyrofilm Corporation, 60 South Jefferson Road, Whippany, New Jersey, 07981, (201) 887-8100 or please circle INFO/CARD #117.**

### Coaxial Assemblies Catalog

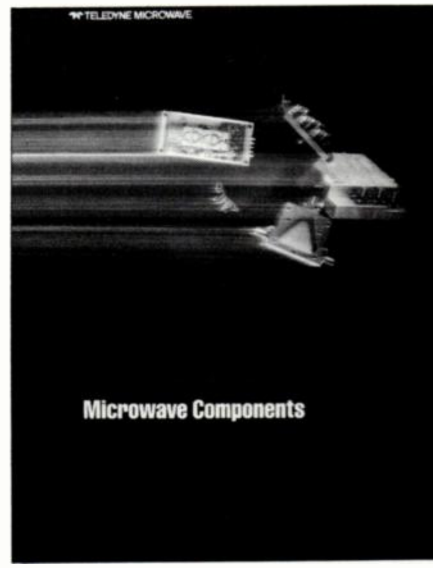
A 20 page catalog features a full line of coaxial cable, coaxial adapters, coaxial connectors, coaxial terminations and coaxial cable assemblies. Pricing on over 1,000 standard catalog items as well as technical specifications are included. **Pasternack Enterprises, 22017 Bushard St., Huntington Beach, CA 92646, (714) 962-9306, or INFO/CARD #114.**

### Pin Diode Design Guide

A new application note titled "Pin Diode RF Switch Design" is available from Frequency Sources Semiconductor Division. The literature gives a clear description of the device and its equivalent r.f. circuit. It describes all important operating parameters at HF, VHF, and UHF in the on and off states and gives tips on low frequency operation and switching speed. Design equations and performance curves are provided for series, shunt, and tee switches along with driver circuits and a designer's selection guide. **Frequency Sources, Semiconductor Division, 16 Maple Road, Chelmsford, MA 01824, (617) 256-8101 or please circle INFO/CARD #115.**

### Microwave Components Catalog

Teledyne Microwave announced that its seventh annual Microwave

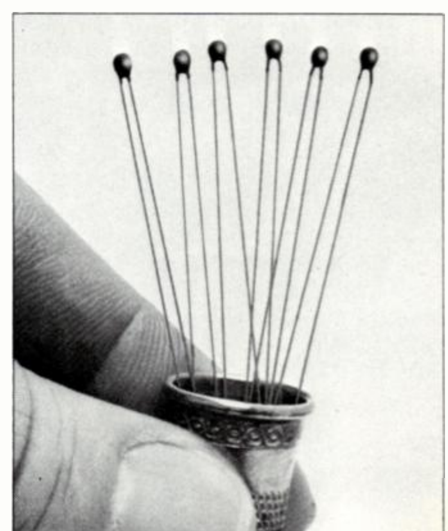



Components Catalog is now available. The catalog includes comprehensive information about Teledyne's film hybrid MIC design and production facility and their complete line of coaxial switches, circulators/isolators, filters/multiplexers, and special products. It also includes a multitude of application notes and ordering information. **Teledyne Microwave, 1290 Terra Bella Avenue, Mountain View, California 94043 (415) 968-2211 or INFO/CARD #116.**



### Thermistor Catalog

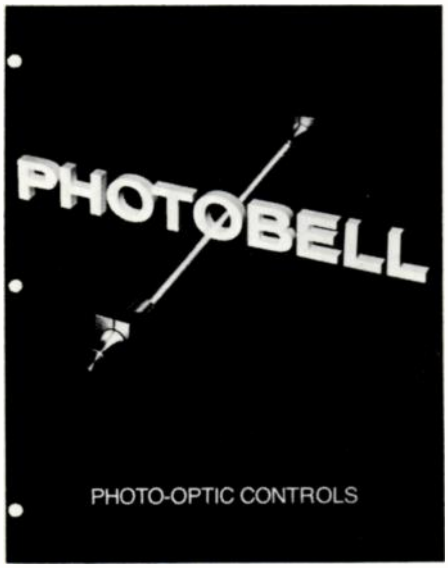
Designers planning to use negative temperature coefficient thermistors will find a wealth of useful information condensed in a new "pocket-size" catalog available from Dale Electronics. The "catalog" details the products of the Dale Western Thermistor Division. Specifications are given for most frequently specified styles, many of which are available for applications where interchangeability is required. **Dale Electronics, Inc., Dept. 860, 2064**



