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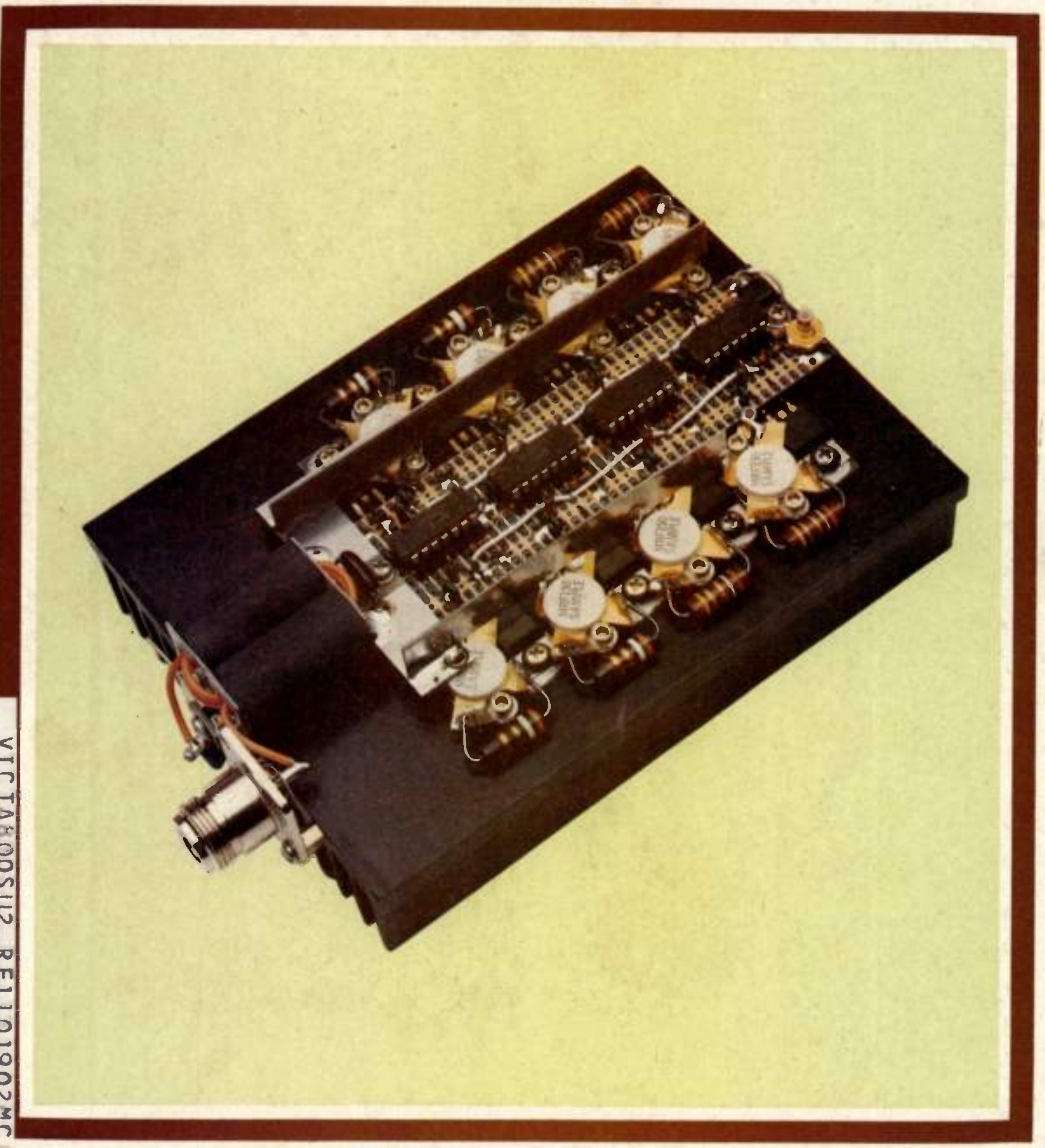

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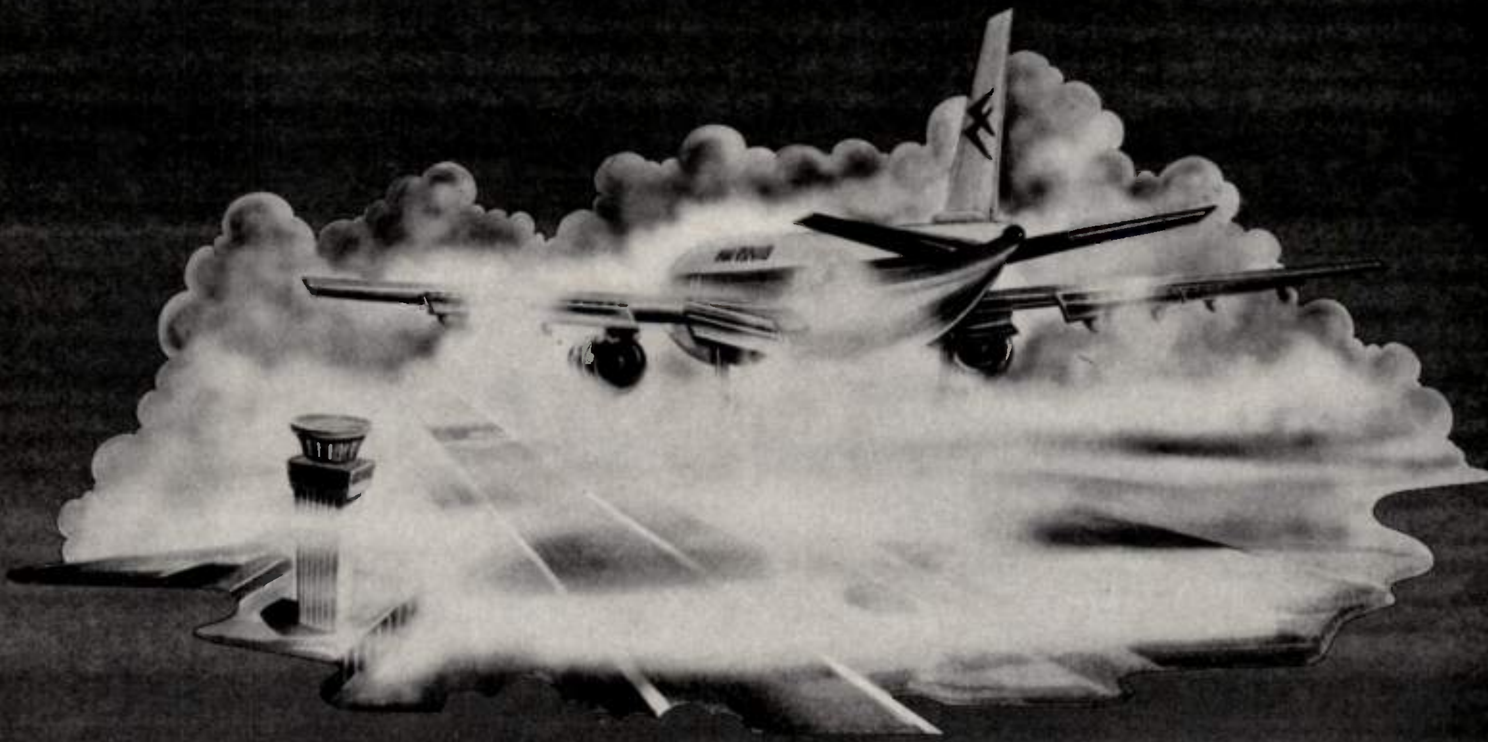

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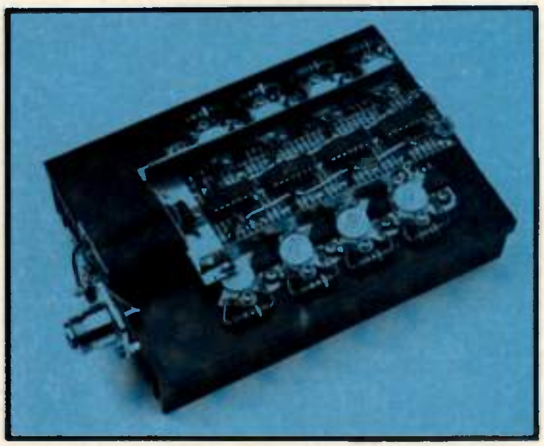
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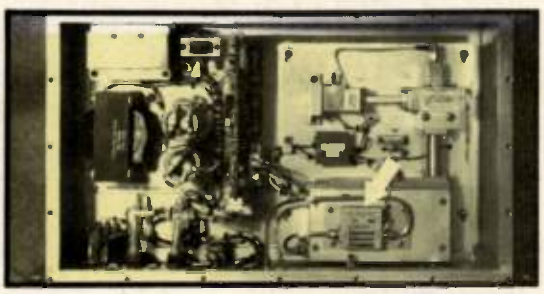
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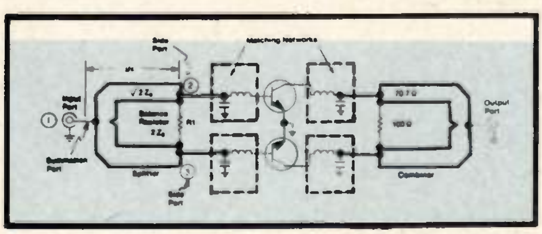
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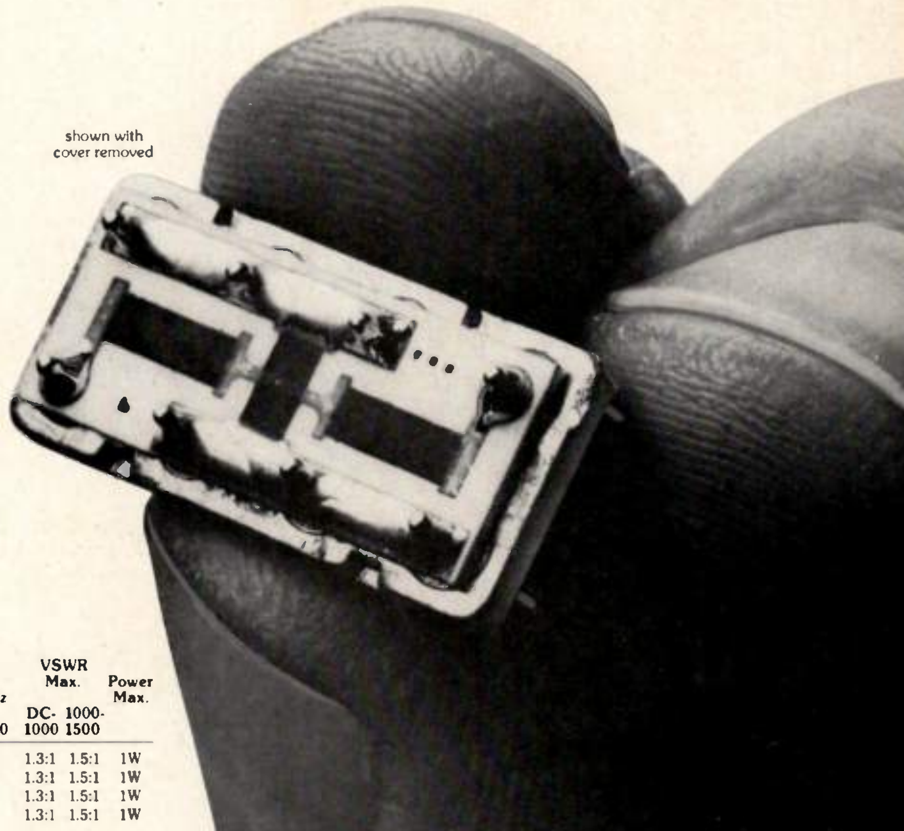

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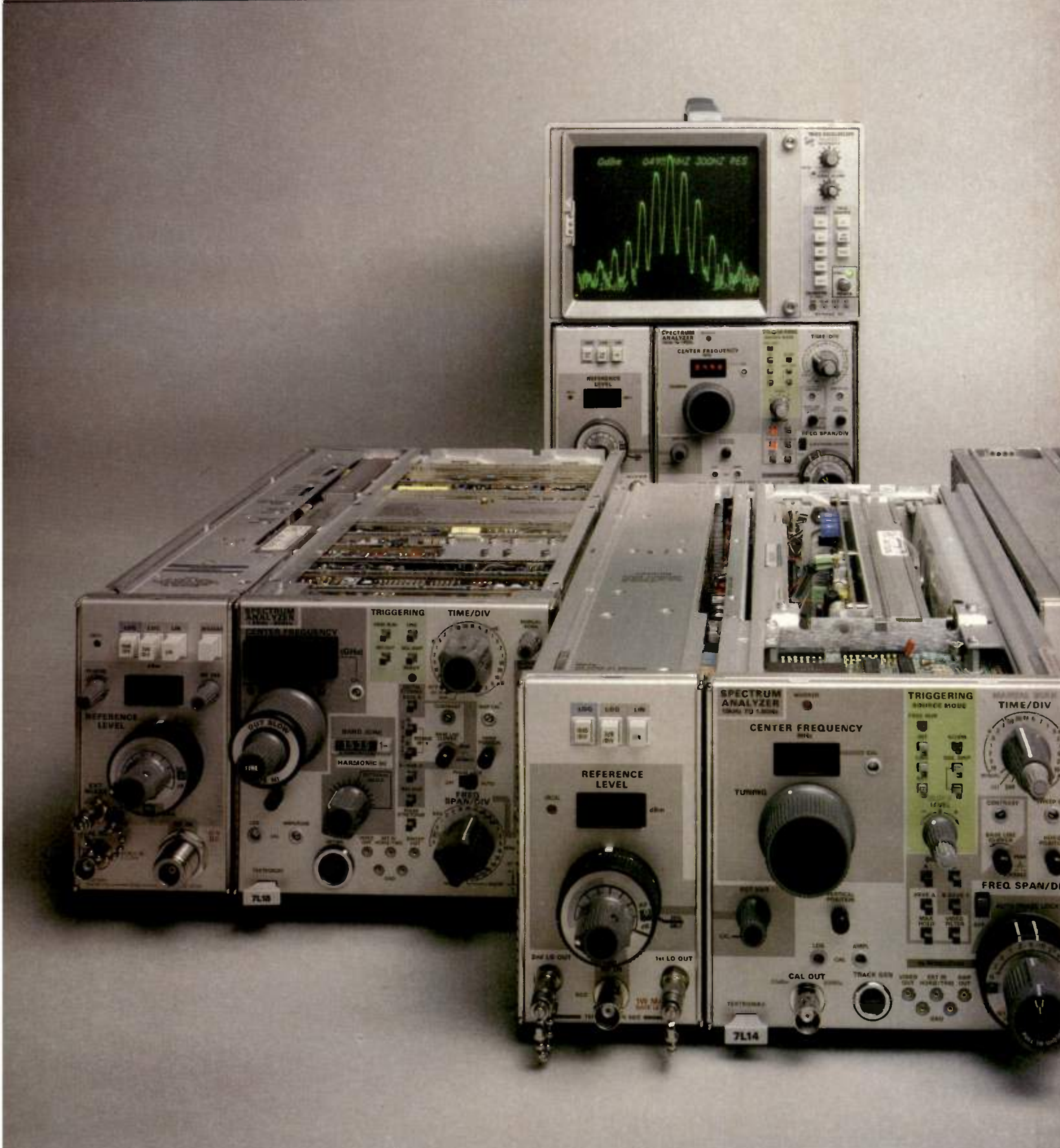

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Power MOS FETs *Versus* Bipolar Transistors

What is better, if anything, with the power FET's if we can get a bipolar transistor with an equal power rating for less than half the price?

By Helge O. Granberg
Sr. Staff Engineer
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Several manufacturers have recently introduced power FET's for RF amplifier applications. Devices with 100 W output capabilities are available for VHF frequencies and smaller units are made for UHF operation. All are enhancement mode devices, which means that the gate must be biased with positive voltage (N channel) in respect to the source to "turn it on." Early designs were so called V-MOS FET s, where the channel is in a V-groove. The V-groove must be etched with a special process, and the silicon material must have a different crystal orientation from the material normally used for bipolar transistors. The difficulty of the etching process in production has led to the development of other types of channel structures such as HEX and T, which are still vertical channel structures, but V-groove is eliminated, and the gate is on a straight surface. Thus,

Table A.

	Bipolar	TMOS FET
Zin, RS/ XS(30 MHz):	0.65 – J0.35 Ohms	2.20 – J2.80 Ohms
Zin, RS/ XS(150 MHz):	0.40+J1.50 Ohms	0.65 – J0.35 Ohms
Zo1 (Load Impedance):	Almost equal in each case, depending on power level and supply voltage.	
Biasing:	Not required, except for linear operation, high current voltage source necessary.	Some gate bias always required. Low current source, such as resistor divider sufficient.
Ruggedness:	Fails usually under current conditions. Thermal runaway and secondary breakdown possible.	Failure modes: Gate punch through, exceeding of breakdown voltages, over dissipation.
Linearity:	Low order distortion depending on die size and geometry. High order IMD a function of type and value of ballast resistors.	Low order distortion worse than bipolar for a given die size and geometry. High order IMD better due to lack of ballast resistors.
Advantages:	Wafer processing easier. Low collector-emitter saturation voltage, which makes devices for low voltage operation possible.	Input impedance more constant under varying drive level. Lower high order IMD. Easier to broadband. Devices or die can be paralleled. High voltage devices easy to implement.
Disadvantages:	Low input impedance with high reactive component. Internal matching required to lower Q. Input impedance varier with drive level. Devices or die can not easily be paralled.	Larger die required for comparable power level. Nonrecoverable gate breakdown. High drain – source saturation voltage, which makes low voltage, high power devices less feasible.

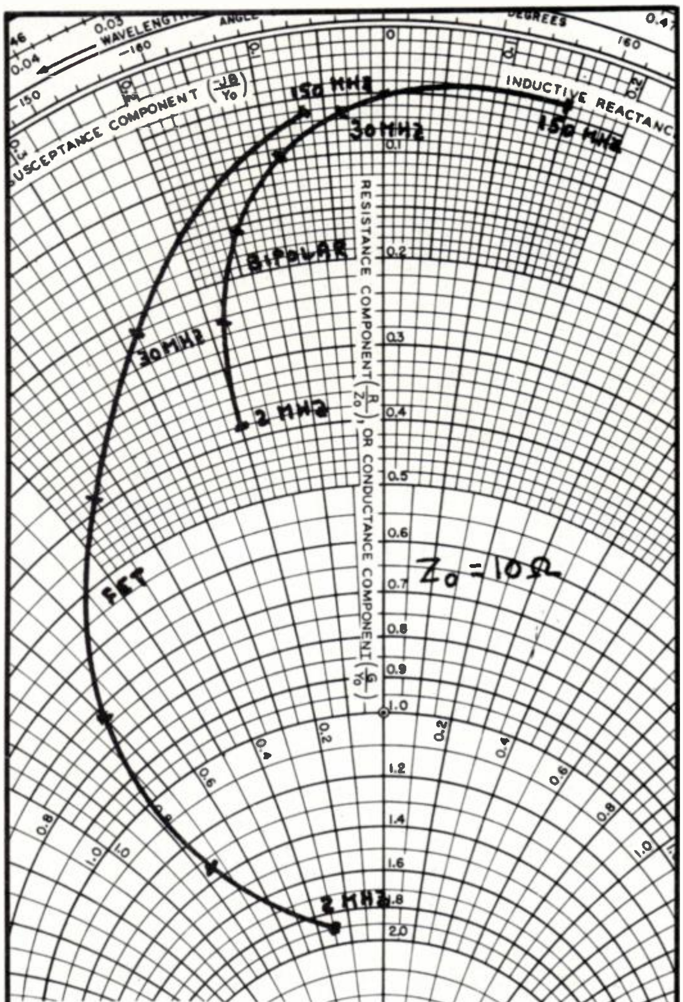


Figure 1. 150 watt MOS FET and bipolar comparison.

for an equal gate periphery, more room on the surface is required. Japanese manufacturers seem to favor geometries with horizontal channels. They are similar to small signal MOS FET's with a number of them paralleled on one chip. This technique represents even more wasteful use of the die surface than HEX or T MOS. Typically a power FET requires 50 to 100 percent more die area than a bipolar transistor for equal power output performance. For T MOS the number is about 50 percent. This is mainly due to the higher saturation voltage, but the geometry also gives some 30 percent less gate periphery than available base area in bipolar.

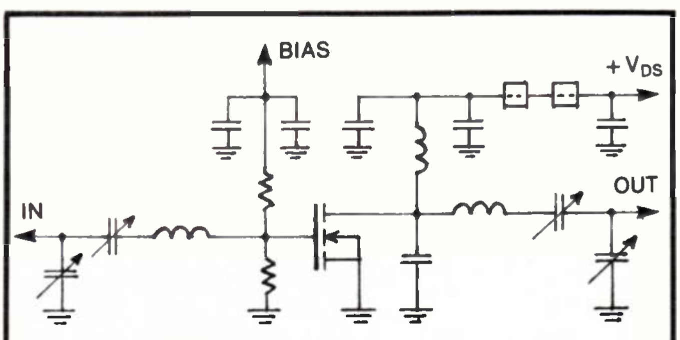


Figure 2. A typical common source MOS FET power amplifier circuit.

Since the price of a solid state device is a function of a die size, we get fewer watts per dollar. This is completely opposite from what the industry has been trying to do in the past years with bipolar transistors. So, one may ask: What is better, if anything, with the power FET's if we can get a bipolar transistor with an equal power rating for less than half the price? This is where we come to the purpose of this article, which is to discuss the characteristics of the FET and bipolar device. Both have the same basic geometry, but with some mask changes one was processed as a MOS FET and the other as a bipolar.

Circuit Configurations

Since the gate of a MOS FET device is essentially a capacitor, which consists of MOS capacitance distributed between the channel and the surface metalization, the input Q is normally extremely high. For this reason, the gate must be de-Q'ed with a shunt resistance or applying negative feedback or a combination of the two. Unless this is done properly, the affect of feedback capacitance (C_{rss}) will result in conditions, where stable operation is impossible to achieve.

Figure 1 shows a Smith Chart plot of a 150 W MOS FET and a bipolar device using the same basic geometry for comparison purposes. The gate of the FET has been shunted by a resistance of 20 ohms. Without the shunt resistance, the input impedance would be a pure capactive reactance, if package inductances are disregarded.

