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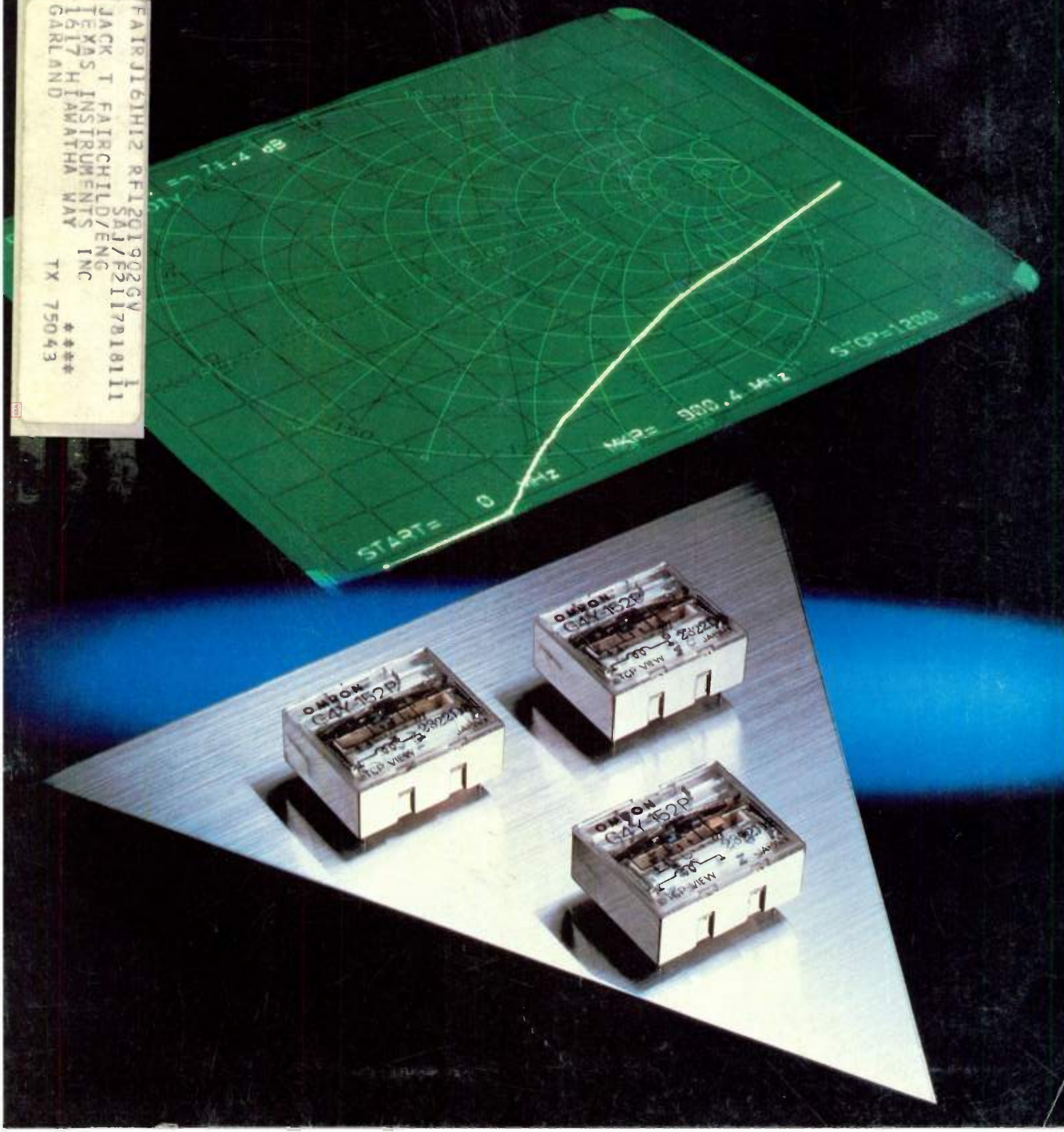