**12th Avenue, Columbus, Nebraska 68601, (402) 371-0080 or please circle INFO/CARD #111.**

## Photo-Optic Controls Catalog

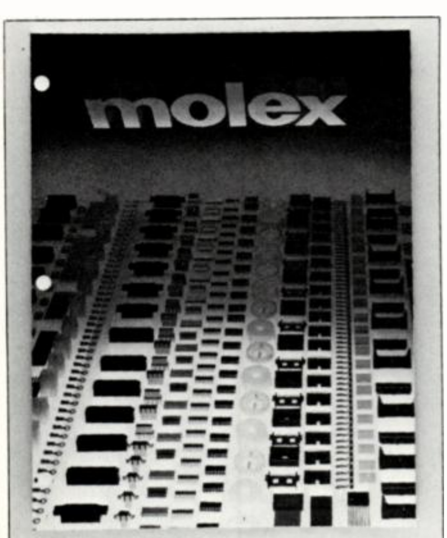
Twelve page catalog contains guidelines for making selection of photo-optic controls and describes comprehensive product line. Termi-




nology and applications for single unit types and 2-unit projector/receiver types are described. Special application systems, integral controllers, external controllers and housing dimensions are also covered. **Photobell Company Inc., 26 Just Road, Fairfield, NJ 07006, (201) 227-3613 or INFO/CARD #110.**

## Molex Catalog

Molex is pleased to announce the publication of a 336-page full product line catalog. Included are photos, 3-dimensional drawings, technical data




and ordering information for 22,000 Molex electrical and electronic components and related tooling. **Molex Inc., 2222 Wellington Ct., Lisle, IL 60532, or INFO/CARD #109.**

## RF Components And Subsystems Catalog

Availability of its recently published 80-page catalog on RF Components and Subsystems for signal processing was announced by Olektron Corporation. The catalog contains detailed technical data and application notes on the company's line of standard components—amplifiers, attenuators, baluns, beamformers, comparators, couplers, hybrids, mixers, modulators, phase shifters, power dividers, power sensors, switches, time delays and transformers. **Olektron Corporation, 61 Sutton Road, Webster, MA 01570. (617) 943-7440 or INFO/CARD #113.**

## LeCroy Shortform Catalog

A new 52-page shortform catalog is now available from LeCroy Research Systems Corporation. The catalog provides descriptions of the integrated product lines available from all five LeCroy Divisions: High Energy Physics (HEP), Acquisition and Control Systems Division (ACS), Fiberoptic (FSD), California (CA), and European Products (EPD). **LeCroy Research Systems, 700 South Main Street, Spring Valley, NY 10977, (914) 425-2000, or please circle INFO/CARD #112.**

## Microwave Components Catalog

A new 44-page Microwave Component Catalog has been published by EMC Technology, Inc., Cherry Hill, New Jersey. The new catalog contains full descriptions of the company's line of microwave connectors, attenuators, terminations, resistors and DC blocks. **EMC Technology, Inc., 1971 Old Cuthbert Rd., Cherry Hill, NJ 08034, (609) 429-7800. Please circle INFO/CARD #106.**

## Transformer Design Manual

A 32-page transformer design manual for use by the telecommunications industry is available from James Electronics, Inc. Complete with mechanical and electrical specifications, the telecommunication transformer and power supply manual covers coupling transformers, signal guard isolation transformers, power transformers (2.4 watt to 12 watt), Electro-Guard shielded power transformers for use in telecommunications and computers and its plug-in power supplies "Plug-Pack" and Series 42 "Wallpack" for use in in-office telecommunications and telephone equipment. **James Electronics, Inc., Marketing Dept., 4050 North Rockwell Street, Chicago, IL 60618, or INFO/CARD #108.**

## Variable Capacitor Catalog

Mepco/Electra's 10-page Variable Capacitor Catalog provides detailed specifications for ceramic and film dielectric capacitors. Product description for each series includes salient design features, device photograph, electrical and mechanical specifications, dimensional outline drawing and mounting dimensions. Series 2800D offers low temperature coefficient, low losses, high DC working voltage, low torque operation and very small size. Series 2500A are available in formats designed specifically for direct installation in P.C. boards or direct solder-in. **Mepco/Electra, Inc., Columbia Road, Morristown, New Jersey 07960, or INFO/CARD #107.** ☐