The input Q is an inverse function of the broadband ability of a device. With the techniques mentioned above, the Q can be controlled to a large degree, but some power gain will be sacrificed, unless only some type of selective negative feedback is employed for that purpose. Amplifiers in the 100 W power level, covering five octaves can be designed, and the limiting factor only seems to be the proper design of the broadband matching transformers.

Due to the lack of base diode junctions inherent to bipolar devices, where the diode forward conductance depends on the drive level, the MOS FET gate impedance varies only slightly with the input voltage amplitude. The gate MOS capacitance should be more or less independent of voltage, depending on the die processing. This is considered one of the advantages with FET's, especially regarding amplitude modulated applications, where a constant load for the driver stage is important. Negative feedback should be limited, since it tends to deteriorate this characteristic. Another advantage is the AGC capability by varying the gate voltage. In common source configuration, depending on the initial power gain, e.t.c., an AGC range of 20 dB is achievable.

Common gate configuration has some advantages, although it is not useful in applications requiring linearity. The load impedance is reflected back to the gate and in effect is in parallel with the source to ground impedance. The total input impedance is more constant with frequency than in common source mode, but varies greatly with output power level and supply voltage. As in a comparable configuration with bipolar transistors, the overall power gain is low, but the unity gain frequency ($f\alpha$) extends higher, which makes the common gate circuit attractive at UHF designs. It also has more tendency for parasitic oscillations, since the input and output are in the same phase. The de-Q'ing of the input can be done in the same manner as in a common source circuit, but negative feedback is not as easy to implement. This circuit also exhibits greater power gain versus bias voltage variation characteristics. In applications, where 40 dB to 50 dB AGC range is required, the common gate configuration should be considered.

A common drain configuration represents the emitter follower in bipolar circuits. In both cases the input impedance is high and the load impedance is effectively in series with the input. The input capacitance, (drain to gate, or collector to base) is lower than in common source or common gate circuits, and several times lower for the FET than bipolar for equal die size. This is due to lack of the diode junction. A MOS FET source follower can not be regarded as having current gain as the emitter follower. The amplification rather takes place through impedance transformation. Due to the fair amount of input de-Q'ing required, the available power gain is lower than in common source circuit for example. Having less than unity voltage gain, the circuit exhibits exceptional stability, and negative feedback is not necessary, nor can it be easily implemented. Push pull broadband circuits for a frequency range of 2 to 50 MHz have been designed for 200 - 300 watt power levels. Their inherent characteristics are good linearity and gain flatness without any leveling networks. High power SSB amplifiers are probably the most suitable application for common drain operation. The AGC range is comparable to that in common source, but higher voltage swing is required. It must be noted that the MOS devices used must have high gate rupture voltage, since during the negative half cycle of the input signal, the gate voltage approaches the level of V_{DS}.

Linearity Aspects

Some literature claims that MOS power FET's are inherently more linear than the bipolar transistors. This is only true up to the point where envelope distortion,

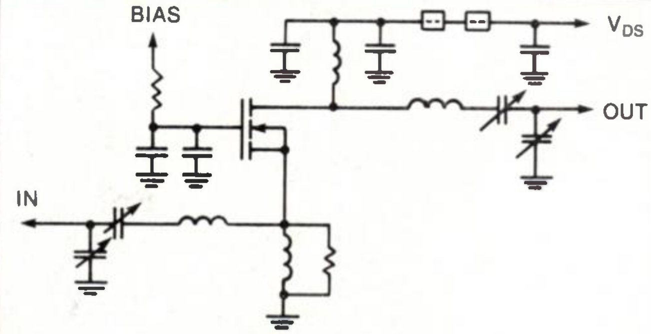


Figure 3. A typical common gate MOS FET power amplifier circuit.

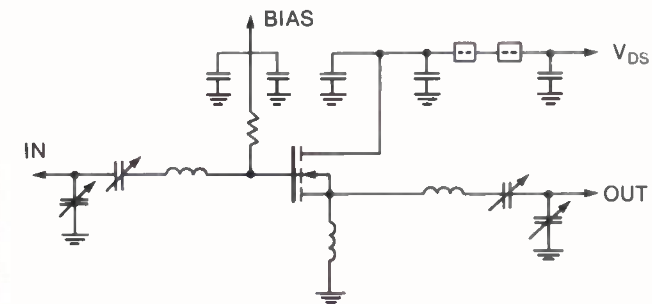


Figure 4. A typical common drain, narrow band MOS FET power amplifier circuit.

caused by saturation, instabilities or other reasons, is not present. It is also a function of the bias current (I_{DQ}). The FET's usually require higher idling currents than the bipolars to get full advantage of their linearity. Bipolars are usually biased only to get the base-emitter diode into forward conduction, whereafter increasing the bias helps little. Class A is an exception, but the device must then be operated at 20 - 25 percent of the rated Class AB level.

Probably the main advantage with the MOS power FET's is their greatly superior high order IM distortion performance. This is mainly due to the fact that ballasting resistors are not required with FET's. In bipolar RF power transistors, nonlinear feedback is distributed to each emitter site through the MOS capacitance from the collector. In devices using diffused silicon resistors, this effect is even worse, and caused by additional nonlinear diode capacitance between the collector and the emitters. The high order IMD (9th and up) is actually in direct relation to the ballasting resistor values, which must be optimized for an even power distribution along the die. Too low values would result in a fragile device, and the opposite would, in addition to the IMD problem, result in high collector - emitter saturation voltage and low power gain.

The feedback capacitance, drain to gate or collector to base for example, also has a secondary effect in IMD. In both cases it is a function of the die geometry, and is usually lower with devices with higher figure of merit, such as the ones made for UHF and microwave applications. A MOS power FET exhibits some five times lower feedback capacitance than a bipolar transistor with a similar geometry. In a bipolar transistor this capacitance partly consists of the collector - base junction, which is highly nonlinear with voltage. This, together with the varying input impedance, generates internal feedback, which is nonlinear and produce high order IMD to some degree. A more noticeable effect is that the low order IMD goes up with reduced drive levels as shown in Figure 6.

This can be related to different turn on characteristics between the two device types. When a bipolar device is biased to Class AB, the bias does not usually, completely overcome the V_{BE} knee. Thus, at lower signal levels, the remaining nonlinear portion covers a larger area of the total voltage swing. Increasing the bias from the normally recommended Class AB values will help and full Class A should eliminate the problem completely.

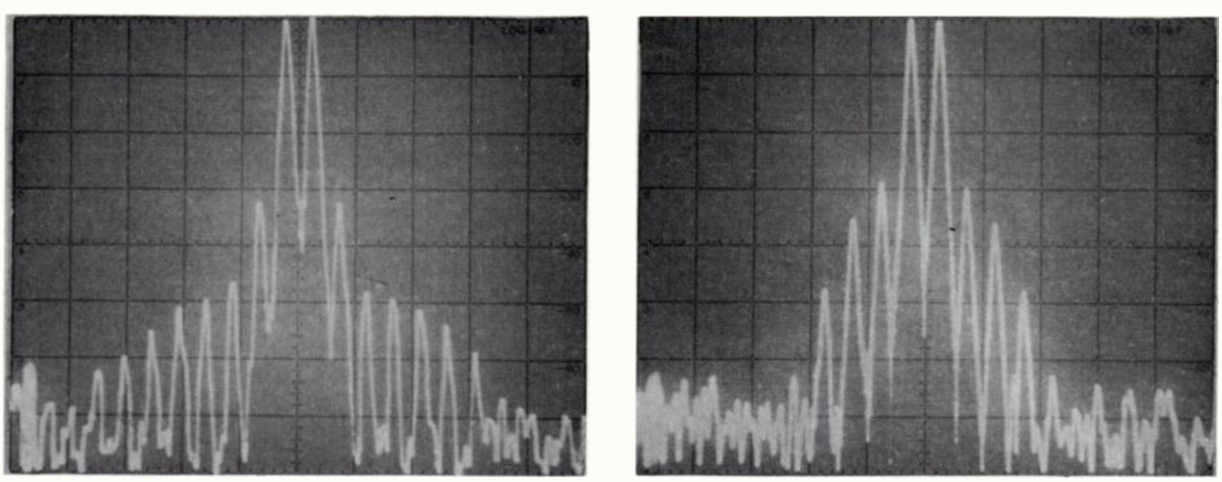
Class D/E Applications

Switching mode RF power amplifiers have only become feasible since the introduction of the power FET. Being a majority carrier device, the FET does not exhibit the storage time phenomena, that limits the switching speed of a bipolar. For a given device, the switching speed is mainly determined by the speed the gate capacitance can be charged and discharged with. If the capacitance is in the order of several hundred pf, a smaller FET is required to provide the fast charge - discharge switch. For low power stages, bipolars can be used, since the storage time is mainly an inverse function of the F_t and device size. The advantages of a Class D amplifier are high efficiency, linearity and ruggedness, since power is ideally dissipated only during the switching transitions.

These amplifiers are readily applicable for FM modulation, after harmonic filtering. The analog gain is obtained by pulse-width modulation of the input switching signal, and demodulation of the output with suitable filters. Linearity is required only from the modulator, which is easy to achieve at small signal levels. The high speed voltage controlled one shot MC10198 should be ideal for a linear pulse-width modulator. By properly adjusting its operating point, low level AM or suppressed carrier double sideband signals can be generated.

General

All MOS FET's can in theory have a positive temperature coefficient on the gate threshold voltage. This means that the gate threshold voltage increases with



a. b.

Figure 5. Two tone spectrographs of 300W PEP, 50V amplifier outputs. a. using bipolar transistors and b. with TMOS power FET's. 500mA of bias current per device was used in each case. Doubling the bias current has a minimal effect in a. but in b. the 7th order products would be lowered by 10-12 dB.

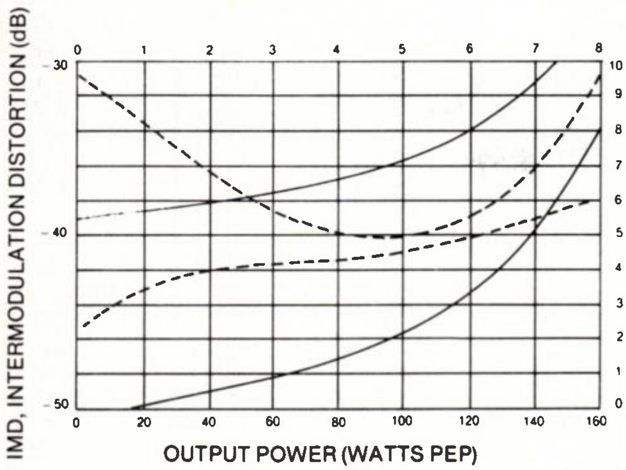


Figure 6. IM Distortion as a Function of Power Output. Solid curves MOS FET, dashed curves bipolar transistor.

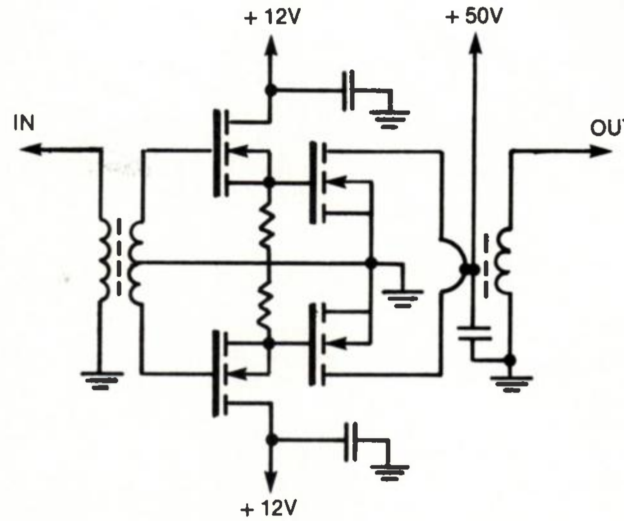


Figure 7. A typical power Mos Fet class D RF amplifier, arranged in push-pull configuration.

temperature, trying to "turn the device off." In addition the G_m will decrease, which also helps in preventing the thermal runaway, which is commonly a problem with bipolars. The coefficient of the gate threshold voltage is also a function of the drain current. Normally the coefficient is negative at low current levels, and turns positive at higher currents. The turnaround point, which can be controlled by doping and other fabrication steps, must be at a current level not to exceed the maximum dissipation rating, taking the derating factor into account. Thus, the power MOS devices can be easily biased to Class A, without fear of a thermal runaway.

Two types of high frequency noise are generated by bipolar transistors. Shot noise is caused by the forward biased junctions, and thermal noise by moving carriers upon flow of electrons. Both have different noise spectrums, and only the latter is present in a FET. In a transmitter, where the devices are biased for linear operation, the shot noise becomes a problem, especially if a receiver is in close proximity, as in transceiver designs. Also, if several stations are operated near each other, the noise can be transmitted through the antenna, disturbing the reception at near by stations. In most instances, the bias of the power devices must be switched on and off during the transmit and receive functions, which will prevent a full break in operation. Measurements of 150 W devices, intended for SSB applications, were performed at 30 MHz, at the proper idling current levels. The difference in the total noise figure between a bipolar and a FET is about three to one, or 7 dB and 2.2 dB respectively. The amount of noise that can be tolerated varies with each situation, and whether the difference above is significant in practice depends on other factors involving the design of the equipment.

Conclusion

From the above we must conclude that it is doubtfull the power FET ever will replace the bipolar transistor in all areas of communications equipment. It will have its applications in low and medium power VHF and UHF amplifiers, eliminating the need for internal matching, and up to medium power low band and VHF SSB, where

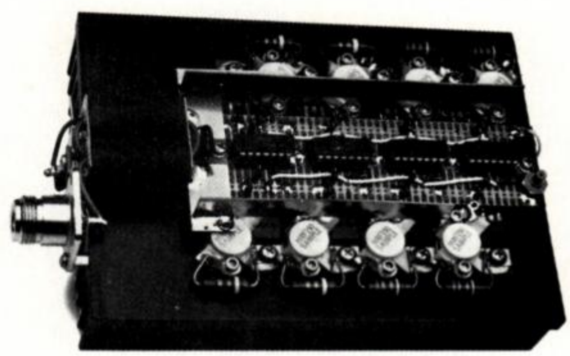
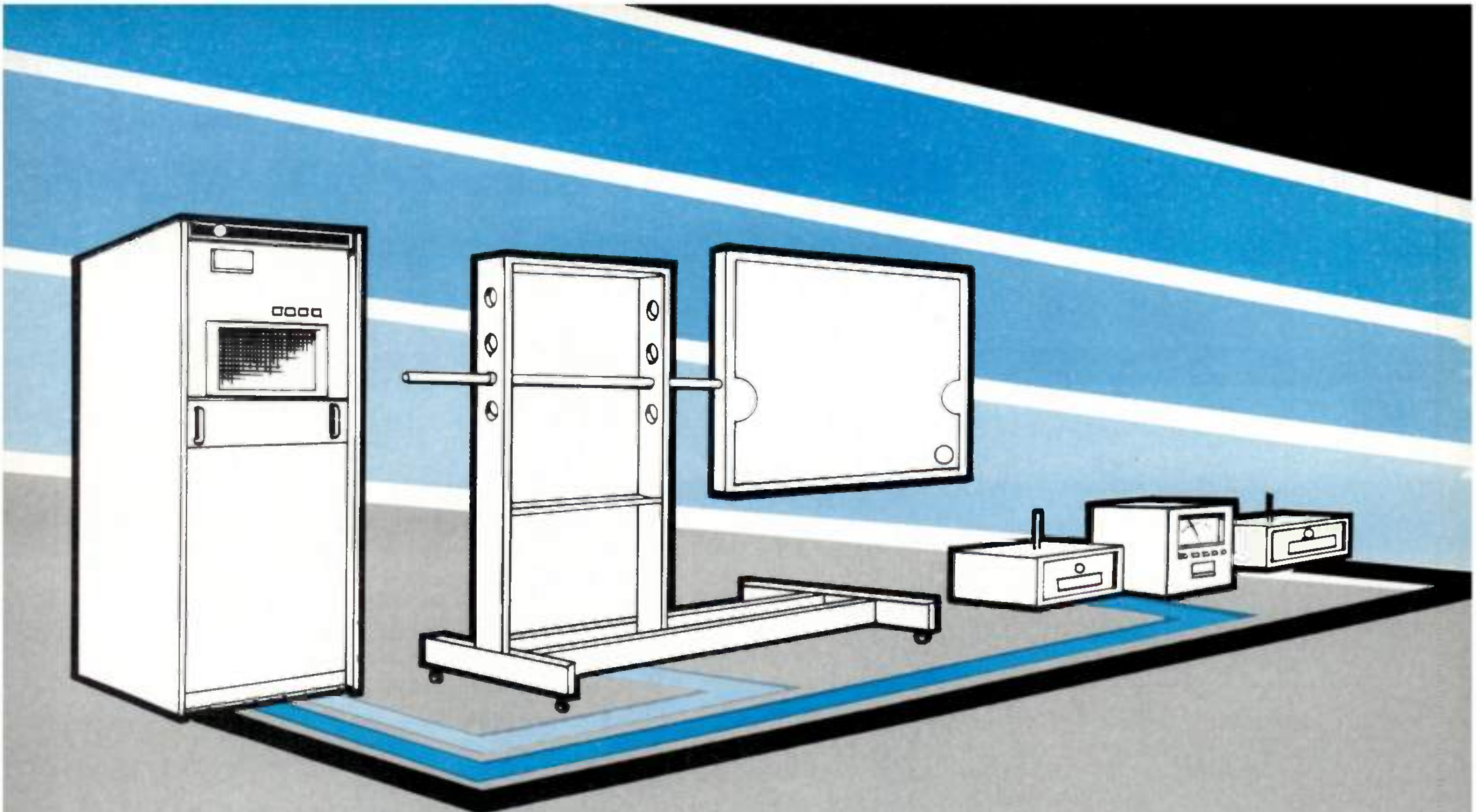


Figure 8. An experimental three stage, one kilowatt class D amplifier. The unit operates up to 10 MHz yielding an efficiency of 85 percent. The power gain is 30 dB.

the high order IMD is beginning to be more and more in emphasis due to the crowded frequency spectrums. The authors personal opinion is that the power FET is the most feasible device for the amplitude compandored sideband (ACSB) applications, proposed for future use in land mobile communications. The system principle requires extreme linearity in the amplifying stages, which in the past has only been achieved with Class A operation. The power FET also opens new applications for high efficiency switching mode power amplifiers, which have not been possible in the past for reasons described earlier. The possible upper frequency limit would be dictated by the physical lay-out of the system.

The author wishes to thank Mr. Bob Johnsen of Motorola Semiconductor Products, Inc. for valuable information related to the wafer processing of services discussed.



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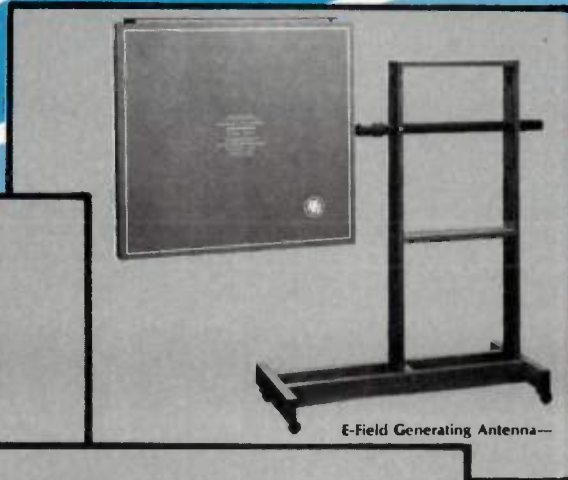
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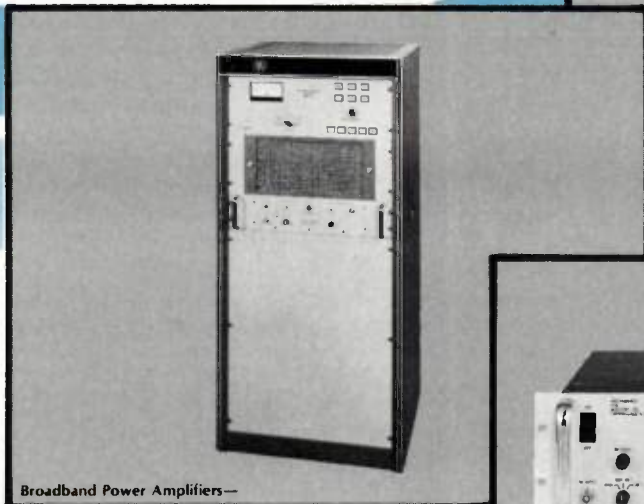
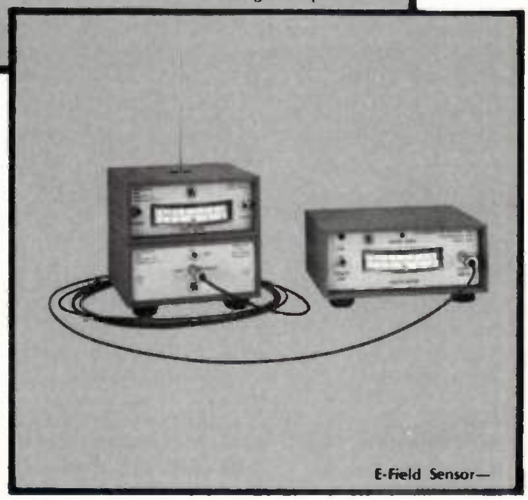

Broadband Power Amplifiers—


Leveling Preamplifier—


E-Field Sensor—

M400 Series

The M400 Series of power amplifiers provide a choice of outputs from 100 watts to 15 kilowatts over a broad bandwidth that ranges from 10 kHz to 220 mHz. Each model can be driven to maximum output by a standard signal generator and there is no tuning or adjustments necessary for true broadband operation. Conservatively rated and designed for long and reliable operation, the M400 Series amplifiers can provide the best in broadband amplifier performance.

LPA-2

Utilizing a preamplifier with over 40 dB of dynamic gain control and a sophisticated microcomputer controller, the LPA-2 provides the automated control necessary to conduct swept-field testing without the need for a dedicated minicomputer. Designed to work in conjunction with IFI E-Field sensors, this Leveling Preamplifier enables the user to complete a closed-loop system for the precise regulation of electric field strength in screen rooms and test chambers.