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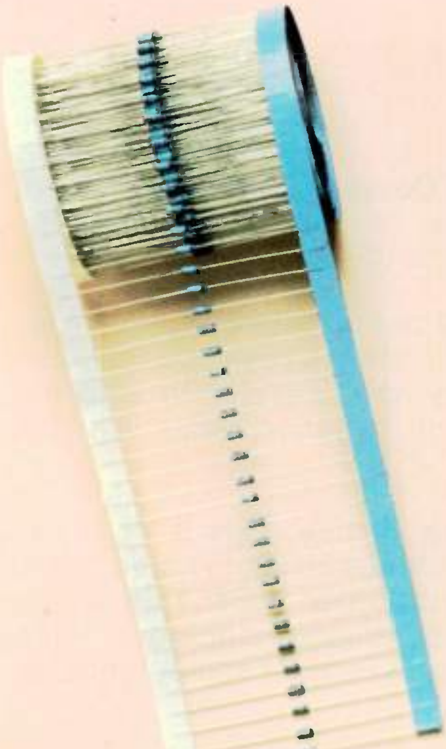

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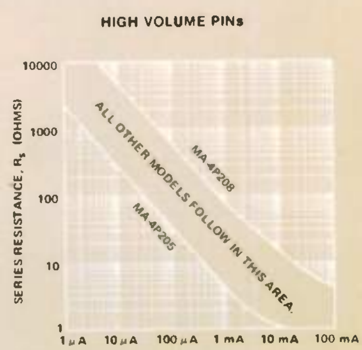

TYPICAL CHARACTERISTICS

Application	Type	Model MA-	V_B (Min.)	V_F (Max. @ 1 mA)	C_T (pF-Max.)	R_S (Ohms @ 100 mA)	R_S (Ohms @ 1 mA)	R_S (Ohms @ 0.01 mA)	T_L (μs)	Max. T_L (ps @ 5 mA)
Switch	PIN	4P205	100	—	1.0 @ 50V	0.4	4	250	1.0	—
Attenuator	PIN	4P208	100	—	0.4 @ 50V	5	90	5000	4.5	—
Detector/ Switch	Schottky	4E2800 (1N5711)	70	0.410	2.0 @ 0V	—	—	—	—	100
Mixer	Schottky	4E2810 (1N5712)	20	0.550	1.2 @ 0V	—	—	—	—	100
Detector	Schottky	4E2835	5	0.340	1.0 @ 0V	—	—	—	—	100

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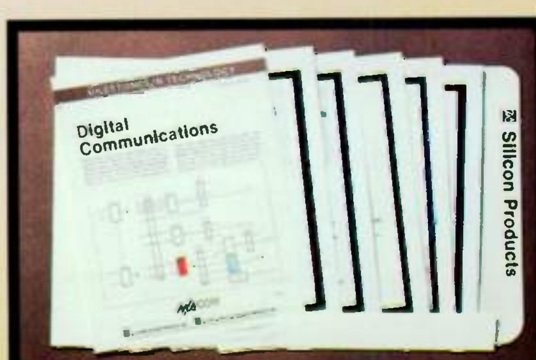

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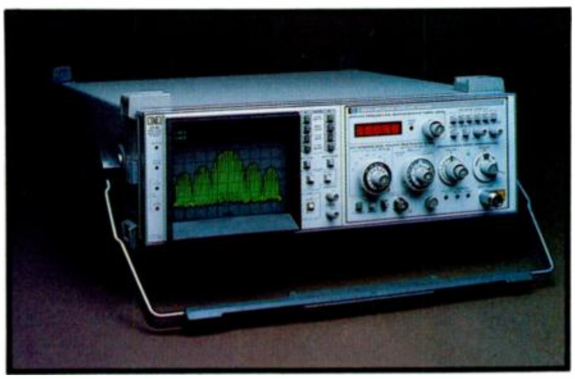