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All information required to design LP antennas without other references. $29.50 to: Smith Electronics, Dept. R 8200 Snowville Rd., Cleveland, OH 44141

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(317) 846-1201

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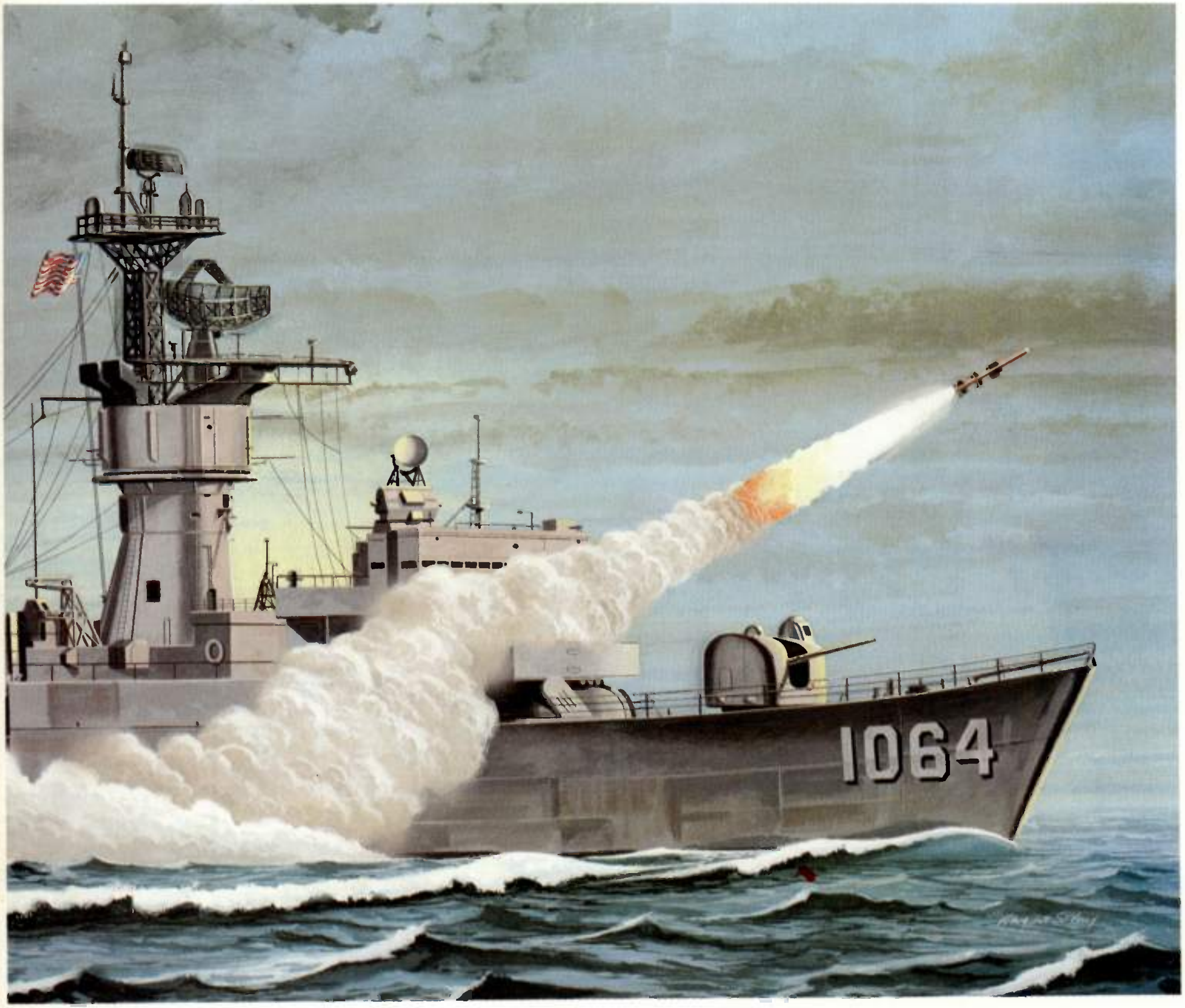
# On Target!

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ANZAC DIVISION

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