EFG-3

Designed to generate strong electric fields for radiated susceptibility testing, the EFG-3 performs efficiently from 10 kHz to 220 mHz, at power levels up to two kilowatts. It is physically small enough for use in most shielded rooms, yet large enough to irradiate a useful area in the broadside mode.

EFS-1

A truly broadband device, the EFS-1 will provide direct readout in Volts/Meter at any frequency between 10 kHz and 250 mHz. It is calibrated in five convenient ranges, from 1 to 300 Volts/Meter and its small size allows positioning in any test area with neglible effects.

The Quality:

Versatile. Accurate. Cost-effective. Whether you require a basic power amplifier that works with any load impedance, or an E-Field Sensor with remote readout, IFI gives you the highest quality, rugged construction for accurate testing, measurement and performance.

And IFI provides the options, accessories and design capabilities to meet any specialized—or critical—requirements.

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INFO/CARD 6

Part II

High-Frequency Transistor Amplifier Design

The second part of a three part, detailed investigation into practical Smith chart design.

By:
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Part 1 of this article (r.f. design, Sept./Oct. 1981) reviewed in detail the Smith Chart, a useful tool which will aid R.F. circuit analysis and design. Among the topics discussed were development of the chart, basic graphical manipulations such as inversion and conjugates, impedance transformation, and matching. Graphical behavior of both lumped and distributed circuit elements was considered.

Also treated in Part I were the characteristics of typical active devices (transistors) used in the design of practical amplifiers. Negative feedback and reactive mismatch were introduced as methods by which the designer can manipulate the terminal behavior of transistor amplifiers. Reactive mismatch, which lends itself readily to both graphical and analytic techniques, is selected for detailed discussion and design examples. Part I concluded after presenting a concept of cascaded passive networks and active devices. The passive network components are selected such that they modify the performance of the active device, resulting in desired amplifier characteristics.

Part II continues development of the reactive mismatch approach, arriving at a generalized design procedure involving constant-gain circles superimposed on the Smith Chart. Active device parameters are evaluated for potential oscillation problems, and corrective measures implemented which ensure an unconditionally stable amplifier. Finally, these techniques are demonstrated in the step-by-step development of two recently produced amplifiers. Confidence is developed in the design techniques by following these two projects from start to finish; initial analysis and design, computer-aided optimization, and prototype testing.

Constant-Gain Circles

As previously stated, the unilateral gain of a single-stage amplifier is equal to the product of G_S, $|S_{21}|^2$, and G_L (see Part I, Fig. 13A). At a given frequency of interest, G_S and G_L may either increase or decrease the amplifier overall gain. This is because the corresponding Γ_S and Γ_L may present either a better or a worse (compared to Z_O) match to the device terminals. The maximum value of G_S occurs when $\Gamma_S = S_{11}^*$ and the "matching" gain is:

$$G_{SMAX} = 1/(1 - |S_{11}|^2)$$

For a desired value of G_S, there exists an infinite number of Γ_S which will provide the correct gain from a given S_{11}. Similarly, there are an infinity of candidate Γ_L to produce a desired G_L. On the Smith Chart, these Γ_S (or Γ_L) are defined by the circumference of a circle. The center of the circle will lie on a line drawn from Z_O (origin) to S_{11}^*. The normalized distance from Z_O to the center of the Γ_S circle is:

$$d_S = \frac{g_S | S_{11} |}{1 - | S_{11} |^2 (1 - g_S)}$$

The radius of the circle is:

$$p_S = \frac{\sqrt{1 - g_S}\,(1 - | S_{11} |^2)}{1 - | S_{11} |^2 (1 - g_S)}$$

where $g_S = G_S$ (desired)/G_{SMAX}

$$G_{SMAX} = 1/(1 - Z_O^2)$$

To design for a required G_S, we first plot S_{11}^*, then construct the proper constant-gain circle. By designing an input network which transforms the generator impedance to a Γ_S on this circle, we achieve the correct matching gain.

Stability Considerations

An active device is said to be unconditionally stable if no combination of passive source and load impedances will cause the circuit to oscillate. A device which has the capability of oscillating into passive terminations may be identified by evaluating the stability factor K.

$$K = \frac{1 + | \Delta |^2 - | S_{11} |^2 - | S_{22} |^2}{2 | S_{21} S_{12} |}$$

Where $\Delta = S_{11} S_{22} - S_{12} S_{21}$

If $K > +1$, the device is unconditionally stable. If $K < 1$, the device is potentially unstable. For the potentially unstable device there are passive impedances which, when used to terminate on port,cause the other port to present a reflection coefficient greater than unity (negative resistance). These passive terminations which may lead to oscillation may be identified by plotting stability circles on the Γ-plane (Smith Chart).

Location of center of source impedance circle:

$$rs_1 = \frac{C_1^*}{| S_{11} |^2 - | \Delta |^2}$$

Radius of source impedance circle:

$$Rs_1 = \frac{| S_{12}S_{21} |}{| S_{11} |^2 - | \Delta |^2}$$

Location of center of load impedance circle:

$$rs_2 = \frac{C_2^*}{| S_{22} |^2 - | \Delta |^2}$$

Radius of load impedance circle:

$$Rs_2 = \frac{| S_{12}S_{21} |}{| S_{22} |^2 - | \Delta |^2}$$

where $C_1 = S_{11} - \Delta S_{22}^*$

$C_2 = S_{22} - \Delta S_{11}^*$

These circles define the boundaries between stable and unstable regions, After plotting, an individual impedance point must be evaluated to determine whether the interior or the circle is the unstable region. This is done by calculating whether terminating one port with the test impedance produces $| \Gamma | > 1$ at the other port. Equations to be used for these calculations are as follows:

Input reflection coefficient with arbitrary load:

$$S_{11}' = S_{11} + \frac{S_{12} S_{21} \Gamma_L}{1 - S_{22} \Gamma_L}$$

Output reflection coefficient with arbitrary source:

$$S_{22}' = S_{22} + \frac{S_{12} S_{21} \Gamma_S}{1 - S_{11} \Gamma_S}$$

The stability calculations required are tedious and involve complex arithmetic. To reduce the possibility of error, it is recommended that the equations be entered into a computer or a programable calculator. For the amplifier design examples to be discussed in this article, stability calculations were performed using a commercial timeshared computer.

Design Procedure

The basic skills reviewed may now be applied to the design of practical amplifiers. The following is an outline of a suggested approach to amplifier design. This procedure will be used for the two design examples. Note: Steps 4 and 6 are optional.

1. Device Selection: Factors affecting choice of transistors are too numerous to list. A good rule is to be conservative within reason. It is false economy when a technician must spend hours extracting the last ½ dB of performance from a device which offers a $5 initial cost savings.

2. Stability: Stability should be investigated for at least two frequencies. In the band of intended use, the device may be matched for near-maximum gain and become unstable. At low frequencies, intrinsic gain may be much higher. This, combined with matching networks which may be highly reactive at these low frequencies, has been the source of out-of-band oscillations in many otherwise sound amplifier designs.

3. Gain Compensation: the amount of gain compensation required may be determined by either of two methods. Assuming a 6 dB/octave device rolloff, we may approxi-

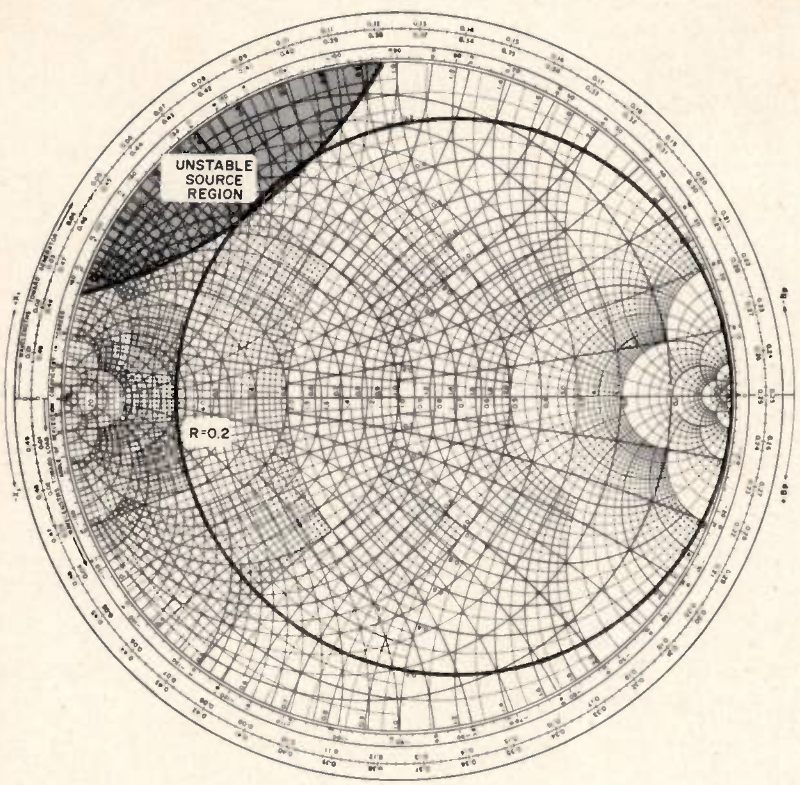

Figure 14. Input Stability.

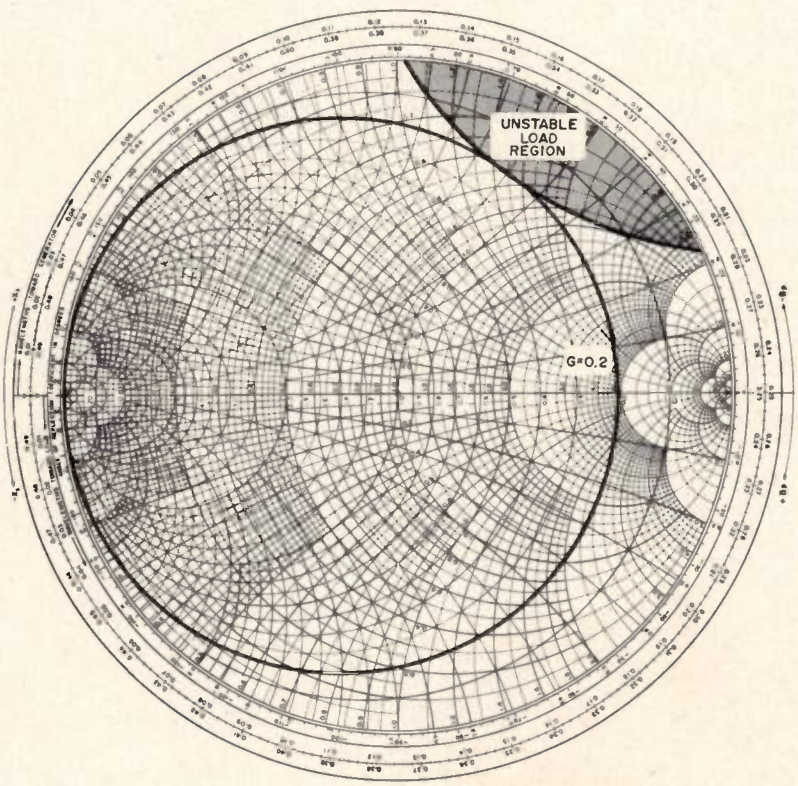

Figure 15. Output Stability.

mate the gain decrease in the frequency interval f_1 to f_2 as follows:

$$\Delta G\ (dB) = 20 \times N \times \log\ (f_1/f_2)$$

where N is the number of transistors. This method is adequate for most designs. For very wide bandwidths, or where the amplifier is to be constructed directly from the initial calculations (no computer-aided optimization), more accuracy may be required. In this case, calculate the exact difference in manufacturer's data for $|S_{21}|$ at the two frequencies. Budget the required compensation between the available input, interstage, and output networks, and plot constant-gain circles. Two sets of circles will be needed, since G_S and G_L requirements will be different at f_1 and f_2. The matching networks designed must satisfy both conditions. Γ_S (or Γ_L) must fall on one circle at f_1 and another circle at f_2. The f_2 circles, of of course, must provide ΔG more gain than the f_1 circles. This will produce a flat gain response in the interval from f_1 to f_2.

4. Optimize component values for gain flatness (if using computer-aided design).

5. Design input and output matching networks (unless already constrained by step 3). If step 4 was performed, match to the input and output impedances obtained from the computer analysis. Otherwise match to S_{11} of the input transistor and S_{22} of the output transistor. An alternate non-computer method, offering more accuracy would be to match to an S'_{11} and S'_{22} calculated from the equations given in the section on stability.

6. Simultaneouly optimize components values for gain flatness and VSWR (if using computer-aided design).

Design Example #1 — Single Stage Amplifier

This unit was designed to implement a required modification in an existing amplifier subsystem. Narrowband parametric amplifiers in the 440 MHz frequency range were being supplied to several customers. Frequent inquiries were received requesting higher gain than was avail-

Frequency	S_{11}	S_{21}
400 MHz	$.62\angle{-100°}$	$11.86\angle 119.8$
440 MHz	$.62\angle{-103°}$	$11.53\angle 117.9$
480 MHz	$.61\angle{-106°}$	$11.20\angle 115.9$

able from the single-stage paramp. The proposed design would be a small transistor amplifier capable of being mounted within the existing paramp enclosure. Instantaneous bandwidth (400-480 MHz) must exceed the tuning range of the paramp. Performance objectives are low noise figure, low input and output VSWR, unconditional stability, and flat gain of 15 to 20 dB. Additionally, it is desired that the circuit contribute as little as possbile to an overall MTBF (mean time between failure) calculated at the operating temperature of 60°C. This will be aided by using the simplest topology which will meet the performance goals, keeping component count to a minimum.

The NE64535 transistor, biased at 8 volts V_{CE} and 7 mA I_C, is chosen for this design. In addition to reducing component stress, this low bias level is indicated by manufacturer's data to provide minimum noise figure. S-parameters in the band of interest are as follows:

Device Stabilization

At 440 MHz, the stability factor K has a calculated value of 0.54, indicating potential instability. Stability circles must now be plotted to determine what impedances can cause oscillation, and steps must be taken to ensure that the device never sees these impedances.

Circle	Center Location	Radius	Stable Region
Input	$2.18\angle 127°$	1.47	Outside
Output	$1.93\angle 57°$	1.20	Outside

The input circle is plotted in Figure 14 and the output in Figure 15. Notice that the source reflection coefficient will not enter the unstable region if its normalized resistance component is 0.2 or greater (Figure 14). The circuit may, therefore, be stabilized at the input by adding a series resistor of 0.2 (50)= 10 ohms. Similarly, it is seen from Figure 15 that a load having a normalized conductance component of 0.2 or less will be stable. The

S_{12}	S_{22}
$.026\angle 36.6°$	$.67\angle -36°$
$.028\angle 36.8°$	$.66\angle -36°$
$.029\angle 36.9°$	$.65\angle -37°$

circuit may be stabilized at the output by adding a shunt resistor of 50/0.2=250 ohms. Note: It is not necessary to include both the input and output resistors (see Figure 16). Either is sufficient to ensure unconditional stability.

Since one of the amplifier design criteria was low noise figure, the transistor will be stabilized using a parallel output resistor. Resistive input losses would add directly to overall noise figure. A lossy output network will degrade the output power capability, but this was not a critical parameter. Both approaches cause a reduction in gain.

Gain Compensation

Expected active device gain variation over the operating bandwidth is:

$$\Delta G\ (dB) = 20 \log \left(\frac{400}{480}\right) = -1.6\ dB$$

The gain could be flattened using the topology of Figure 17, allowing G_S and G_L to each provide 0.8 dB mismatch at 400 MHz and a conjugate match at 480 MHz. However, the 0.8 dB mismatch loss represents a reflection coefficient of 0.41 at 400 MHz. The corresponding VSWR of 2.4:1 does not satisfy the design goal of "low" VSWR.

An alternate technique for gain compensation of amplifier designs constrained by VSWR requirements is the absorbtive diplexer. Stability requirements have already necessitated a lossy output network. Why not make its loss frequency-dependent, such that the difference in power absorbed at the low and high frequencies exactly compensates for the transistor's gain variation (see Figure 18). The input network may now provide a lossless conjugate match. Stability will be preserved, provided the equivalent shunt resistance presented to the transistor is 250 ohms or less. Circuit complexity will be reduced, since a single network will be used to implement stabilizing, gain flattening, and output impedance matching functions. The topology used to implement the diplexer may also serve as a bias feed to the RF transistor.

Diplexer Design

A simple shunt diplexer may be constructed from two components (see Figure 19A). The ratio of voltage across R to applied voltage will decrease by 6 dB/octave, giving an absorbtion characteristic which may be used to compensate for the single-pole rolloff of device gain.

For ease in graphical manipulation and circuit analysis, the spot-frequency parallel equivalent circuit will be derived for the diplexer (refer to Figure 19B).

$$Y_P = Y_S \text{ (at one frequency only)}$$

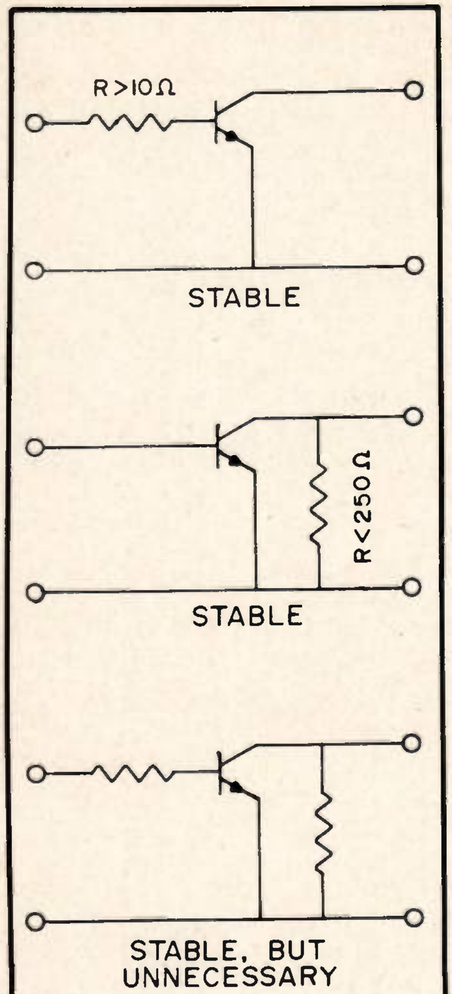

Figure 16. Stability Circuit Options.

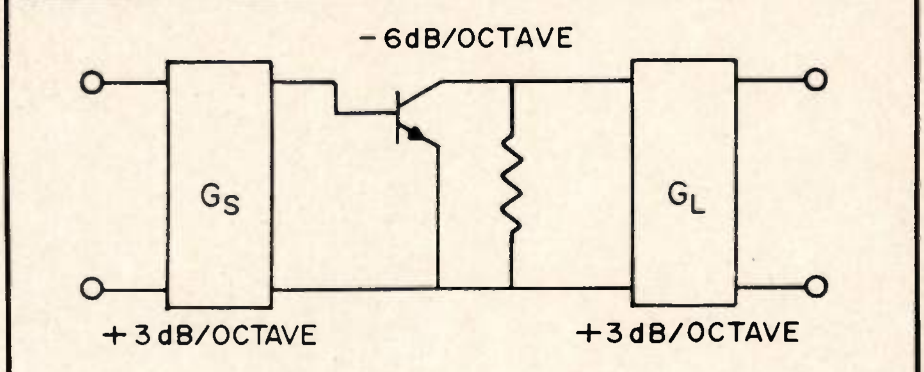

Figure 17. Candidate Gain Compensation.

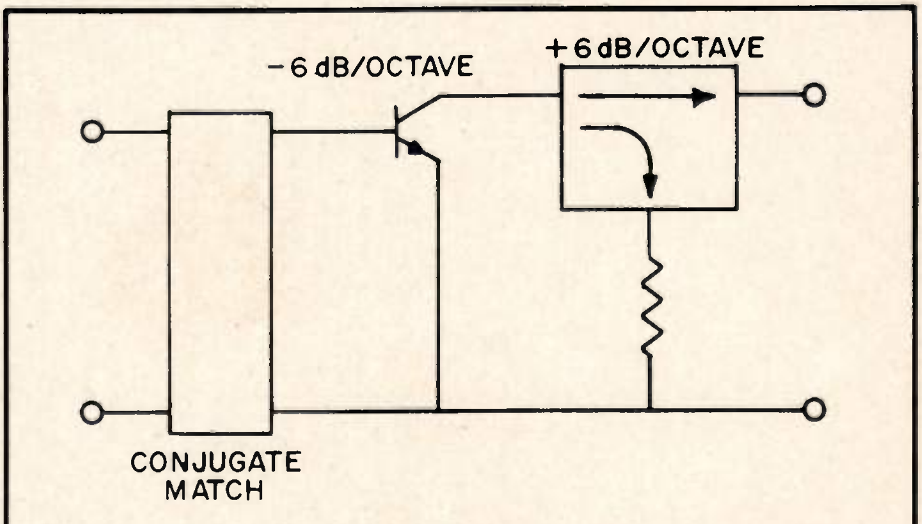

Figure 18. Alternate Gain Compensation.

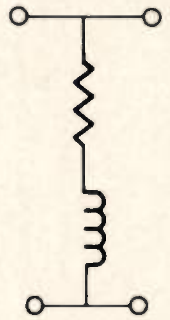

Figure 19A. Shunt Diplexer.

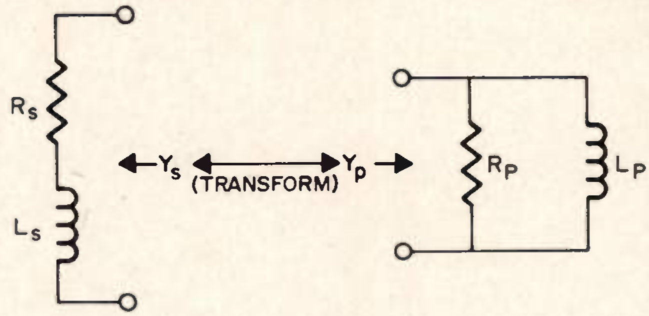

Figure 19B. Series To Parallel Transformation.

$$\frac{1}{R_P} + \frac{1}{j\omega L_P} = \frac{1}{R_S + j\omega L_S}$$

$$\frac{R_S - j\omega L_S}{R_S^2 + \omega^2 L_S^2}$$

Equate, separately, the real and imaginary parts of Yp and Ys. Solve for Rp and Lp in terms of Rs, Ls, and ω.