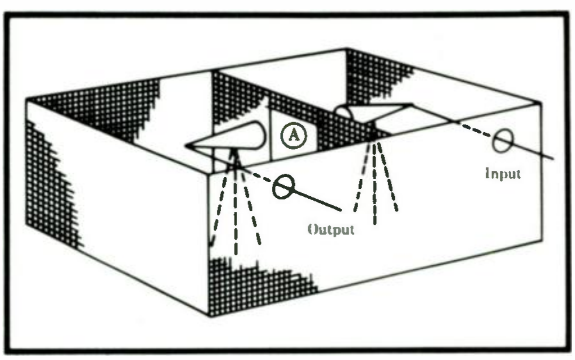

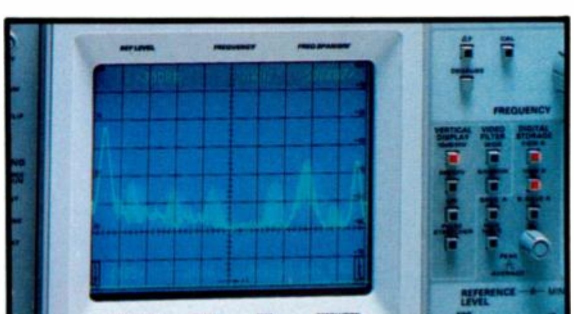

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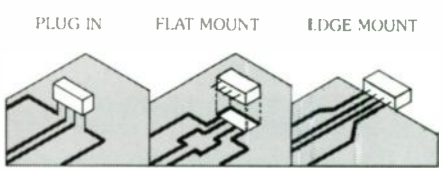

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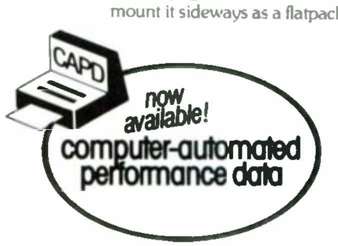
MODEL	FREQUENCY, MHz		CONVERSION LOSS dB, TYPICAL		ISOLATION dB, TYPICAL						PRICE	
			1 Octave from Band Edge	Total Range	Lower Band Edge		Mid Range		Upper Band Edge		$	
	LO/RF	IF			LO-RF	LO-IF	LO-RF	LO-IF	LO-RF	LO-IF	EA.	QTY.
TFM-2	1-1000	DC-1000	6.0	7.0	50	45	40	35	30	25	11.95	(1-49)
TFM-3	.04-400	DC-400	5.3	6.0	60	55	50	45	35	35	19.95	(5-49)
TFM-4	5-1250	DC-1250	6.0	7.5	50	45	40	35	30	25	21.95	(5-49)
•••TFM-11	1-2000	5-600	7.0	7.5	50	45	35	27	25	25	39.95	(1-24)
•••TFM-12	800-1250	50-90	—	6.0	35	30	35	30	35	30	39. 95	(1-24)
▪▪TFM-15	10-3000	10-800	6.3	6.5	35	30	35	30	35	30	49.95	(1-9)
▪▪TFM-150	10-2000	DC-1000	6.0	6.5	32	33	35	30	35	30	39.95	(1-9)

•••If Port is not DC coupled
▪▪+10 dBm LO, +5 dBm RF at 1dB compression
*Units are not QPL listed



For complete specifications and performance curves refer to the Microwaves Product Data Directory, the Goldbook, EEM or Mini-Circuits catalog

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INFO/CARD 5

68 REV. A

Errata

A portion of the HP-41 computer program accompanying Andrzej Przedpelski's article on *Loop Antenna Design* (November/December 1982, Page 14, Table II) dropped out in printing. We have reprinted a solid copy of it here.

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

R_{00}	
R_{01}	R_O
R_{02}	C_S
R_{03}	C_{TS}
R_{04}	L
R_{05}	Q_L
R_{06}	f
R_{07}	[2πf]
R_{08}	[ ]
R_{09}	[ ]
R_{10}	[ ]
R_{11}	[ ]
R_{12}	
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Getting the most out of your SPECTRUM ANALYZER

Combining a tracking generator with a spectrum analyzer to create a high dynamic range display network analyzer

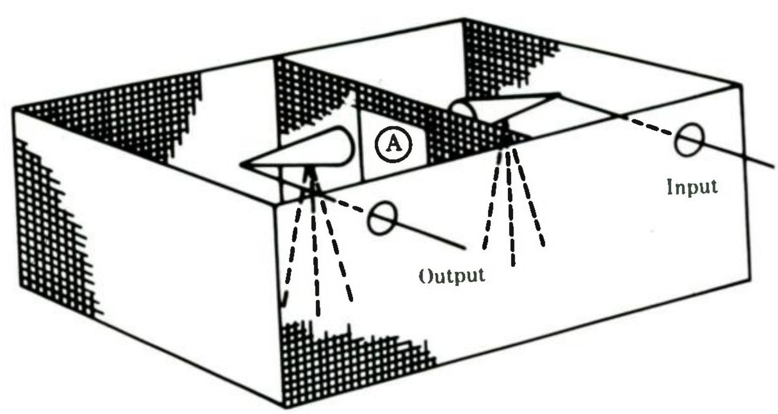

Figure 1. Attenuation Test Chamber with Partition.