Real:

$$\frac{1}{R_P} = \frac{R_S}{R_S^2 + \omega^2 L_S^2}$$

$$= \frac{R_S^2 + \omega^2 L_S^2}{R_S}$$

Imaginary:

$$\frac{1}{j\omega L_P} = \frac{-j\omega L_S}{R_S^2 + \omega^2 L_S^2}$$

$$j\omega L_P = \frac{R_S^2 + \omega^2 L_S^2}{-j\omega L_S}$$

$$L_P = \frac{R_S^2 + \omega^2 L_S^2}{\omega^2 L_S}$$

The inverse transforms are derived similarly, giving

$$R_S = \frac{\omega^2 R_P L_P^2}{R_P^2 + \omega^2 L_P^2}$$

$$L_S = \frac{R_P^2 L_P}{R_P^2 + \omega^2 L_P^2}$$

Design procedure will be to graphically determine, at center frequency, values for the parallel equivalent circuit that provide good output impedance match. After checking

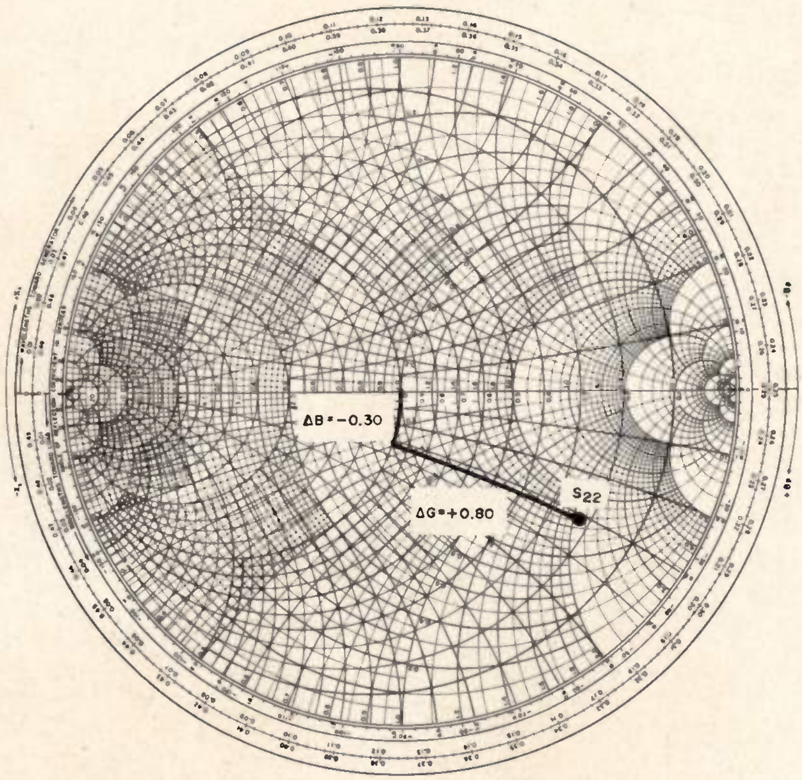

Figure 20. Diplexer Design.

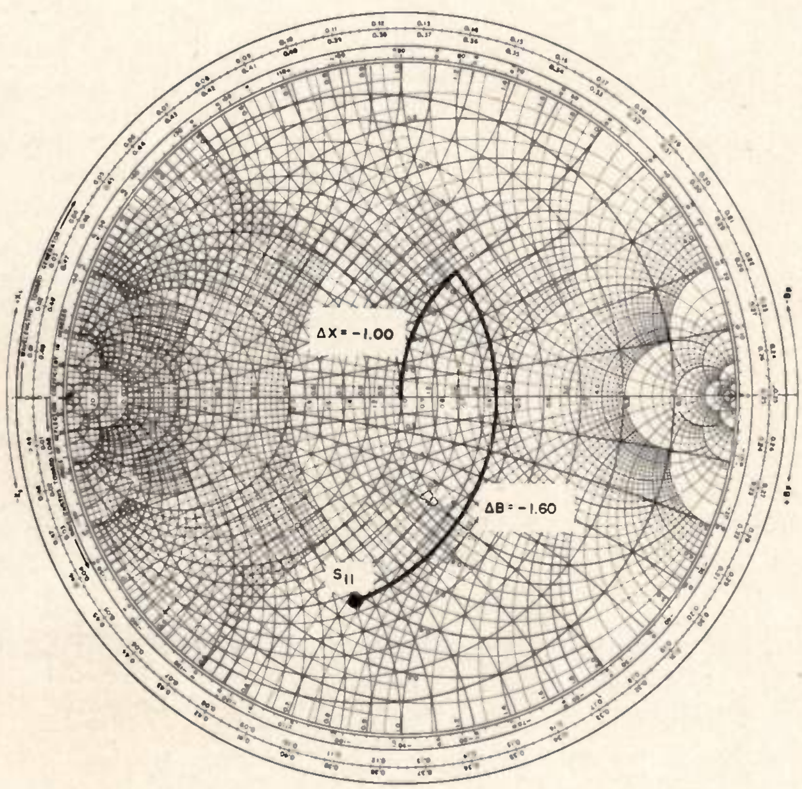

Figure 21. Input Match at 440 MHz.

that Rp will provide stability (250 ohms or less), the inverse transforms are used to determine the actual values (Rs and Ls) to be used in the circuit. The parallel circuit solution is illustrated in Figure 20. Component values are calculated using the formulas presented in Part I of this article (r.f. Design, Sept/Oct 1981).

$$R_P = \frac{50}{\Delta G} = \frac{50}{0.8} = 62.5 \text{ ohms (Stable)}$$

$$XL_P = \frac{50}{\Delta B} = \frac{50}{0.3} = 167 \text{ ohms}$$

$$L_P = \frac{167}{2\pi \times 440 \times 10^6} H = 60.4 \text{ nH}$$

From the inverse transforms, $L_S = 7.42$ nH and $R_S = 54.74$ ohms.

Input Match

Input match will be designed at center frequency using the simplest network possible, then evaluated for acceptable VSWR at the band edges. Firgure 21 diagrams the input match design. Here, careful selection of matching network topology can reduce component count. The parallel coil doubles as the base bias feed to the transistor and the series capacitor also provides DC isolation from the input connector. Component values are calculated as follows:

$$X_C = 50(\Delta X) = 50(1.0) = 50 \text{ ohms}$$

$$C = \frac{1}{50 \times 2\pi \times 440 \times 10^6} F = 7.23 \text{ pF}$$

$$X_L = \frac{50}{\Delta B} = \frac{50}{1.6} = 31.25 \text{ ohms}$$

$$L = \frac{31.25}{2\pi \times 440 \times 10^6} H = 11.3 \text{ nH}$$

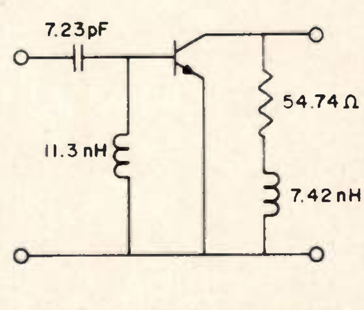

Figure 22. R.F. Schematic.

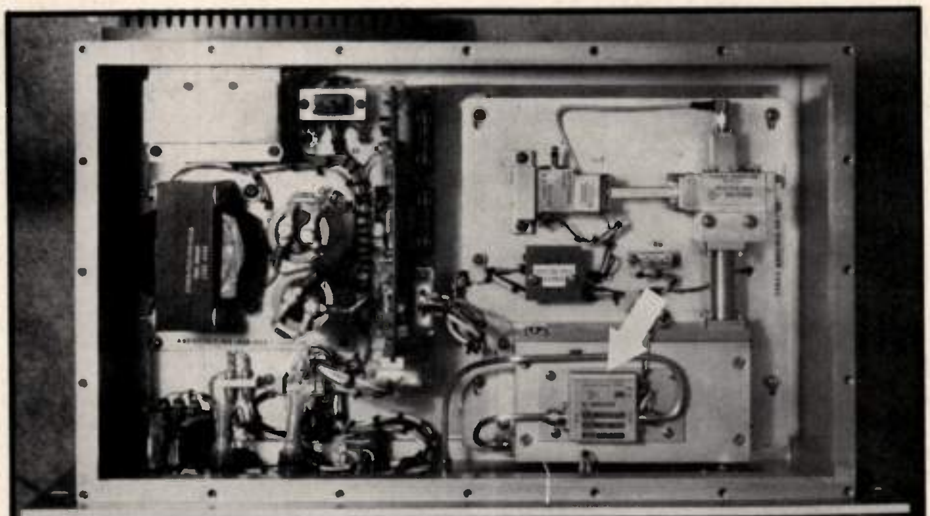

Figure 23. Mounted Amplifier.

Computer Analysis				
Frequency	**VSWR (In)**	**VSWR (Out)**	**Gain**	**K-Factor**
400 MHz	1.99:1	1.35:1	19.20 dB	2.50
440 MHz	1.28:1	1.32:1	19.49 dB	2.34
480 MHz	1.20:1	1.27:1	19.28 dB	2.30

Results

Figure 22 is a schematic of the complete RF circuit. To check the accuracy of the design, a computer analysis was performed using the graphically determined component values. Notice how little error was introduced by our unilateral assumption.

A prototype unit was constructed and, after slight "tweaking," the following performance was measured over 430-450 MHz (actual paramp tuning range) at the elevated temperature of 60°C:

Gain :	18 dB min
Input VSWR :	1.5:1 max
Output VSWR :	1.15:1 max
Noise Figure :	1.4 dB max

In its initial application, gain of the amplifier was reduced to a requirement of 12 dB by inclusion of a resistive "π" attenuator. Figure 23 is a photograph of the 440 MHz amplifier as installed within the paramp enclosure.

Next issue: Design example two-stage amplifier.

Errata to Part I

Page 27, second paragraph, first sentence, should be:

...consider $\Gamma = 0.5 \angle -150°$. The complex conjugate, Γ^*, is therefore $0.5 \angle +150°$.

Page 30, series transmission lines, statement 3 should be:

Plot Z_1 and rotate...

Page 30, quarter-wave transformer, statement 2, should be:

$$R'_1 = \frac{1}{R_1} = \sqrt{Z_0 R/R}$$

Page 32, an entire equation is omitted in the feedback analysis, should be:

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

$$\frac{V_2}{V_1} = \frac{100}{1 + 100\,(0.1)} = \frac{100}{11} \approx 9.09\ (19.17\ dB)$$

Suppose A is the...

Quarter Wavelength Taper Line Matching

HP25 programs for designing tapered and multisectional

By Andrzej B. Przedpelski
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A.R.F. Products, Inc.
Boulder, Colorado

It is difficult to find in the literature a comparison of the quarter wavelength and exponential taper line matching transformers. In particular, the treatment of the tapered line transformer and the multisection quarter wavelength line transformer is either difficult to apply to everyday applications or is theoretically too difficult for the average designer to apply to his work.

This article compares the different types of transformers on the same basis and provides comparatively easy approaches to solutions of the more difficult calculations. With the advent of the simple programmable calculator, these solutions become within reach of the designer who does not have access to a computer. The programs are based on the HP-25 calculator capabilities, since up to 49 steps are required. The programs are not very sophisticated so that they can be modified easily for use with other calculators or for special applications.

The calculations deal with the design of line matching transformers using microstrip or stripline techniques. Methods to determine their bandwidth are also described. Since the microstrip or stripline techniques are used mainly at higher RF frequencies, the bandwidth presentation in terms of reflection coefficient, Γ, which can easily be converted to other parameters of interest.

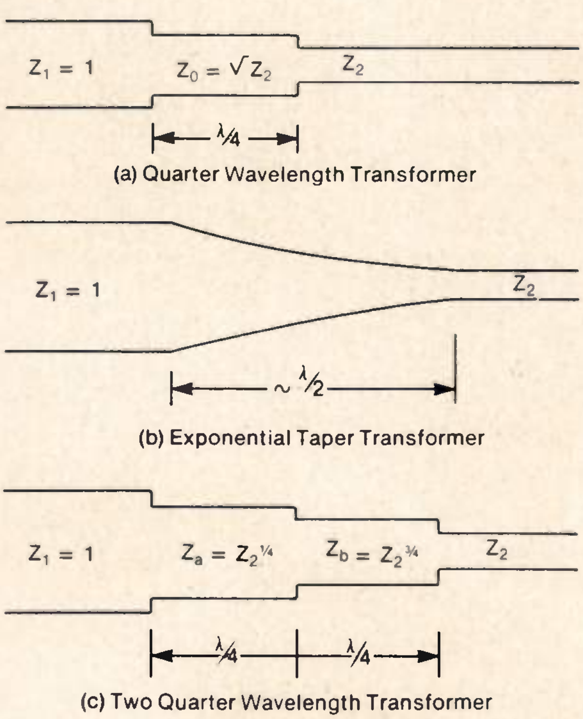

Figure 1. Transmission Line Transformers (Normalized with respect to Z_1)

Quarter Wavelength Matching Transformer

The quarter wavelength matching transformer is probably the one most commonly used at higher RF frequencies. The two real impedances are matched by a length (one quarter of the wavelength) of a transmission line with a characteristic impedance:

$$Z_0 = (Z_1 \times Z_2)^{1/2}$$

where Z_1 is the real input impedance Z_2 is the real output impedance. The perfect match ($\Gamma=0$) is obtained only at frequencies whose wavelength is 4, 12, 20, 28, etc. times the line length. (Usually, however, only the condition where the line is one quarter the wavelength is of interest.) At all other frequencies a mismatch occurs.

and Exponential Transformers

quarter wavelength line transformers.

This mismatch can be calculated by first determining the complex input when terminated with impedance Z_2:

$$Z_{in} = R_{in} + jX_{in} = Z_0 \left[\frac{Z_2 + jZ_0 \tan\theta}{Z_0 + jZ_2 \tan\theta}\right]$$

where θ = electrical line length in degrees or radians and then converting to the reflection coefficient:

$$\Gamma = \frac{Z_1 - (R_{in} + jX_{in})}{Z_1 + (R_{in} + jX_{in})} = \frac{1 - \frac{Z_2}{Z_1}}{1 + \frac{Z_2}{Z_1} j2 \sqrt{\frac{Z_2}{Z_1}} \tan\theta}$$

Curves for three values of Z_2/Z_1 (2, 10 and 100) are shown in Fig. 2. The program used to obtain these curves is shown in Table I and can be used to obtain data for other values of Z_2/Z_1.

Tapered Line Matching Transformer

Several possible configurations of the tapered line transformer exist. However, only the most common configuration, as shown in Fig. 1 (b), will be discussed. Basically, the line has an exponential taper, i.e. the line impedance changes exponentially with length and the characteristic impedance at the two ends is the same as the two impedances to be matched.

Arnold, Bailey and Vaitkus (Ref. 2) derive an expression for the reflection coefficient of a tapered line from the basic relationships cited by Womack (Ref. 1). The resulting equations are difficult to use but can be simplified as shown below:

$$\Gamma = \frac{\ln \frac{Z_2}{Z_1} \tan \tfrac{1}{2} \sqrt{4\theta^2 - (\ln \frac{Z_2}{Z_1})^2}}{\sqrt{4\theta^2 - (\ln \frac{Z_2}{Z_1})^2} + j2\theta \tan \tfrac{1}{2} \sqrt{4\theta^2 - (\ln \frac{Z_2}{Z_1})^2}}$$

This equation avoids the use of hyperbolic functions and interrelated equations as derived by Arnold et al. Its main disadvantage is that data cannot be calculated for values of θ less than ½ln Z_2/Z_1. Fortunately, this area is usually of little interest, since for these line lengths the reflection coefficient approaches the zero length value very closely, as will be shown later.

The reflection coefficient of an exponentially tapered

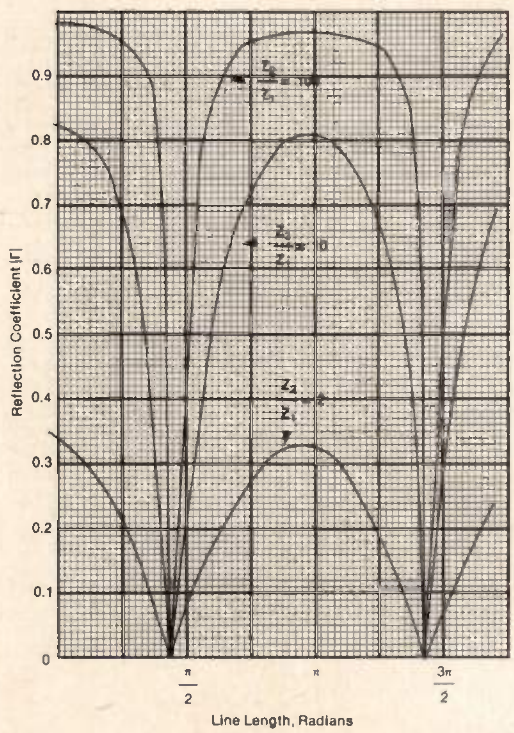

Figure 2. $\frac{\lambda}{4}$ Line Transformer

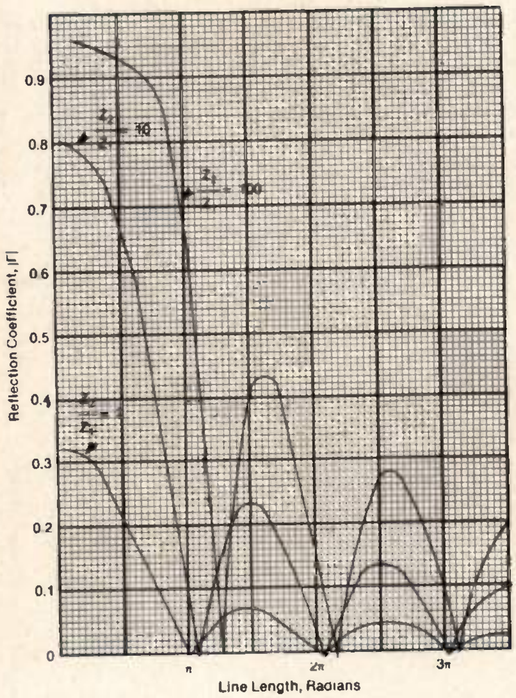

Figure 3. Exponential Taper Line Transformer

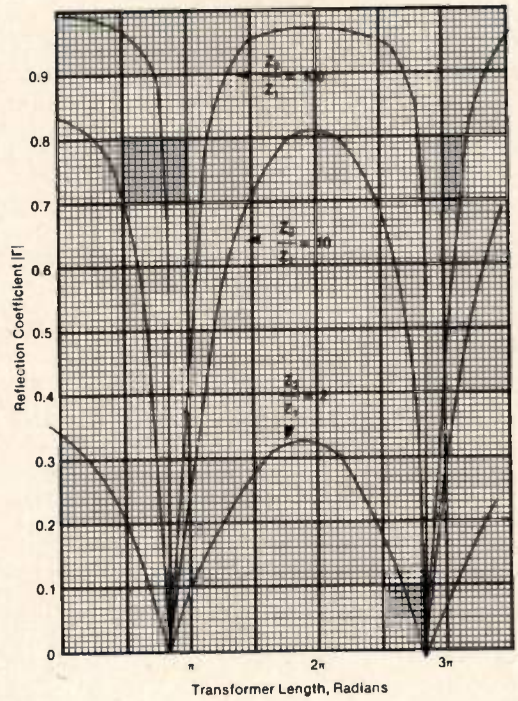

Figure 4. $2 \times \frac{\lambda}{4}$ **Line Transformer**

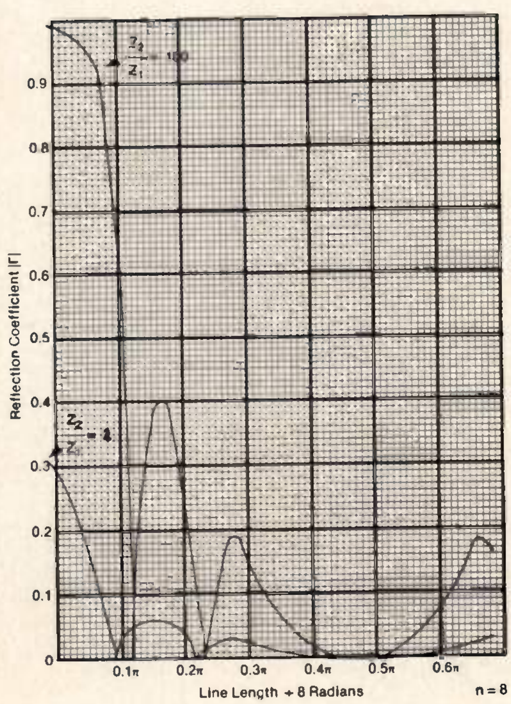

Figure 5. $8 \times \frac{\lambda}{4}$ **Line Transformer**

Display Line	Display Code	Key Entry	x	y	z	t	Comments
00							
01	24 06	RCL 6					
02	14 06	(f) tan					
03	02	2					
04	61	x					
05	24 03	RCL 3					
06	14 02	[illegible]					
07	61	x					
08	24 03	RCL 3					
09	01	1					
10	51	+					
11	15 09	(g) →P					
12	24 03	RCL 3					
13	32	CHS					
14	01	1					
15	51	+					
16	71	÷					
17	15 22	(g) [illegible]					
18	15 03	g ABS					
19	13 00	GTO 00					
20							

Registers: R ___ ; R ___ ; R ___ ; R Z_2/Z_1 ; R ___ ; R ___ ; R ___

Step	Instructions	Input Data/Units	Keys				Output Data/Units
1	Store	Z_2/Z_1	STO	3			
		θ	STO	6			
2	Calculate		R/S				Γ

Table I
Quarter Wavelength Line Reflection Coefficient

transformer, as shown in Fig. 1 (b), is plotted in Fig. 3 for three values of Z_2/Z_1 (2, 10 and 100). Reflection coefficients for other Z_2/Z_1 ratios can be derived using the program shown in Table II. If values of 0 less than ½ln Z_2/Z_1 are tried, the calculator will give an ERROR indication. Thus erroneous values cannot be obtained.

It will be noted from Fig. 3 that the matching points do not occur at regular intervals as for the quarter wavelength line matching. This is particularly noticeable for larger values of Z_2/Z_1. The program of Table III can be used to obtain the exact matching points in accordance with equation:

$$\theta_n = \sqrt{n^2 \pi^2 + (\tfrac{1}{2} ln \frac{Z_2}{Z_1})^2}$$

where n = 1, 2, 3 etc.

Multi Quarter Wavelength Matching Transformer

Figs. 2 and 3 show that the tapered line transformer is at least twice as long as the simple quarter wave line transformer. To make the comparison between the two more valid a dual quarter wave transformer, as shown in Fig. 1 (c), was also considered.

To obtain the reflection coefficient of this transformer, equation (2) is used first to obtain the complex impedance at the junction of the two quarterwave sections. (Z_b is used in place of Z_o). Then, these values for R and X are used in:

$$R_{in} + jX_{in} = Z_a \left[\frac{R + j(X + Z_a \tan\theta)}{(Z_a - X \tan\theta) + jR \tan\theta} \right]$$

to obtain the complex input impedance to Z_a. Equation (3) is then used to obtain the reflection coefficient of the transformer. (The same 0 is used in equations (2) and (6) and is the length of one line section. Thus, the total length of the transformer is 20 radians). Reflection coefficient for the same Z_2/Z_1 ratios (2, 10 and 100) are plotted in Fig. 4. The reflection coefficient for other Z_2/Z_1 ratios can be calculated using the program shown in Table IV. This program takes up th 49 steps available in the HP-25 calculator, thus equation (2) has to be solved separately,

Display Line	Display Code	Key Entry	X	Y	Z	T	Comments
00							
01	15 33	(g) RAD					
02	24 07	RCL 7					
03	02	2					
04	61	×					
05	23 06	STO 6					
06	15 02	(g)x²					
07	24 01	RCL 1					
08	14 07	(f) 1n					
09	23 05	STO 5					
10	15 02	(g) x²					
11	41	−					
12	14 02	(f) √x					
13	23 04	STO 4					
14	02	2					
15	71	÷					
16	14 06	(f) tan					
17	23 03	STO 3					
18	24 05	RCL 5					
19	61	×					
20	24 03	RCL 3					
21	24 06	RCL 6					
22	61	×					
23	24 04	RCL 4					
24	15 09	(g)→P					
25	21	x⇌y					
26	32	CHS					
27	74	R/S					
28	23 03	STO 3					
29	22	R↓					
30	71	÷					
31	23 02	STO 2					
32	74	R/S					
33	24 03	RCL 3					
34	21	x⇌y					
35	14 09	(f) →R					
36	23 04	STO 4					
37	74	R/S					
38	21	x⇌y					
39	23 05	STO 5					
40	13 00	GTO 00					
41							

Registers	
R_0	
R_1	Z_2/Z_1
R_2	r
R_3	θ
R_4	a
R_5	b
R_6	
R_7	θ

Step	Instructions	Input Data/Units	Keys				Output Data/Units
1	Store		STO	1			
			STO	Θ			
2	Calculate Θ		R/S				Θ
	r		R/S				r
	a		R/S				a
	b		R/S				b
3	Recall		RCL	2			r
			RCL	3			Θ
			RCL	4			a
			RCL	5			b
	r Θ = a + jb						

Table II
Exponential Taper Line Reflection Coefficient

Display Line	Display Code	Key Entry	X	Y	Z	T	Comments
00							
01	24 00	RCL 0					
02	01	1					
03	51	+					
04	23 00	STO 0					
05	15 73	(g) π					
06	61	×					
07	15 02	(g) x²					
08	24 02	RCL 2					
09	24 01	RCL 1					
10	71	÷					
11	14 07	(f) 1n					
12	02	2					
13	71	÷					
14	02	(g) x²					
15	51	+					
16	14 02	(f) √x					
17	23 06	STO 6					
18	74	R/S					
19	01	1					
20	08	8					
21	00	0					
22	61	×					
23	15 73	(g) π					
24	71	÷					
25	23 07	STO 7					
26	13 00	GTO 00					
27							

Registers	
R_0	n
R_1	
R_2	Z_2
R_3	
R_4	
R_5	
R_6	θ RAD
R_7	θ DEG

1	Store	Z_1	STO	1			
		Z_2	STO	2			
2	Calculate		R/S				θ_1 (RAD)
			R/S				θ_1 (DEG)
			R/S				θ_2 (RAD)
			R/S				θ_2 (DEG)
			R/S				θ_n (RAD)
			R/S				θ_n (DEG)
3	Recall		RCL	0			n
			RCL	6			θ_n (RAD)
			RCL	7			θ_n (DEG)

Table III
Exponential Taper Line Transformer Length

using the program shown in Table V. For the two section transformer discussed above, use n=2 in the Table IV program.

The question: what happens if you use more sections? comes up naturally. Fortunately Table IV program is of the iterative kind and will work with any number of equal length sections by making n equal to the number of line sections. As an example, an 8 section transformer (n=8) was calculated and the curves are shown in Fig. 5. Note that the matched operating point is still at a section length equal to quarter wavelength, thus making the transformer two wavelengths long. It should be noted that with n=8 the Table IV program calculation takes about 40 seconds per point, which is considerably less than with non-memory type calculators. In the recent past this type of a calculation could be reasonably done only with a computer.

Display Line	Display Code	Key Entry	X	Y	Z	T	Comments
00							
01	31	Enter ↑					
02	15 33	(g) RAD					
03	14 06	(f) tan					
04	23 00	STO 0					
05	24 06	RCL 6					
06	23 03	STO 3					
07	24 07	RCL 7					
08	14 02	(f) $\sqrt{x}$					
09	71	÷					
10	23 05	STO 5					
11	00	0					
12	23 04	STO 4					
13	24 00	RCL 0					
14	24 03	RCL 3					
15	61	x					
16	24 05	RCL 5					
17	24 00	RCL 0					
18	24 04	RCL 4					
19	61	x					
20	41	—					
21	15 09	(g) → P					
22	23 01	STO 1					
23	21	x ⇌ y					
24	23 02	STO 2					
25	24 00	RCL 0					
26	24 05	RCL 5					
27	61	x					
28	24 04	RCL 4					
29	51	+					
30	24 03	RCL 3					
31	15 09	(g) → P					
32	24 01	RCL 1					
33	71	÷					
34	24 05	RCL 5					
35	61	x					
36	21	x ⇌ y					
37	24 02	RCL 2					
38	41	—					
39	21	x ⇌ y					
40	14 09	(f) → R					
41	23 03	STO 3					
42	21	x ⇌ y					
43	23 04	STO 4					
44	24 07	RCL 7					
45	23 71 05	STO ÷ 7					
46	24 05	RCL 5					
47	01	1					
48	14 41	(f) x ≠ y					
49	13 13	GTO 13					

Registers	
R_0	($\tan\theta$)
R_1	
R_2	
R_3	R_{in}
R_4	X_{in}
R_5	(Z_n)
R_6	Z_0
R_7	$Z_L^{1/n}$

1	Enter θ (RAD)	θ					
2	Calculate		R/S				1
3	Recall		RCL	3			R_{in}
			RCL	4			X_{in}

Table IV — Multisection Line Input Impedence

Bandwidth Considerations

It is difficult to make a general statement as to which configuration offers larger bandwidth for a given maximum allowable reflection coefficient. Each case has to be treated individually, since the bandwidth is a complex function of impedance ratios, type of tranformer and maximum allowable reflection coefficient.

A typical problem may illustrate this better and show some of the basic trends. Let us consider the requirement of matching two real impedances with impedance ratios of 2, 10 and 100, using the discussed transformer configurations with a maximum allowable VSWR of 1.5:1. Using the relationship:

$$|\Gamma| = \frac{VSWR - 1}{VSWR + 1}$$

Display Line	Display Code	Key Entry	X	Y	Z	T	Comments
00							
01	31	Enter ↑					
02	23 01	STO 1					
03	74	R/S					
04	23 02	STO 2					
05	01	1					
06	31	Enter ↑					
07	24 01	RCL 1					
08	41	—					
09	15 09	(g) → P					
10	21	x ⇌ y					
11	22	R↓					
12	24 02	RCL 2					
13	24 01	RCL 1					
14	01	1					
15	51	+					
16	15 09	(g) → P					
17	21	x ⇌ y					
18	22	R↓					
19	71	÷					
20	13 00	GTO 00					
21							
22							
23							
24							

Registers
R_0
R_1
R_2
R_3
R_4
R_5
R_6
R_7

1	Enter	R_{in}	R/S				
		X_{in}					
2	Calculate		R/S				Γ

Table V — Calculation of Reflection Coefficient

Transformer Type	Z_2/Z_1 100	10	2
Quarter Wavelength	± 2.6%	± 9.4%	± 39.2%
Exponential	± 6.0%	± 16.7%	∞ (highpass)
Two Section	± 13%	± 25%	± 54%
Eight Section	± 53%	—	± 86%

Table VI — Transformer Bandwidth Comparison For VSWR = 1.5 maximum the maximum allowed reflection coefficient is 0.2. Using the curves for the quarter wavelength matching transformer (Fig. 2) we can draw a line at $|\Gamma| = 0.2$. We can see that for $Z_2/Z_1 = 2$, this requirement will be met when the line length is between 0.304 π and 0.696 π (0.5 π= λ/4). Thus the bandwidth is $\pm$39.2 percent. Similarly, the bandwidths can be calculated for the other values of Z_2/Z_1 and other transformer configurations using Fig. 3, 4, and 5. The calculated values of bandwidth for a VSWR of 1.5:1 are shown in Table VI.

It should be noted that the first line resonance is used in these calculations. The higher order resonances will give narrower bandwidths. For instance the 3/4λ line transformer will have ⅓ the calculated bandwidths shown in Table VI.

The exponential taper transformer has very large bandwidths for low Z_2/Z_1 ratios. For the lowest reflection coefficient at moderate bandwidths the multisection λ/4 transformer offers better performance for the same overall transformer length. The optimum performance (VSWR of 1:1), of course, is theoretically obtained with an infinite number of λ/4 sections.

Thus, the designer has some choice in selecting the transformer configuration most suitable for his particular requirements.

Transformer Length

The λ/4 transformer is the shortest, providing its bandwidth is adequate. Arnold et al. states that the tapered line transformer may be very attractive for narrowband complex impedance matching where space is important. They show an example of matching a 50 ohm line to a transistor with an input impedance of 10+j15 ohms. A 5:1 tapered line with an electrical length of 60° is required. However, the same matching can be accomplished with a 14.79 ohm nontapered line with an electrical length of 38.27°. Thus the tapered line transformer is always longer than the nontapered. However, it may be very useful where its highpass characteristics can be utilized. □

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Acknowledgement

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Wilkinson Hybrid With Quarter-Wave Offset

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When deciding on methods of combining the output power from two transistor amplifiers at UHF and microwave frequencies using microstrip, a choice is usually made between the Wilkinson hybrid (1, 2) and the 90° Branch Line hybrid (3). There exists another possible combiner, the Wilkinson hybrid with a quarter-wave offset (4), which has some of the characteristics of each as a compromise hybrid. The Offset Wilkinson hybrid displays the wide bandwidth and isolation inherent with the Wilkinson hybrid while at the same time offering the advantages of improved summing port return loss and intermodulation performance available in the Branch Line hybrid. It does however suffer from unequal insertion loss in each side port when used over a wide bandwidth, in a similar manner to the 90° Branch Line hybrid.

Need of Combiners/Splitters

Power combining may be considered on one of two general levels; the device level and the circuit level. Device level combining is accomplished by clustering several devices in a region whose extent is small compared with a wave length. Transistor vendors are continually increasing the available output powers from a single package. Combining several transistor dies within the same package eventually runs into problems of impedance matching, concentrated heat dissipation and reactance interaction or power hogging. By directly paralleling two transistors, the resistive part of the input and the output impedance goes down yielding a higher Q, lower bandwidth part. The advantage of using a hybrid for power combining centers around the isolation achieved

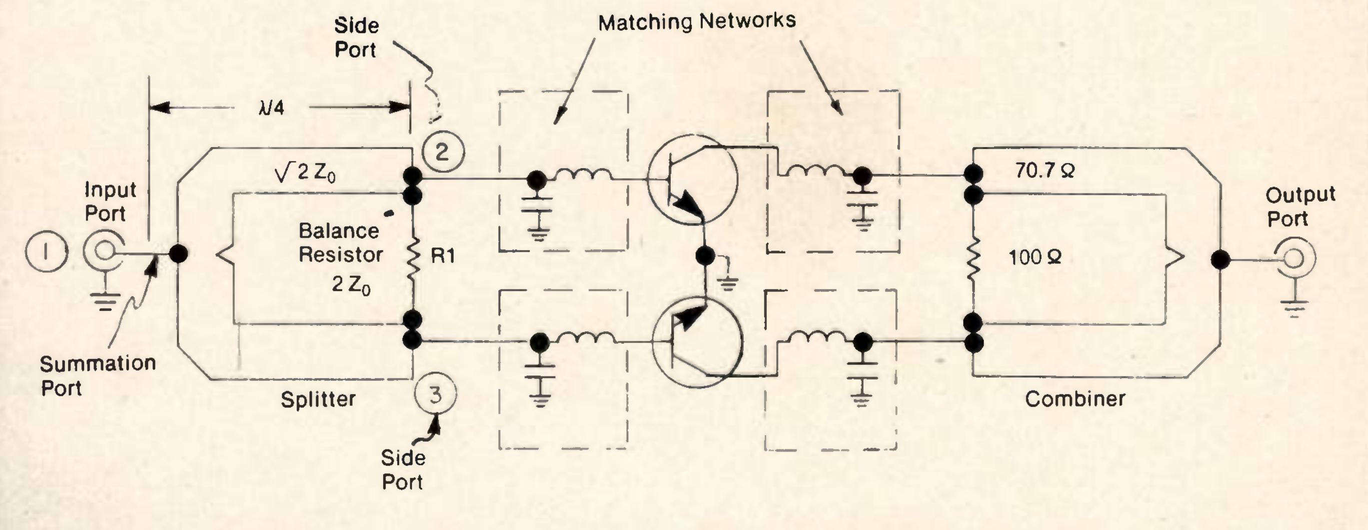


Figure 1. The Wilkinson splitter/combiner is composed of two quarter-wave transmission lines and a balance resistor.

By adding a quarter-wave line to one output port of a Wilkinson hybrid one can expect an improvement in input return loss and output intermodulation performance. If the bandwidth is limited to approximately 10 percent, the power unbalance due to unequal insertion losses in each leg should not be a problem.

between the input ports. Stability of both transistors is enhanced when the reactive components of one transistor do not affect the parallel transistor.

Wilkinson

Power entering the summing port of a Wilkinson hybrid emerges from both side ports with equal amplitude and phase. The hybrid is reciprocal as shown in Fig. 1 where it serves both as a splitter and as a combiner. The Wilkinson hybrid is constructed using two identical quarter-wave lines connected between a summation port and two side ports. Each quarter-wave matching transmission line has a charactersitic impedance, Z_O, equal to the square root of the product of the source and load impedances.

$$Z_O = \sqrt{Z_S \cdot Z_L}$$

Where Z_S equals source impedance and Z_L equals the load impedance. For the case of a 50 ohm input/50 ohm output hybrid, we need to transform 100 ohms at the source end of the quarter-wave line down to 50 ohm at the side ports. By connecting the 100 ohm end of two quarter-wave lines in parallel at the summation port the 50 ohm input impedance is achieved.

One notes that the side arms (labelled 2 and 3) are one-half wave apart. If a reflection or mismatch occurs at one of the side ports, port 2 for instance, the reflected signal splits with part of the energy travelling back to the summation port 1. There it splits again and continues to the opposite side port 3. This reflected wave arrives at port 3 from two paths. The path through the $\lambda/4$ lines amounts to 180° of travel. This signal will then be cancelled by the 0° signal which travelled directly through the balancing resistor. The value of the resistor is chosen so that the two signal appearing at port 3 both suffer 6dB loss and thus are equal in amplitude and 180° out-of-phase and will be nulled out to achieve a high value

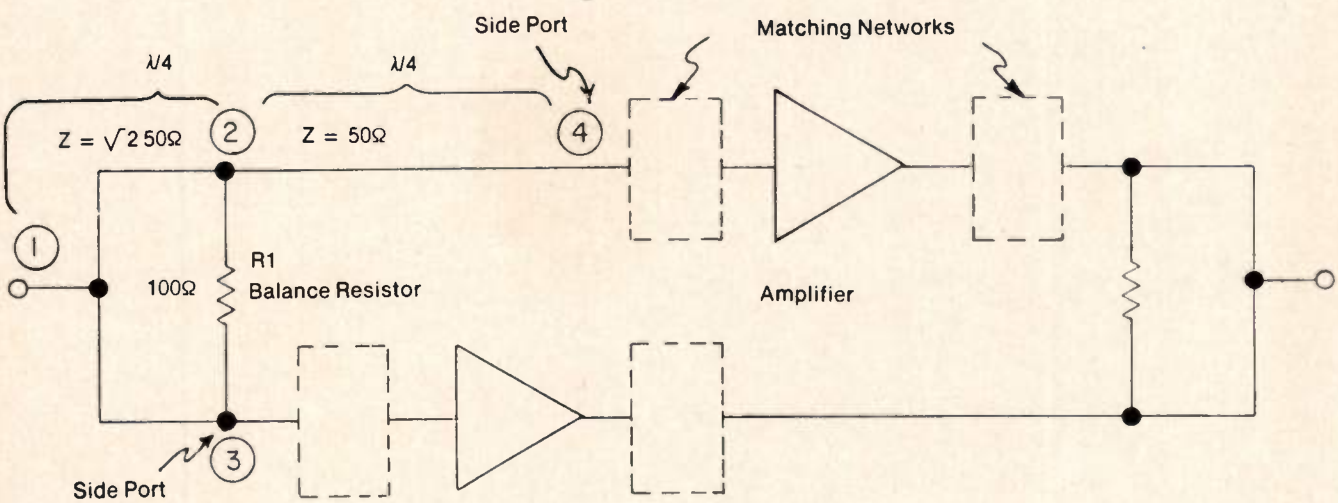


Figure 2. A quarter-wave offset section is added to one arm of the standard Wilkinson hybrid to improve the input return loss.

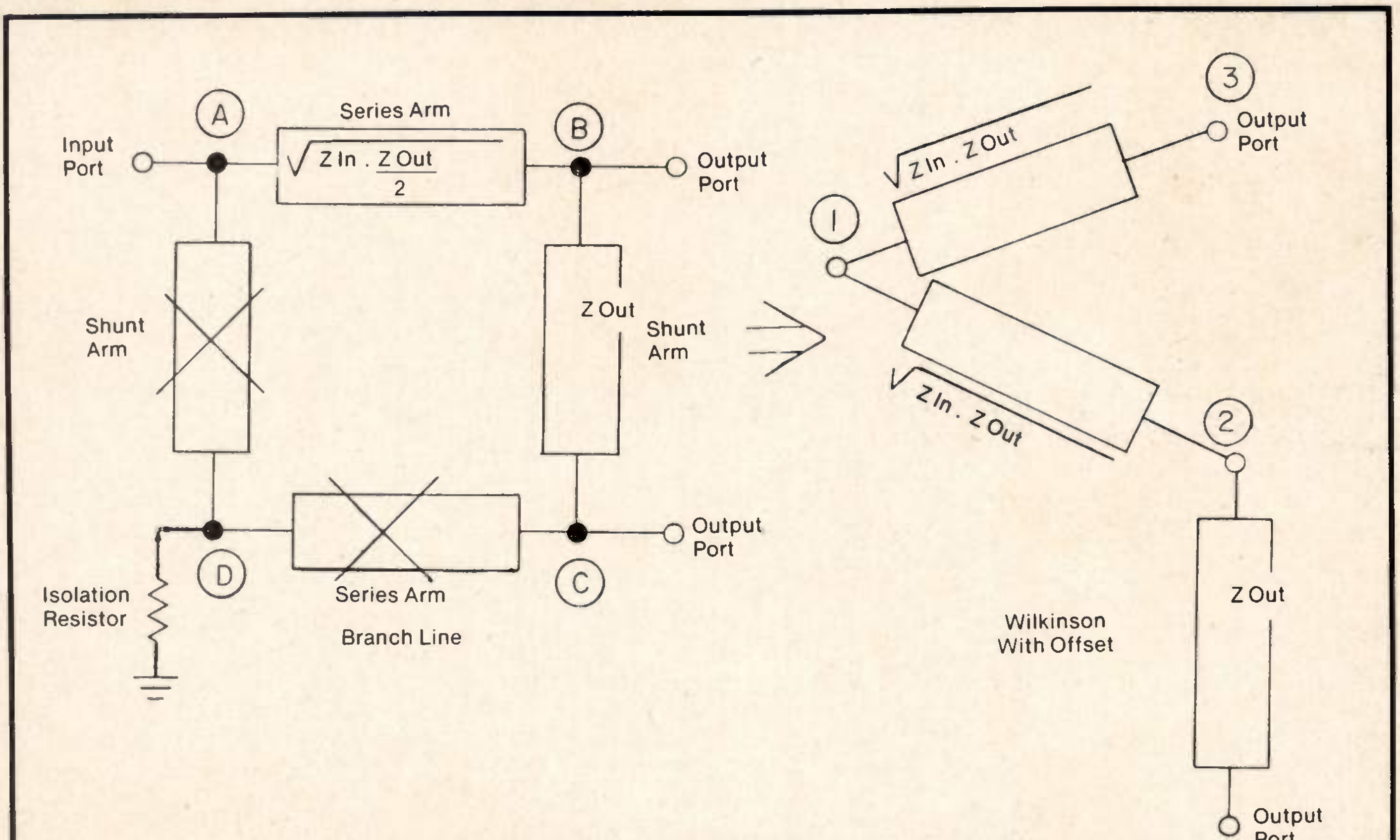


Figure 3. The Wilkinson with offset may be thought of as one-half of a Branch Line hybrid.