Frequency	Power (Max)	Sensitivity (Min)	Dynamic Range
10 kHz - 2 GHz	– 5 dBm	– 105 dBm	100 dB
2 GHz - 4 GHz	– 5 dBm	– 110 dBm	105 dB
4 GHz - 8 GHz	– 15 dBm	– 105 dBm	90 dB
8 GHz - 12.4 GHz	– 25 dBm	– 100 dBm	75 dB
10 GHz - 22 GHz	< – 25 dBm	– 90 dBm	65 dB

Table I. Output Power vs. Analyzer Sensitivity vs. Dynamic Range.

By D. Krautheimer, M/A-COM MPD, Inc. and A. Freeland, Eaton Corp., Electronic Instrumentation Division.

The combination of a spectrum analyzer with a tracking generator creates an easy-to-use high dynamic range display (to 100 dB on screen) network analyzer from 10 KHz to 12.4 kHz. Unlike standard network analyzers made up of a wide band sweep generator and a broadband input network analyzer, the tracking generator, spectrum analyzer combination is free of spurious and harmonic responses. This is because the undesirable signals fall outside the analyzer's passband and thus will not be seen. The tracking generator will not add any broadband noise thus leaving the analyzer's sensitivity unaffected. While a sweeper will usually deteriorate a broadband network analyzer's sensitivity by up to 20 dB or more.

The spectrum analyzer has a sensitivity of $\geq$ 100 dB up to 12.4 GHz. Specifically in the frequency range up to 2 GHz, the tracking generator has an output level of approximately – 5 dBm and the analyzer a sensitivity of – 105 dBm allowing up to 100 dB display of attenuations. Table I contains a detailed listing of output power vs. analyzer sensitivity vs. dynamic range.

Most broadband network analyzers have a sensitivity of about −60 dBm. Since the analyzer's sensitivity varies between −90 and −100 dBm, an advantage of 30 to 50 dB in sensitivity is achieved. The analyzer's dynamic display range of 100 dB can be fully utilized by controlling the output level of the tracking generator. For instance, in the range from 4 – 8 GHz, the output of the tracking sweep generator is typically −15 dBm. The sensitivity of the analyzer is −105 dBm which equals about 90 dB of dynamic range. If an amplifier of 10 dB or more were placed after the tracking generator to boost its output level above −5 dBm a dynamic range of 100 dB could be achieved.

Often test systems have a built in loss due to cable losses, matching pads, mismatches, insertion loss, etc. An inherent loss of 20 dB would subtract from the dynamic range of the measurement unless compensated for by an amplifier. How do you know if the signal level is sufficient for 100 dB dynamic range? If the waveform deflection reaches the top of the display (0dB) without saturation and the internal noise level of the analyzer is at the 100 dB line, a dynamic display of 100 dB can be achieved. If the waveform deflection reaches only the −40 dB graticule line, a dynamic range of 60 dB is obtained. Therefore, amplification of the output level of the tracking generator by 40 dB would restore the dynamic range back to 100 dB.

Measuring Shielding Effectiveness or Insertion Loss of Materials

A specific application to measure the shielding effectiveness of a plastic material used as a calculator case involved testing the material in a special test chamber with antennas

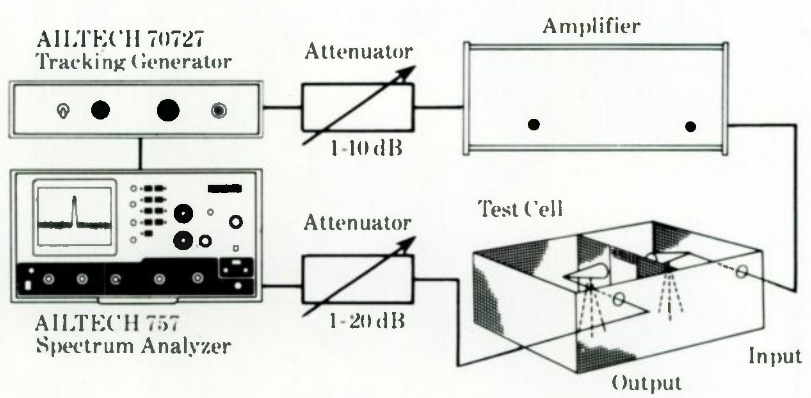

Figure 2. Spectrum Analyzer/Tracking Generator System.

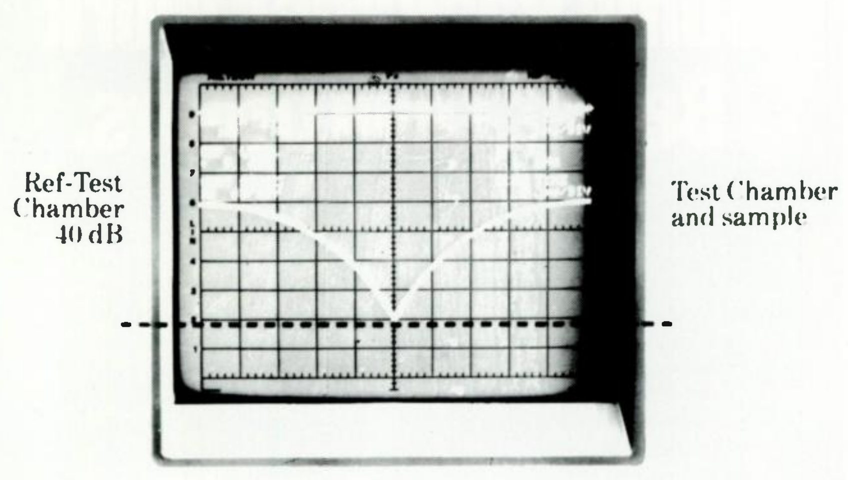

Figure 3. Spectrum Analyzer Display.

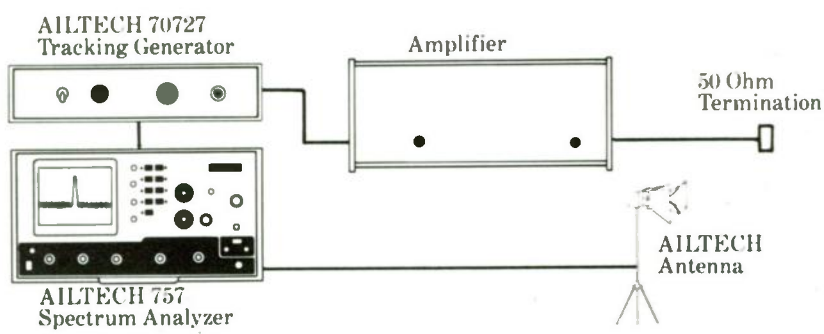

Figure 4. Generator/Analyzer System Use with Coaxial Cable.

on either side of the material. Previously the measurement had been done using a power oscillator at a single frequency, an oscilloscope as an indicator and an attenuator to measure the effectiveness.

First a reference level was established on the oscilloscope with the sample across the hole (A). With the sample removed, the attenuator was increased until the reference line was reached. The amount of attenuation added equalled the shielding effectiveness of the material. Attenuations of about 40 – 50 dB were to be expected in the range up to 1 GHz.

Since the dynamic display range of the analyzer tracking generator system is 100 dB up to 2 GHz, no significant problems were to be expected. On connecting the tracking generator to the input of the test chamber and the analyzer to the output of the chamber, a loss of 80 dB up to 300 MHz was found in the test chamber with no sample across the hole (A) in the partition. Above 300 MHz, reflections were so large that no meaningful information could be obtained. With test samples across the hole (A) in the partition, a dynamic range of 20 dB was available. No correlation could be made to the power oscillator procedure.

A linear power amplifier was added to the output of the tracking generator making possible an output level of up to more than 50 watts. A 1 to 10 dB variable attenuator was added before the amplifier to provide fine control of the amplifier output level.

The analyzer tracking generator system (Fig. 2) allowed the user to measure the insertion loss of the material quickly over a wide frequency range. Fig. 3 shows the displayed results as seen on the Spectrum Analyzer. Similarly, the shielding effectiveness of coaxial cables can be measured using the analyzer tracking generator system.

Correction factors or losses of antennas or other pick up devices can be as high as 20 dB or more. Depending upon the frequency range, an amplification of between 20 and 55 dB would be necessary at the output of the tracking generator to achieve full dynamic range.