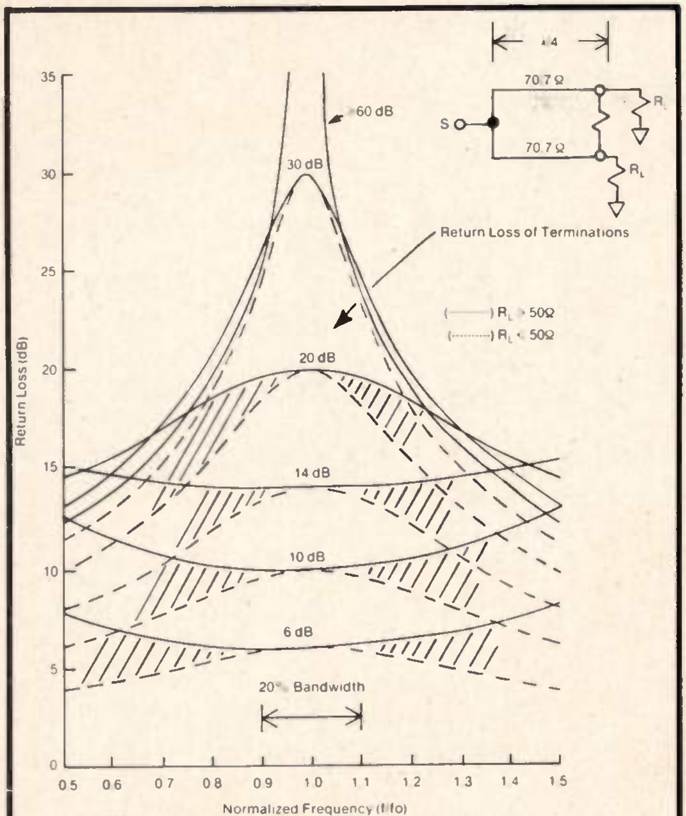


Figure 4. The Wilkinson provides an input return loss approximately equal to the terminations over a 20 percent bandwidth.

of isolation. Reflections at one side port do not appear at the other side port.

Wilkinson With 90° Offset

By adding an additional quarter-wave line in series with one of the output ports of a standard Wilkinson hybrid, the Offset Wilkinson is formed, Fig. 2. The input return loss of the Offset Wilkinson splitter at the summation port will be better than the standard Wilkinson hybrid for two reasons. Any reflections from a mismatch at port 4 appears back at port 2 180° out-of-phase with the incident wave which caused it. Thus the reflected signals from equal mismatches present at ports 3 and 4 will be 180° out-of-phase. The reflected signal will be partially dissipated in the balance resistor.

The second reason for improved return loss is due to cancellation of reflected signals at the summation port. The round trip distance of the incident and reflected signal in the offset path 1-4 is one-half wave greater than path 1-3. This 180° relationship will cause cancellation of the reflected signal.

Derivation

The Wilkinson hybrid with offset may be thought of as a combination of the Wilkinson and the Branch Line hybrids. The series arm C-D and the shunt arm A-D of the Branch Line hybrid have been eliminated, Fig. 3, to form the Wilkinson hybrid with an offset quarter-wave line. The series arm A-B of the Branch Line hybrid may be thought of as a two $\lambda/4$ lines in parallel. For a 50 ohm system the impedance of the series quarter-wave matching section in the Branch Line hybrid should be equal to the geometric mean of the product of the input impedance and the impedance at point B. The 25 ohm impedance at B is formed by the parallel combination of the 50 ohm termination at port B and the 50 ohm shunt arm B-C. The Wilkinson hybrid with offset is formed if the Branch Line series arm A-B is split into two 70.7 ohm lines with one arm containing the 50 ohm $\lambda/4$ offset.

Input Return Loss

One of the principal advantages of adding the quarter-wave offset to the Wilkinson combiner is the improved input return loss at the summation port. The sum port return loss of a Wilkinson hybrid without offset is shown in Fig. 4 for various equal mismatched terminations at the side ports. We have shown the region of input return loss bounded for all phase angles. The terminations were chosen to be equal because most transistor amplifiers will be built with transistors from the same manufacturered batch. If each side port is terminated by equal mismatched loads, the input return loss at center frequency is the same as the load mismatch and generally degrades for operating frequencies away from center frequency.

Over a 20 percent bandwidth one might expect the individual transistors in a typical amplifier to maintain at least a 10 dB input return loss. Thus when they are combined using a Wilkinson hybrid, the input return loss over the same bandwidth will also be about 10 dB (VSWR $\leqslant$1.9:1). The input return loss when using the Wilkinson hybrid with an offset however, Fig. 5, is greater than 20 dB (VSWR $\leq$1.2:1) over the same 20 percent bandwidth.

The improvement in stability of the driver due to the increased input return loss of the hybrid with an offset is very important. If the power level of an amplifier is changed due to variations in temprature, supply voltage, output load, or power turndown, the transistor amplifier input impedance will also change. This change in amplifier input impedance must not be allowed to affect the stability or match of the driver stage.

A UHF 500 MHz hybrid model was constructed using semi-rigid coaxial lines to verify the theoretical predications of input return loss, Fig. 6. With the side ports terminated with 50 ohm loads, the Wilkinson hybrid performed equally well with or without the quarter-wave offset (curve A). If both side ports are terminated with 6 dB return loss loads, the Wilkinson shows a fairly flat 6 dB input return loss across the band (curve B) while the offset hybrid maintains greater than 15 dB return loss (VSWR $\leq$ 1.43:1) over a 20 percent bandwidth, assuming the transistors exhibit at least a 10 dB input return loss.

The coupling loss of the Wilkinson hybrid with the 90° offset however is equal in each arm only within 10 per cent of the center frequency, Fig. 8, when both arms are identically mismatched. If both terminations are greater than 50 ohms, the insertion loss in the offset arm decreased when the operating frequency is different from the design center frequency while the loss in the other arm increased. This differential insertion loss labelled "Δ" in Fig. 8, means that the transistor attached to one arm is driven harder than the other transistor. This difference in power will be magnified at the outputs of the transistors by the gain of the transistors. The differential output power will then appear across the balance resistor. Over the same 20 percent bandwidth, assuming equal 10 dB input return losses to the tranisitors and equal gain, an input power difference of 0.1 dB would appear across the output arms of the hybrid. If one were combining the output power from two 50 watt transistors to form a 100 watt power amplifier, there would be a 5

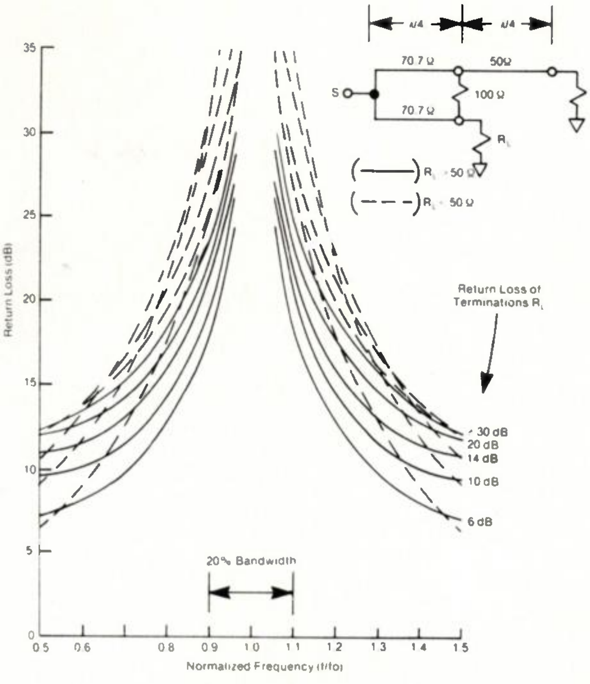


Figure 5. The addition of a quarter-wave offset improves the input return loss to greater than 20 dB over a 20 percent bandwidth for a 10 dB return loss load.

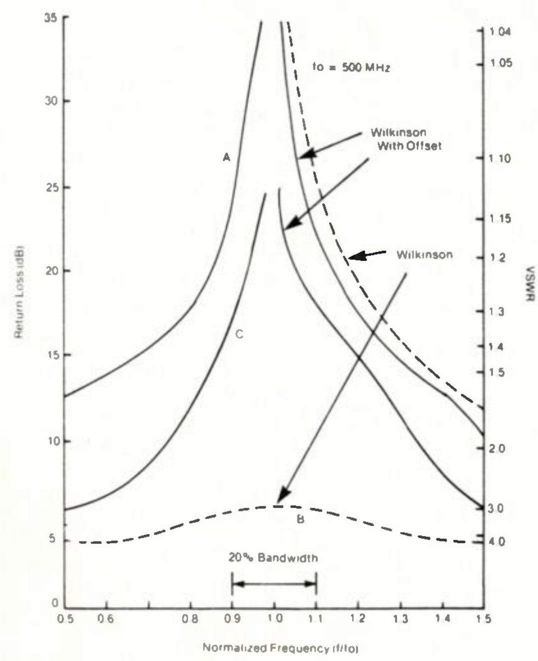


Figure 6. The experimental model matches the theoretical results of Figure 5 quite well; (A) matched terminations, (B + C) 6 dB return loss termination, R_L = 26 ohms.

watt difference at the individual transistors. The balance resistor would absorb 2.5 watts of wasted power. This would exceed the capability of an inexpensive 2 watt carbon composition balance resistor. However, if the operating bandwidth is reduced to 10 percent then the above problem vanishes.

Catastrophic Failure

As the output power level required from a solid-state amplifier is increased, it becomes necessary to achieve this power by combining the output power from identical modules using hybrids. If one module should catastrophically fail and present an open or possibly a short at the input or output terminal, it is desirable that no other modules are adversely affected and that the imredance to the driver must not induce catastrophic failure. Isolation between output ports is identical for the Wilkinson hybrid with and without the offset. The isolation is greater than 19 dB over a 40 percent bandwidth, Fig. 9, even when the load resistors present a 6 dB return loss.

The input return loss presented to the driver for the Wilkinson hybrid with or without the offset arm is shown in Fig. 10. If both arms are shorted or if both are opened, the input return loss of the standard Wilkinson hybrid is 0 dB (curve A). By adding the quarter-wave offset arm, the return loss at the sum port is improved to greater than 15 dB over a 20 percent bandwidth when both ports are either open or shorted (curve B & C). The driver stage would barley notice that the final stage had been completely wiped out. If, however, only one transistor is wiped out, the input return loss for either hybrid is reduced to approximately 6 dB over a 20 percent bandwidth (curves D, E, F, G). Thus over a 40 percent

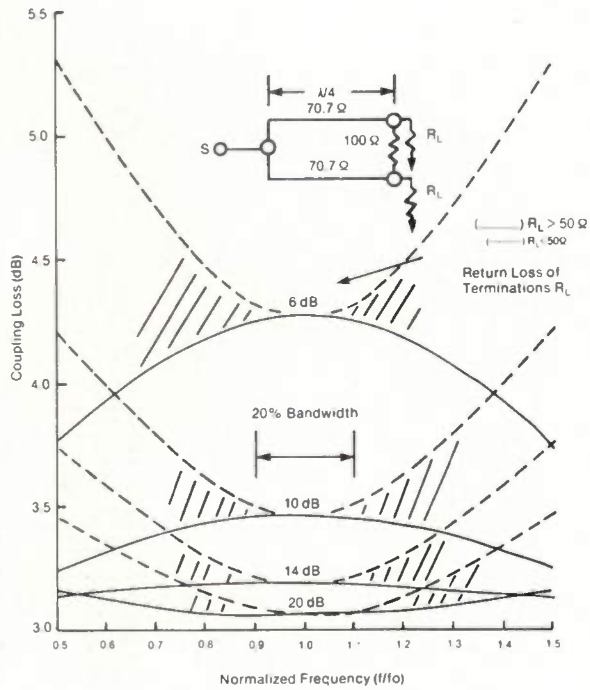


Figure 7. The coupling loss for a standard Wilkinson is equal in each arm for identical terminations.

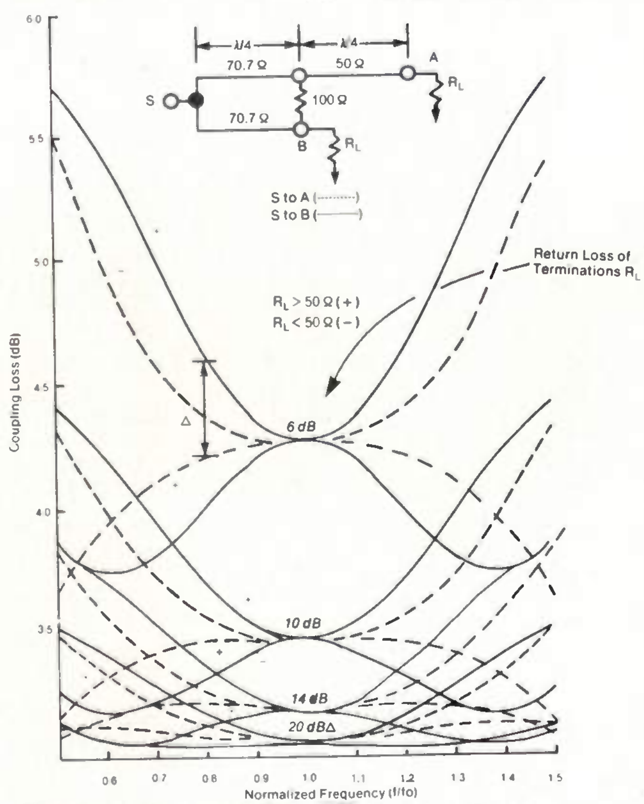


Figure 8. The coupling loss for an offset Wilkinson is different in each arm. This difference is greater as the mismatch is increased.

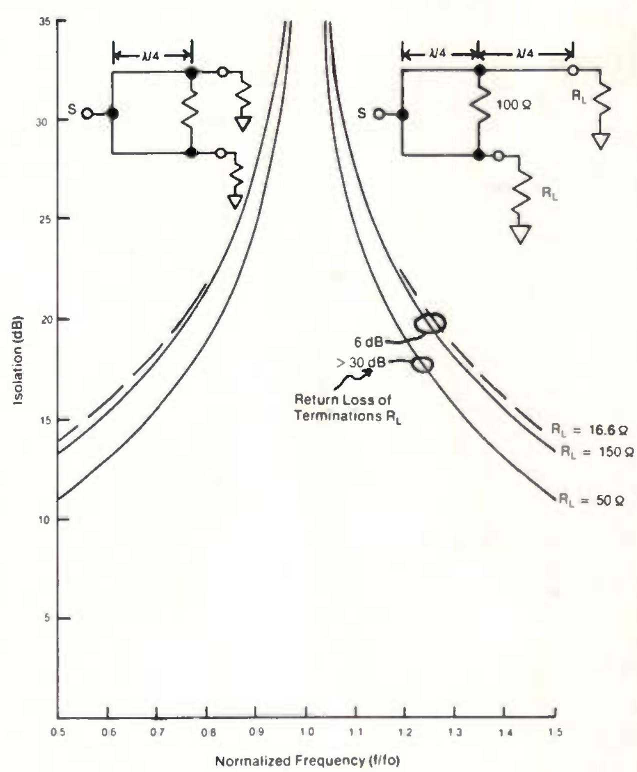


Figure 9. The isolation between side ports is equal for the Wilkinson with and without the offset line.

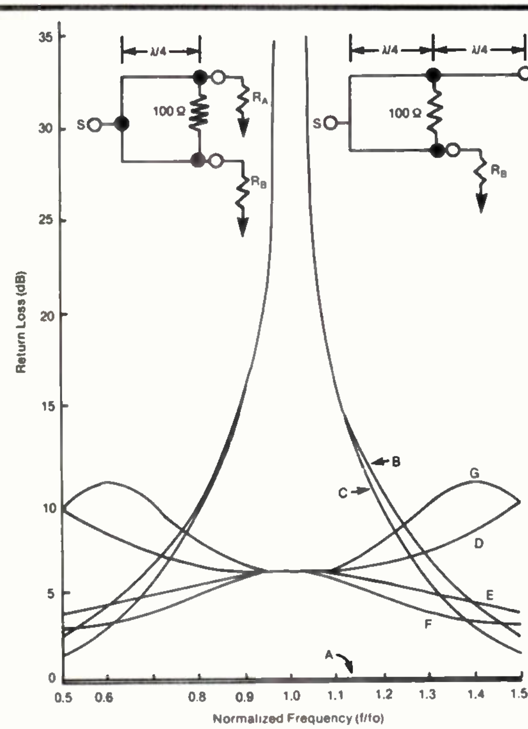
bandwidth the input return loss, no matter what happens to either or both transistors, will always be greater than 5 dB (VSWR ⩽ 3.5:1) for the Offset Wilkinson. The return loss of the sum port for the Offset Wilkinson hybrid is identical to the Wilkinson hybrid without offset if both arms are terminated in their characteristic impedance. This impedance is merely shifted one-quarter wave length and effectively appears directly at the balance resistor.

Intermodulation Performance

The final amplifier of a mobile radio is operated in a power-efficient, but highly non-linear, class-C mode. Transmitter intermodulation occurs when the signal, F2, from a nearby transmitter antenna is coupled through the transmitter antenna back to the collectors of the final amplifier. This interference signal mixes with the second harmonic of the main signal, 2XF1, to form mixing or intermod products which are then re-radiated from the same antenna. This third order product appears at 2F1-F2, Fig. 11.

An isolator (Ferrite circulator plus termination) is normally installed in a base station as a one-way valve to reduce the interfering signal F2 of neighboring transmitters from reaching the collector circuitry.

By adding a quarter-wave section to the output Wilkinson combiner, the beat products from each collector will be summed at the output 180° out-of-phase with each other. This is due to the quarter-wave delay of the incoming interference signal in the Wilkinson combiner offset arm added to the quarter-wave delay of the intermod product travelling back through the same offset. Thus the intermod product will be cancelled out at the summing port due to this half-wave phase difference of



Wilkinson: RA = RB = 0 or ∞ ohms; Wilkinson plus offset: (B) RA = RB = ∞ ohms, (C) RA = RB = 0 ohms; Wilkinson: (D) RA = 50 ohms, RB = ∞ ohms, RB = 50 ohms, (E) RA = 50 ohms, RB = 0 ohms or RA = 0 ohms, RB = 50 ohms; Wilkinson plus offset: (D) RA = 50 ohms, RB = ∞ ohms; (F) RA = ∞ ohms, RB = 50 ohms; (E) RA = 50 ohms, RB = 0 ohms; (G) RA = 0 ohms, RB = 50 ohms.

Figure 10.

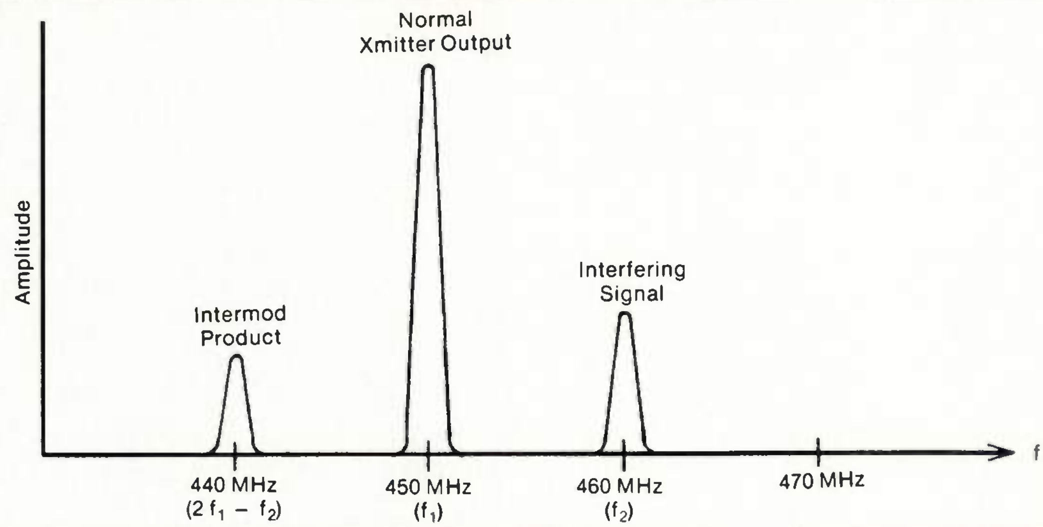


Figure 11. A nearby interfering transmitter signal F2 mixes in the non-linear collector circuit with the second harmonic 2F1 to form an intermodulation product.

the intermod products from each collector summed at the output port.

Two UHF amplifiers were constructed to demonstrate this intermod improvement when using the Wilkinson plus quarter-wave offset, Fig. 12. The intermodulation performance showed a 5 to 23 dB improvement. As expected, the shape of the curve is similar to that of Fig. 5, Input Return Loss.

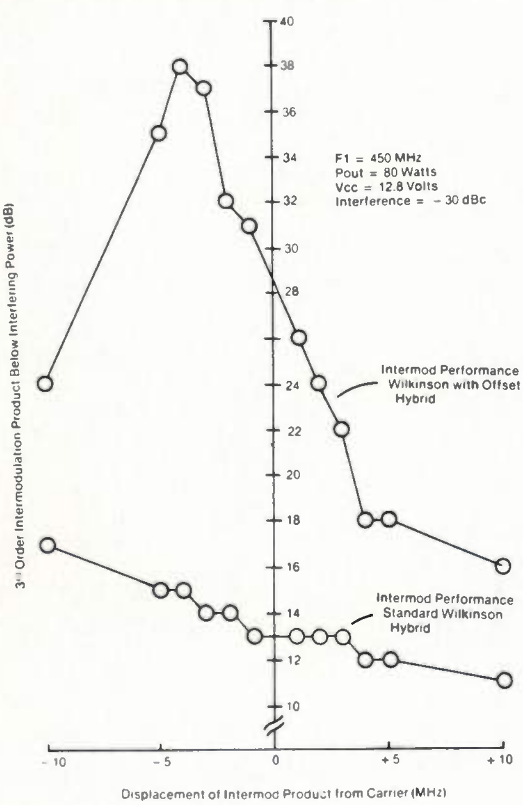


Figure 12. The intermodulation performance of an amplifier is improved when the 90° offset arm is added to a Wilkinson combiner.

Microstrip Combiner Resistors

The power-handling capability of a Wilkinson power splitter/combiner with or without offset is limited by the balance resistor, R1, Fig. 2 If both arms of a Wilkinson splitter are matched or identically mismatched, no power is dissipated in the balance resistor. Any unbalance appears across the balance resistor. If one side port of a Wilkinson combiner is completely lost (shorted or open transistor), then one-half of the output power is dissipated in the balance resistor.

The designer has a choice of available combiner resistors. If cost is the main criterion, the familiar 1 and 2 watt carbon composition resistors may be used. As individual transistor output power levels are increased above 50 watts these resistors cannot handle the power unbalance due to differences between transistor phase and gain imbalance. Above this power level combiner resistors which heat sink to the chassis are needed.

Thin-film combiner resistors are available from many sources in microstrip configurations which gracefully mate with the Wilkinson hybrid in microstrip form. These power resistors are formed by depositing nickel-chromium on a berullia-oxide ceramic substrate and then hard brazing onto a standard copper flange or stud. This is very similar to the emitter ballast resistors used in RF power transistors. □

References

1.) E.J. Wilkinson, "An N-Way Hybrid Power Divider", *IRE Trans. On Microwave Theory and Techniques*, vol. MTT-8, January 1960, pp. 116-118.
2.) S.B. Cohn, "A Class of Broadband Three-Port TEM-Mode Hybrids", *IEEE Trans. On Microwave Theory and Techniques*, vol. MTT-16, February 1968, pp. 110-116.
3.) J. Reed and G. J. Wheeler, " A Method of Analysis of Symmetrical Four-Port Networks", *IRE Trans. On Microwave Theory and Techniques*, vol. MTT-4, October 1956, pp. 246-252.
4.) W. C. Tsai and D. M. Wong, "Integrated Solid State Transmitter Uses Dual-Gate and Power Field Effect Transistors", *Microwave System News*, February 1981, pp. 72-76.

Microprocessor-Based Microwave Signal Generators

Solid-state technology combines with spectrally-pure cavity tuning to provide high performance test signals from Hewlett-Packard's newest HP 8683A/B and HP 8684A/B Microwave Signal Generators. Instead of klystrons, the active oscillator elements are bipolar transistors for the HP 8683 and GaAs FET's for the HP 8684.

Communications applications are covered in two frequency bands: 2.3 - 6.5 GHz (HP 8683A), and 5.4 - 12.5 GHz (HP 8684A). These models feature 0 to −130 dBm power output, calibrated, low distortion FM with ±5 MHz peak deviation at rates from DC to 10 MHz, and calibrated, metered AM, 0 - 70 percent modulation depths at rates from DC to 10 kHz.

Radar and military electronics applications use the "B" Models, HP 8683B (2.3 - 6.5 GHz) and HP 8684B (5.4 - 12.5 GHz). In addition to the above capabilities, these models have increased power output of 10 dBm and a high performance pulse modulator and generator with rates from 10 Hz to 1 MHz in 5 ranges; pulse widths from 100 ns to 100 ms in 7 ranges, and pulse delays from less than 50 ns to 100 ms in 7 ranges. The RF pulse itself provides greater than 80 dB on-off ratio and less than 10 ns rise and fall times.

These two frequency ranges now cover applications bands which previously required four older HP klystron models. They are compact and highly portable at less than 40 pounds each.

Price (U.S.):

HP 8683A, 8684A (Communications Version)	$12,000
HP 8683B, 8684B (Radar Version)	$15,000

Delivery is 12 weeks after receipt of order. Circle INFO/CARD #140.

TEM Cells

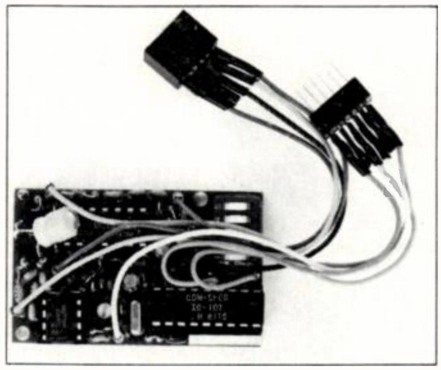
Instruments For Industry, Inc. (IFI), a leading designer and manufacturer of broadband amplifiers and test equipment, now offers its new brochure covering its line of TEM cells for electromagnetic testing. Included in the brochure is the CC-series of TEM cells. The brochure describes the various features, applications and specifications for each of these products.

For additional information on these products, or to obtain this brochure, contact: Ronald Richards, Instruments For Industry, Inc., 151 Toledo Street, Farmingdale, New York 11735; (516) 694-1414. Circle INFO/CARD #139.

CTCSS Encoder-Decoder

Communications Specialists, Inc. introduces their new TSS-32 direct plug-in CTCSS encoder-decoder. The TSS-32 will directly replace the TN-14 tone board which is used in the Standard 766L. It may also be used as a direct plug-in for other Standard radios which take the TN-10 and TN-14 factory tone boards. This plug-in board is based on the field proven programmable TS-32 and will greatly reduce installation time. The price is $64.50. A catalog is availabel on request.

Communications Specialists, Inc.,



426 West Taft Ave., Orange, CA 92667; (800) 854-0547, (714) 998-3021. Circle INFO/CARD #137.

UHF Surface Acoustic Wave Resonator

RF Monolithics, Inc. of Dallas, Texas, announces the availability of new surface acoustic wave resonators operating in the UHF range.

The SAW resonator is a frequency control element for VHF-UHF oscillators in applications such as TV, CATV, mobile communications, STV and others, with frequencies of 674 MHz and 680 MHz available in large quantities.

The device offers crystal stability directly at UHF frequencies, eliminating crystal multiplier chains, while occupying less space and consuming less power than other methods of frequency control.

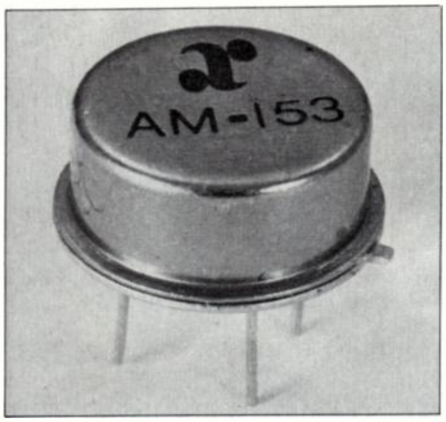
The company has announced receipt of a contract to supply Scientific-Atlanta, Inc. with SAW resonators for use in the Scientific-Atlanta Series 6700 Set-Top Terminal. This is stated to be the first large volume application of SAW resonators in the CATV industry.

Samples are now immediately available and quantity quotations are available upon request.

For further information contact RF Monolithics, Inc., 4441 Sigma Road, Dallas, TX 75234; (214) 233-2903. Circle INFO/CARD #136.

"Lossless Feedback" Amplifier

Using a torriodial coupler as the major feedback element, Anzac's AM-153 can provide significantly lower noise figure performance then



is currently available in TO-8 amplifiers. Typical noise figure performance is shown in Figure 1 with a maximum guaranteed 3.5 dB over temperature. Along with this noise figure performance, the lossless feedback provides improved third order intercept of over +15 dBm with only 200 mW of DC power applied.

The AM-153 is a thick film amplifier designed to meet the rugged environments of today's military systems. These units, as well as all Anzac standard products, are available from stock at $135 each in 1-5 quantities and can be supplied screened to MIL-STD-883B.

Anzac Division, Adams-Russell Company, Inc., 80 Cambridge Street, Burlington, MA 01803; (617) 273-3333. INFO/CARD #135.

Susceptibility Testing Equipment

Instruments For Industry, Inc. (IFI), a leading designer and manufacturer of broadband amplifiers and test equipment is now offering its new brochure covering its line of E-Field sensors. Included in the brochure are IFI's Models EFS-1, EFS-2, EFS-3, and remote fiber optic readout components. The brochure describes the various features, applications and specifications for each of these products.

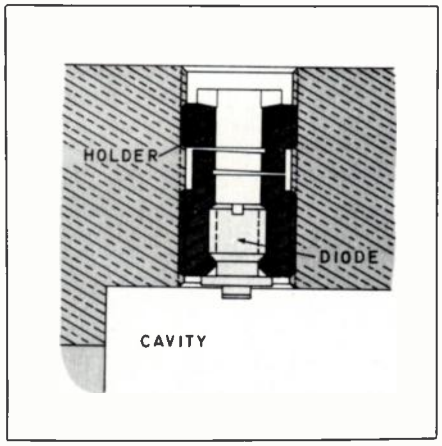
For additional information on these products, or to obtain this brochure, contact: Instruments For Industry, Inc., 151 Toledo Street, Farmingdale, New York 11735; (516) 694-1414. Circle INFO/CARD #138.

Diode Holders For Microwave Applications

Johanson microwave diode holders are an inexpensive and efficient means of securing and tuning microwave diodes into precise attitudes for optimum circuit performance.

The essence of the diode holder design is embodied in the patented Johanson slotted rotor screw which permits adjustment under dynamic operating conditions without locknuts or set screws.

The self-locking constant torque drive mechanism of the rotor screw maintains very low dynamic tuning noise and high Q throughout the full excursion of the diode holder travel. Benefits to the user include: relaxation of costly mechanical tolerances in circuit fabrication; one-hand tuning in cramped quarters; ultra-precise adjustment resolution; excellent diode-to-holder thermal conductivity.



Typical applications include impatt diode transmitters, high-power waveguide limiters and switches, Gunn oscillators, detectors and mixers.

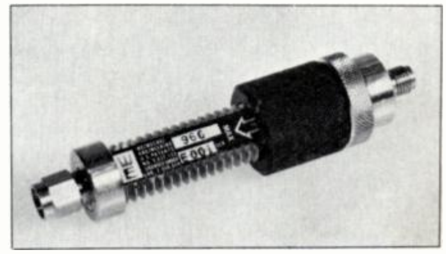
Johanson Manufacturing Corporation, 400 Rockaway Valley Road; Boonton, NJ 07005; (201) 334-2676. Please circle INFO/CARD #133.

Continuously Variable Lossy-Line Attenuators

Weinschel Engineering Variable Lossy-Line Attenuators are designed for signal level control in the 2.5 to 18 GHz frequency range. The Series 960, available in 12 models, is ideal for level setting and channel balancing where flatness of attenuation and minimum phase shift vs. attenuation are of prime consideration.

The uncalibrated 960 and calibrated 961 are thumb wheel controlled and are ideally suited for bread board applications. The 962 is designed for systems applications requiring low RF leakage and are varied by way of a screw adjust while the 963 is especially suited for measurement system level setting due to its fast movement slider control. Each model is available in 3, 6 or 10 dB values.

Miniature in size for performance



these attenuators are designed for low linear phase shift: 1°/dB × f(GHz); low insertion loss: less than 1.0 dB; low frequency sensitivity: less than 1.0 dB together with low power and temperature sensitivity.

For additonal information contact Weinschel Engineering, One Weinschel Lane, Gaithersburg, Maryland 20877; (301) 948-3434. Please circle INFO/CARD #134.

Log-Periodic Antennas

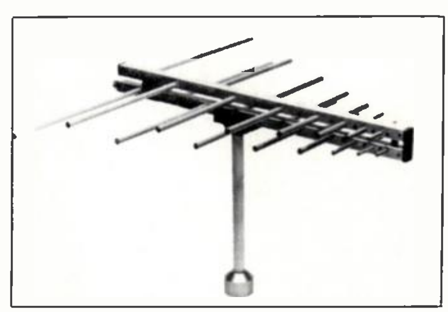
The Electro-Mechanics Company of Austin, Texas, is introducing their new 3140 Log-Periodic Antennas for applications in compliance and susceptability EMC testing.

Designed for specific EMC testing frequency coverages, the Models 3145 and 3146 Log-Periodic Antennas are lightweight and compact enough for use in applications where space is a vital consideration. Such applications would include testing in anechoic chambers and shielded enclosures.

These antennas are fully calibrated to include gain and antenna factors which comply with testing procedures involved in FCC Part 15. All antennas have a gain greater than 5 dB, are linearly polarized and have the relatively constant input impedance characteristic of Log-Periodics.

When used for susceptability testing, the Series 3140s can handle up to 1000 Watts of input power which create fields of over 200 volts per meter.

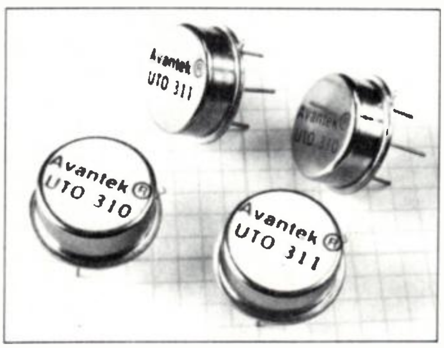
The 3145 has a frequency range of 100-1000 MHz with a maximum width of 1.5 meters (59"). The Model 3146 covers the frequency range of 200-1000 MHz and has a maximum width of 75 cm (29½").



Contact Mike Hart, The Electro-Mechanics Company, P.O. Box 1546, Austin, Texas 78767. Please circle INFO/CARD #131.

TO-8 Amplifier Modules

Avantek, Inc., Santa Clara, CA is now producing two TO-8 packaged, 10-300 MHz thin-film hybrid amplifier modules that combine guaranteed noise figures as low as 2.0 dB, with up to +17 dBm output power, +31 dB third-order intercept point (+40 dBm, second-order) and 8 dB gain; yet consume only 15 to 25 mA at +15 VDC. This combination of high output power, low noise figure and wide dynamic range makes these modules suitable for use in the IF amplifiers of superheterodyne receivers to complement the wide dynamic range of today's high-performance mixers, or as VHF preamplifiers. Their low current consumption makes them ideal for battery-operated equipment.

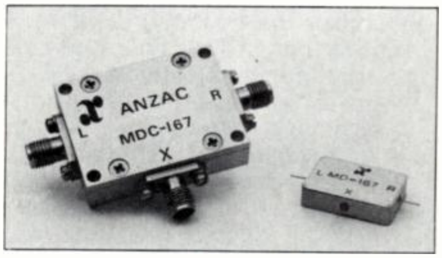


One unit, designated UTO-311, offers a typical (at 25°C) 2.0 dB noise figure, +20 dBm output power (1 dB gain compression), +31 dBm third-order intercept point and 8.5 dB gain. The second, designated UTO-310 offers a typical 1.5 dB noise figure, +14 dBm output power (10-200 MHz, +12 dBm full-band) and 9 dB gain. Both units feature ±0.5 dB typical gain flatness, 1.3:1 typical input and output VSWR and guaranteed performance over the −55° to +85°C temperature range.

The UTO-310 is priced at $130 and the UTO-311 at $160 in 1-9 piece quantities, with delivery 30 days ARO. Contact Harvey Huffman at Avantek. INFO/CARD #130.

EW Mixer Covers 0.5 - 18 GHz

Anzac Division of Adams-Russell Company has developed a double-balanced mixer for the EW market. The MD-167 provides RF port coverage of 0.5 - 18 GHz with overlapping



IF coverage to 8.0 GHz. The mixer features 8.0 dB typical, 10.0 dB max conversion loss at any RF frequency with IF's to 8 GHz. The mixer also provides 20 dB typical isolation and 1 dB compression point of _9.5 dBm. The MD-167, like all of Anzac's Microwave Mixer line, is designed for military environments and comes in a hermetic drop in module or connectorized housing. Pricing starts at $565 with delivery from stock.

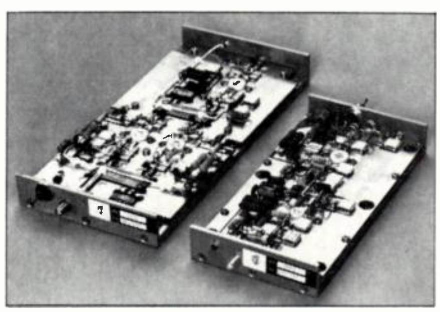
Anzac Division, Adams-Russell, 80 Cambridge Street, Burlington, MA 01803; (617) 273-3333. Please circle INFO/CARD #132.

100 Watt VHF And UHF OEM Wideband Amplfiers

RF Power Labs' Models V100 and U100 wideband amplifiers will permit operation on 100-160 MHz and 225-400 MHz in 100 Watt increments. The units require 28 VDC and 10 Amps maximum and 2-4 Watts drive for Class AB operation. Harmonics are minus 20 dB.

The V100 and U100 may be combined in pairs or quads to achieve a compact, high power wideband system.

Each module is constructed of highly reliable, multi-source components, which are carefully tested and mounted on a quarter inch of copper.



1-3 pc. at 100-160 MHz is $695 ea., or at 225-400 MHz is $1295 ea.

RF Power Labs, Inc., 21820 - 87th S.E. Maltby Industrial Village, Woodinville, WA 98072; (206) 481-8833. Circle INFO/CARD #129.

VHF IC

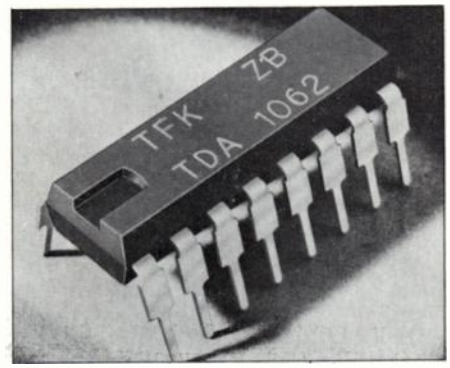
Samples of the TDA1062S IC — the successor of Telefunken's recently introduced integrated VHF Front End IC — are presently available from AEG-Telefunken's Semiconductor Division.

The IC remains the only commercially available monolithically integrated VHF Front End capable of operating up to 250 MHz (min.), with an integrated RF amplifier, local oscillator, double balanced mixer and voltage regulating circuitry on one chip.

In addition, the TDA1062S features an on-Chip local oscillator buffer, with either fixed or mask programmable gain. Thus, the device becomes particularly suitable for applications in PLL tuned commercial and communications receivers, where a high local oscillator output level is often required for interfacing with prescaling stages. A typical L.O. buffer output level is 150mV.

Similarly to its predecessor, the TDA1062S' operating voltage range is 9...15 V, with a typical current consumption of 30 mA. The typical power gain is 30 dB.

Applications include high performance FM tuners, scanners, communications receivers, marine and Weather Band receivers, UHF



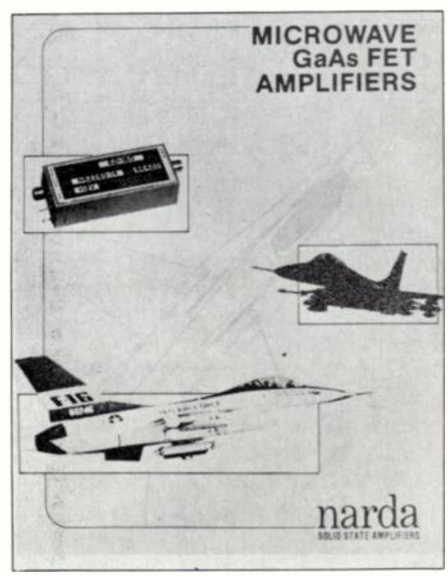
down-converters and fast interface to fiber optic communications links.

The IC, made in Telefunken's fast bipolar technology is housed in a standard 16-Pin DIP package and is priced at $1.96 each in quantities of 100.

For further information contact Chris Nilsson at (201) 722-9800, Ext. 224. INFO/CARD #127.

New Amplifier Guide Book

The entire line of microwave GaAs FET amplifiers is described in a new 20-page Guide Book just published by The Narda Microwave Corporation. Introductory technical data pages provide an overview of Narda's amplifier features, their key characteristics and examples of special amplifiers built by Narda. Detailed specifications and outline drawings are presented for the company's complete product line which includes a large selection of inventoried amplifiers, tested and stocked for immediate delivery.



Narda's low-noise thin-film GaAs FET amplifiers span the frequency spectrum from 2 to 18 GHz. The standard product line specializes in broadband octave and multi-octave

(Continued on page 45.)

(Continued from page 42.)
amplifiers for use in Electronic Warfare Systems as well as narrow band amplifiers for use in Radar, Telecommunications and Telemetry applications.

Contact The Narda Microwave Corporation, 75 Commercial Street, Plainview, NY 11803; (516) 349-9600. Circle INFO/CARD #128.

20 Watt 2-Way Power Divider

A high power resistive power divider for stripline has been announced by KDI Pyrofilm. The PPD 20-2 has been developed for use in leveling loop applications. It has the advantage over directional couplers or transformer type dividers by providing an improved source match, flat tracking between outputs and operating in the DC to 2.5 GHz range. The PPD 20-2 performs at power levels up to 20 watts with an effective source VSWR of 1.10:1 maximum. The price for the quantity of 1-9 pieces is $49.75 each with delivery from 4-6 weeks.

Send for KDI Pyrofilm Technical Bulletin No. 6, KDI Pyrofilm Corporation, 60 South Jefferson Road, Whippany, NJ 07981; (201) 887-8100. Circle INFO/CARD # 126.

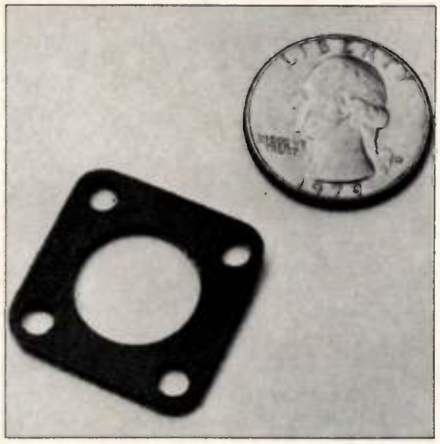
Connector Shielding Flange Gasket

A connector RFI/EMC shielding flange gasket with adhesive back for simple installation has been introduced by ITT Pomona Electronics, Pomona, California.

Available in two styles, model 4780 fits standard BN/BNC flanges. Model 4795 fits standard type "N" and UHF flanges. Both feature monel fibers for positive electromagnetic shielding.

ITT Pomona Electronics manufactures electronic test accessories including banana plugs, jacks and patch cords, test clips and molded accessories.

ITT Pomona Electronics, 1500 E.



Ninth Street, Pomona, CA 91766; (714) 623-3463. INFO/CARD #124.

Alkaline and Mercury Battery Packs

Multiplier Industries Corp., a leading manufacturer of batteries for use in the communications industry has introduced alkaline and mercury battery packs for the Slimline and Omni models of the Motorola HT200, HT220, and MT500 transceivers.

These moderately priced batteries are direct replacements for use with Motorola, Regency Radios, RF Communications, Aerotron, Sonar Radios, Leacom, Public Systems, Multitone, Cook and other transceivers.

Alkaline and mercury batteries are engineered to deliver extended capacity and can be stored for long periods with little loss of rated capacity.

Batteries are in stock and can be shipped next day.

Multiplier Industries Corp., P.O. Box 29, Mount Vernon, NY 10550. (914) 699-0990. INFO/CARD #123.

EMP/EMI Protection

"Solutions for Survivability" is the apt title of the new 4-color brochure published by Chomerics, Inc. Providing an historical perspective of the advancing state-of-the-art, this brochure describes the family of materials that provide solutions for shielding against the effects of EMP, EMI, RFI, Tempest, lightning, static discharge and radiation.

Chomerics is a specialty chemical company which develops and manufactures engineered materials and laminates to solve electronic packaging problems. In addition to offering EMP hardening, these materials typically are used for electromagnetic shielding, grounding, insulating, heat dissipating, sealing and interconnecting. INFO/CARD #122.

Brochure Describes ATE Compatibility of Sweep Generators

Key features of a new family of sweep generators that are designed especially for the ATE market are described in a new 18-page brochure. The programmability of the WILTRON 6600 Series is demonstrated by a complete listing of the mnemonics recognized by the sweep generator. Also included is a sample program, written for the Model 85 controller, that shows how the mnemonics are used. Graphics illustrate ATE applications and the benefits of the programmability. Performance characteristics that exceed those of previously available sweepers and make the 6600 particularly well suited to ATE systems include frequency accuracy, stability, and spectral purity.

Also included in the brochure are important features of the 560 Scalar Network Analyzer, the natural component to the 6600, that are useful to designers. Descriptions of alternative data transfer modes and mnemonics demonstrate the capability to control all aspects of data collection.

Contact Walt Baxter, WILTRON Company, 805 East Middlefield Road, Mountain View, CA 94043. (415) 969-6500, TWX: 910-379-6578. Circle INFO/CARD #121.

3.7 - 4.2 GHz Power Divider

Engelmann Microwave Company has announced the availability of a new 16 way isolated stripline Power Divider, covering the frequency range of 3.7 - 4.2 GHz. Model D1634M provides a minimum output isolation of 20 dB, between adjacent boards with typical isolation figures as high as 25 - 30 dB. The model D1634M can be used either as a Power Divider or Power Combiner with a maximum VSWR of 1.25 in either case.

The total passive insertion loss is 1.2 dB maximum and a typical amplitude balance over any 40 MHz band width is less than .01 dB. The full frequency band amplitude balance is ±0.3 dB. Phase symmetry of balance is maintained over the full frequency band within ±4° maximum.

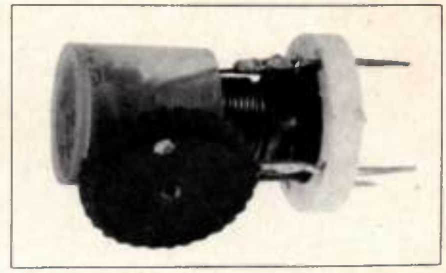
Model D1634M is intended for commercial applications, but is fabricated to complete Mil specifications including a temperature range of −55° to +125°C, and power handling capabilities up to 50 watts CW, 3 kw peak.

Pricing in small quantities is $275 per unit and delivery is guaranteed from stock. Alternate models with the same specifications are available with type N female nickel plated connectors.

Contact Mr. Carl Schraufnagl, Engelmann Microwave Company, 662 Myrtle Ave., Boonton, NJ 07005, or call (201) 334-5700. Please circle INFO/CARD #125.

Variable Linearity Coil

Prem Magnetics, Inc. has introduced a new variable, magnetically-biased linearity coil for TV, computer terminal and word processing terminal applications. The patent-pending design consists of a pair of stationary magnets mounted on ends of a coil and core assembly. A rotatable



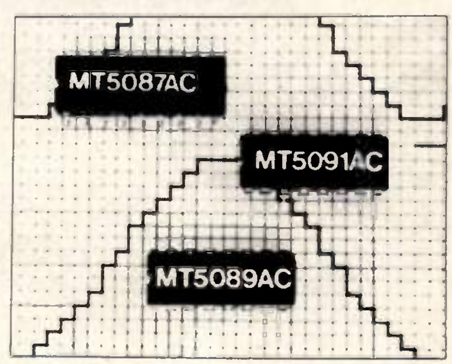
magnet is mounted in proximity to one of the fixed magnets, providing an adjustable mahnetic field. This simplified construction of a readily adjustable linearity coil provides an improved linear display on CRT screens. Write or call Prem Magnetics, Inc., 3521 N. Chapel Hill Road, McHenry, IL 60050. (815) 385-2700. INFO/CARD #120.

ISO2-CMOS DTMF Generator

These monolithic integrated circuits, manufactured by Mitel Semiconductor, are pin and functional replacements for the industry MK5087/89/91 types.

Packaged in 16 pin or 18 pin DIPs, these IC's operate between 2.7 V and 10 V, and take less than 1 microamp of standby supply current. Inputs from either 2-of-8, or Class A single-contact push-button keyboard are converted into a mixed tone comprising of 2-of-8 possible frequencies, each representing a column or row of the keyboard.

The Reference oscillator is completed by an external 3.58 MHz crystal, and auxiliary control input/output options are Mute output, Tone disable input, Transmitter switch output, single tone inhibit input, any key down detect output.



The output tone, regulated or unregulated, is driven by an open emitter bipolar transistor except in the case of the MT5091 (18 pin version) which has an on-chip uncommitted bipolar driver. This feature facilitates external addition of discrete filtering components necessary to meet the European CEPT

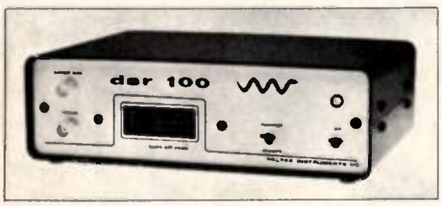
Harmonic distortion recommendations.

In addition to DTMF signalling telephone sets, other applications for these devices includes mobile radio transmitters, remote control or information terminals.

Contact L. Thurlow, Product Marketing Manager, Telecommunications ICs, Semiconductor Marketing, Mitel Corporation. (613) 592-5280. INFO/CARD #119.

Digital Squelch Reader

The DSR-100 decodes digital coded squelch signals and presents the number on a LED three digit readout. It includes an indicator light that lights when the turn-off code is received. The DSR-100 also provides a synchronizing pulse to permit oscilloscope observation of the digital code train. This permits use of an oscilloscope in the analysis of problems in Digital Squelch systems.



The unit functions with the popular digitally coded squelch systems having a 23 Bit continuous code stream, 3 digit Octal ID number and data rate of 134 Bits per second. The DSR-100 will operate over an input voltage range of 15 mV to 10 volts Peak to Peak.

Delivery on the DSR-100 is 2 to 3 weeks ARO. The price is $329. Contact Helper Instruments Company, P.O. Box 3628, Indiatlantic, FL 32903; (305) 777-1440. Please circle INFO/CARD #118.

High Frequency Piston Trimmer Capacitor

The OXLEY[R] range of high frequency piston trimmer capacitors,

type PT6/- has been extended to include devices having values of C max up to 29 pF in a variety of mounting styles compatible with accepted international patterns.

The Trimmers feature a unique construction with the inherent hydrophobic nature of the PTFE dielectric guaranteeing improved performance even in severe environments.

Excellent mechanical stability and resistence to soldering operations is provided by the low aluminia insulation and proprietary HMP solder bonding.

Qualification Approval for military applications against BS9093 F0022 in progress.

Contact Oxley Inc., 6290 Sunset Blvd./Suite 1126, Los Angeles, CA 90028. (213) 463-5120, TWX 910 3212903. INFO/CARD #117.

Hand-Held Digital Multimeter

Weston Instruments, the Newark, NJ manufacturer of test instrumentation and panel meters, announces a new Hand-Held Digital Multimeter, part of the "Roadrunner ADMM"

family. This Model 6120, designated "Roadrunner II" includes all of the standard Roadrunner audible multimeter features, and a combination of new functions unavailable in any other hand-held DMM.

The basic measurement functions include voltage to 100 VDC, 750 VAC, current to 200 mA DC and AC, and resistance to 20 megohms. Basic accuracy is ±0.1 percent.

The unique, audible test function featured in this instrument family operates with very precise thresholds to check both ohms and volts values quickly, for testing semiconductors, checking continuity, and for many other troubleshooting functions in the field.

Contact Catherine Gessner, Weston Instruments, 614 Frelinghuysen Ave., Newark, NJ 07114; (201) 242-2600. INFO/CARD #107.

Miniature Phase Locked Oscillators

RFD, Inc. manufactures a series of Phase Locked Oscillators that cover the frequency range of 0.4 to 5.2 GHz. The sources are fundamental frequency devices, utilizing a high Q stabilized oscillator. RF power output of standard units is +20 dBm between 0.4 and 2.0 GHz and +13 dBm between 2.0 and 5.2 GHz. Internal reference stability is ±0.003

percent over the operating temperature range of 0 to +50°C. Units may be supplied to operate on either an internal or external reference with automatic switch over between the two. Harmonics are specified at −20 dBc and spurious at −80 dBc. Power supply requirements are either positive or negative (12 to 28 VDC) at 150 mA typical. Options include lock limit alarm, custom mounting, operation from −30 to +70°C and higher RF output power up to +26 dBm from 0.4 to 2.0 GHz and +16 dBm from 2.0 to 5.2 GHz. Overall size is 1.25" × 2.25" × 2.25" excluding projections. Delivery is 10 weeks.

Contact RFD, Inc., 5024 Nassau Street, Tampa, FL 33607. Please circle INFO/CARD #116.

10 Microsecond Frequency Synthesizer

Zeta Laboratories' Model 6802 direct frequency synthesizer covers 750 to 1000 MHz in one MHz steps with +15 dBm output. Switching time is 10 microseconds; spurious outputs are less than −50 dBc and harmonics less than −25 dBc. Phase

noise measured in a one Hz bandwidth is −75 dBc at 100 Hz, −85 dBc at 1 kHz and −100 dBc at 100 kHz.

Inputs are a 10 MHz reference, which establishes the long term stability, +28 and +5 volts, and TTL compatible, BCD commands on 10 lines. Size is 40 cubic inches.

Contact Zeta Laboratories, Inc., 3265 Scott Blvd., Santa Clara, CA 95051. (408) 727-6001, TWX 910-338-7336. INFO/CARD #105.

Susceptibility-Testing Antenna

By treating the shielded test room as a cavity and taking on the role of cavity exciter, the AT2000 CAVITENNA™ radiator from Amplifier Research succeeds in handling high power inputs and providing high-intensity fields while taking up one-quarter the space of a comparable log-periodic antenna. The system will handle up to 1250 watts from 30 to 1000 MHz, with maximum power inputs of 3500 watts at the lower end of the spectrum (30-250 MHz) and 2000 watts in the center portion (250-500 MHz).

At rated power, the CAVITENNA system will generate fields that peak at well over 600 Volts/meter in the 200 MHz region.

The CAVITENNA radiator is unique in that it is not intended as a free-space radiator, but is designed specifically for the cavity-like environment of the shielded rooms used for RFI and EMI testing. The new system uses a wall or ceiling of the room as its ground plane, providing very-wide-bandwidth performance in a small package.

The new CAVITENNA radiator is less than four feet long, while a comparable log-periodic antenna for 30 MHz operation would require at least a 16-foot span.

Further information and complete specifications, including the results of tests conducted in a typical shielded room, are available from Amplifier Research, 160 School House Road, Souderton, Pennsylvania 18964. (215) 723-8181. Please circle INFO/CARD #115.

New Vectorscope With CRT Generated Targets

Leader Instruments Corporation, of Hauppauge, NY, recently introduced the Model 5850 Vectorscope.

Phase-amplitude targets are generated by the CRT so that they are illuminated and easy to see as the vector points themselves.

Because the target boxes and the vectors are both produced electronically, there is no error induced by CRT aging. And because the

targets can easily be seen from afar, even in a dimly-lit control room, a simple coaxial switching system permits one Vectorscope to be used remotely to monitor several video sources.

Phase and amplitude adjustments are made more accurate by the illuminated inner target display which represent error limits of +/– 2.5° and +/– 2.5 IRE units. Two loop-through inputs are included that can be selected for display by front-panel push buttons. A test circle pattern is also selectable. The phase reference is chosen from either of the two composite video inputs, and one of these can be switched to phase lock to a subcarrier unit. Another front panel push button switch selects either 100 or 75 percent saturation levels. A gain control permits continuous adjustment as well as a detented calibrated position and a phase control allows you to rotate the display through a full 360° circle. The unit is available in a protective carrying case or is compatible with the industry's standard half-rack mounting configuration.

Contact Leader Instruments Corp., 380 Oser Ave., Hauppauge, NY 11788, or call (516) 231-6900 or toll free (800) 645-5104. Please circle INFO/CARD #114.

Bench Attenuators Cover DC To 2GHz

A new cost effective, high performance series of Weinschel Engineering Bench Attenuators cover the DC - 2 GHz in six attenuation ranges and steps. Models are available with ranges to 140 dB and step resolutions of 0.1, 1 and 10 watt average, 100 power rated at 1 watt average, 100 watts peak.

Each is packaged in a convenient bench type housing with weighted base for stability.

Known as the 3050 Series they exhibit low VSWR: 1.20 to 1.35 maximum depending on model; frequency sensitivity of 0.1 to 0.2 dB up to 2 GHz and repeatability of less than 0.1 dB over their 1,000,000 step life at +75°C. RF leakage is greater than 85 dB below the input level.

Contact Weinschel Engineering, One Weinschel Lane, Gaithersburg, MD 20877. (301) 948-3434. Please circle INFO/CARD #113.

IC Packaging Panels Catalog

The new 24-page Catalog P-81 from Electronic Molding introduces EMC's state-of-the-art press-fit connector backpanel system. Also included: high-density Nurl-Loc® boards and adaptors, as well as EMC Wire-Wrapping capabilities. Copies of the catalog are available from Electronic Molding Corp., 96 Mill St., Woonsocket, RI 02895.

Circle INFO/CARD #108.

Coaxial Cable Assemblies Catalog

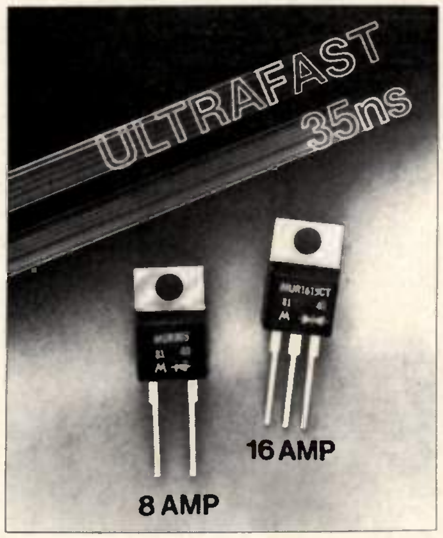
An 18-page catalog features a full line of coaxial cable, coaxial adapters, coaxial connectors, coaxial terminations and coaxial cable assemblies. Pricing on over 1,000 standard catalog items as well as technical specifications are included.

Contact Pasternack Enterprises, 22017 Bushard St., Huntington Beach, CA 92646. (714) 962-9306.

Circle INFO/CARD #111.

35 Nanosecond 8 And 16 Ampere Rectifiers In TO-220 Package

Motorola Rectifiers is expanding its Switchmode Power Rectifier product offering to include ULTRA-



FAST devices with 35 nanosecond recovery time. These are the first of many products in this category and are available in both single (8 ampere) and dual chip (16 ampere) configurations. They are packaged in the popular TO-220 package and have an operating junction temperature of 175°C.

These new Switchmode power rectifiers, which have the major parameters characterized in the following table, will see primary usage in switch power supplies with greater than 5 V outputs. Motorola's volume TO-220 production capability will allow these devices to be offered at very low introductory prices.

The new ULTRAFAST power rectifiers are also designed for use as inverters and as free wheeling diodes. Their faster switching allows these devices to operate more efficiently.

Forward voltage is very low. . . less than 0.85 V at 8 amps, TC = 150°C. The low leakage current of less than 250 μA at 150°C allows safe operation without fear of thermal runaway.

Contact Motorola Semiconductor Products Inc., P.O. Box 20912, Phoenix, AR 85036. INFO/CARD #110.

Low-Cost Multimeter Comes In Kit Form

A precision hand-held 3½-digit multimeter in quick assembly kit form has been introduced by Tomar Ltd. of San Jose. The compact instrument is intended for field and laboratory use, and is designed around the industry-standard ICL7106 A/D converter integrated circuit from Intersil, Inc. Price of new DMM813 is $49.95 in kit form or $54.95 fully assembled.

The DMM813 meter features measurement functions which include DC voltage, AC voltage, DC current and resistance. Readout is on a ½ inch liquid crystal display, and includes parameter and polarity indication and low battery warning. All functions are protected against overload by a diode-transistor network.

Contact Tomar Ltd., 6322 Mohave Dr., San Jose, CA 95126. (408) 997-7685. INFO/CARD #109. □

Classifieds

RF DESIGN ENGINEER

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Primary duties involve design and maintenance of scientific instrumentation and supervision of electronics technician. Experience in radio-frequency design is helpful. Excellent salary and professional benefits, including TIAA/CREF Annuity. Deadline: Dec. 15, 1981. Contact Professor B.R. Ware, Department of Chemistry, Syracuse University, Syracuse, New York 13210, Telephone (315) 423-4645.

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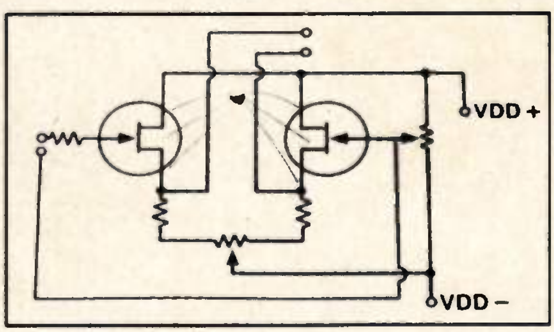

This dual FET is designed for low level amplifiers with input noise voltage typically $1.4nV\sqrt{Hz}$ at 1 kHz. Device has min. Gm of 25,000 μMho per side, assuring voltage gain of 25 min. with 1K drain load. The 10mA operating point is easily held due to low pinch-off voltage, as source follower, CD860 has typical output impedance of 24 ohms. Gm is matched to ±5% and VPO to ±25mV.

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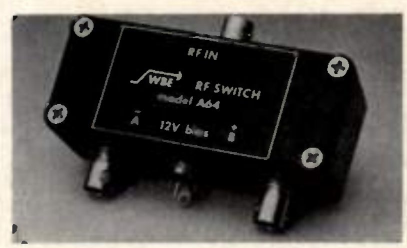

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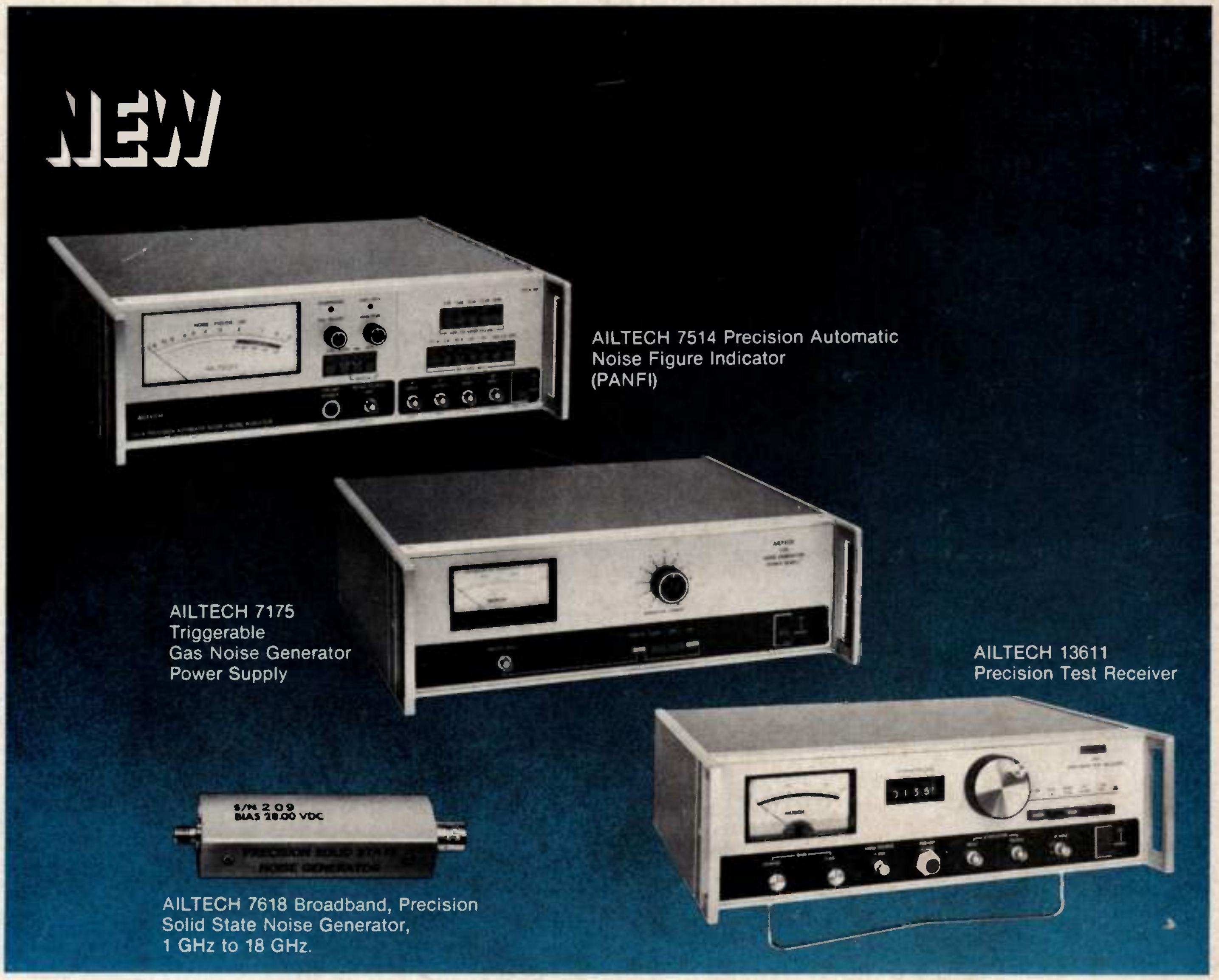

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