By using the Spectrum Analyzer's digital storage capability, reference levels can be stored and displayed simultaneously, with the attenuation level of the device under test. In addition a normalization feature allows the user to subtract the frequency response of his device under test from that of the memory. This virtually cancels out errors that would exist due to mismatches or the like. □

REF LEVEL 10 dBm

INPUT ATTEN 10 dB ATTEN

LOG SCALE 10 dB/

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SPECTRUM ANALYSIS BASICS PART 2

Part I of this article, appearing in March/April 81, discussed basic types of spectrum analyzers, equipment limitations, safety procedures and spectrum analyzer selection criteria. Part II will cover basic measurement techniques, interface considerations and computer aided measurements.

By James Beck
Eagle (Consultants)
Fallbrook, CA

Basic Measurement Techniques

The quality of spectrum analyzer measurements will only be as good as the spectrum analyzer measurement system itself. The system does not consist of just the spectrum analyzer but all of the interfaces to the device under test as well. Additionally, in the presence of high level fields, indirect radiation or conduction paths must be considered, i.e. power supply feedthrough and radiation through the case of the analyzer.

Figure 1 illustrates a proper installation. This system insures that the front end of the spectrum analyzer is protected from overdrive. The screened enclosure protects the measurements from direct radiation and the main filter eliminates conducted interference. If the DUT is well shielded and the pad inserts minimal attenuation, the shielded enclosure is not required.

The following guidelines should be followed for best performance:

1. Coax "B" should be as short as possible.
2. Coax "A" should be routed away from the case of the spectrum analyzer.
3. Utilize the minimum amount of external attenuation required to present a safe level to the analyzer.
4. DUT and power supply should be located at least 1 wavelength away from the spectrum analyzer when frequency is above 100 MHz.

The external attenuator is very important and must be properly selected to insure good test data. To determine the proper amount of attenuation required, divide the input power from the DUT by the amount of power that the spectrum analyzer can safely handle. Convert the result to dB and select a pad that has at least that amount but no more than 10 dB over that amount.

$$\text{REQUIRED ATTENUATION} = 10 \log \frac{\text{TEST POWER}}{\text{ANALYZER RATED POWER}}$$

Using a pad with excessive attenuation decreases the immunity to direct feedaround via ground loops and other means.

Figure 2 illustrates a typical immunity test of the measurement system to direct radiation pickup. Initially the transmitter was connected directly into the pad. The analyzer was then adjusted so that the signal indicated a 0.0 dB reference. Then the interconnecting cable was removed and an antenna installed at the transmitter output. At this point the analyzer indicated −70 dB. It is very reasonable to assume that the antenna was at least 30 dB higher than normal case radiation which means that the radiated pickup is at least −100 dB below a direct input. This test was performed at 175 MHz with a power level of 100 watts. With at least 100 dB isolation this system will be adequate for harmonic measurements to at least -80 dB.

In a second test the antenna was removed and a dummy load (not the 20 dB pad) was installed. In this test no signal from the transmitter could be detected on the spectrum analyzer, which indicates that the actual case radiation was far better than the -30 dB used above.

Enhanced Harmonic Measurements

Spectrum analyzers, like all equipment, have certain limitations that must be considered before attempting measurements. In harmonic measurements, the spurious free dynamic range must be considered. If the analyzer to be used has a 70 dB dynamic range it will be useful in analyzing signals with a dynamic range of 60 to 65 dB. Additionally, the 70 dB specification occurs at a specific input power to the spectrum analyzer's mixer. By various combinations of internal and external attenuators this power level can usually be obtained.

In making a harmonic measurement of 75 dB the above limitation becomes apparent. By carefully limiting the input power to the maximum undistorted level, the noise floor of the analyzer is around -70 dB, effectively masking the harmonic at -75 dB. Removing 10 dB of attenuation causes the harmonic to increase 20 dB which is a definite indication of mixer overload.

The harmonic dynamic range can easily be increased via the use of a notch filter. Refer to Figure 3 illustrating the implementation of this method. The rules for this system are the same as for the system illustrated in Figure 1. Keep both "B" coaxes short, or use double male adapters, which is preferable.

Tune the notch filter to the fundamental frequency, remove 10 or 20 dB of input attenuation as required and the harmonic at -75 dB can now easily be seen. Overload is no longer a problem because the notch filter is attenuating the large fundamental that was causing the overload.

The accuracy of the harmonic measurement is now dependent on the amount of notch filter insertion loss at the harmonic frequencies. Therefore the insertion loss should be minimal or known at all harmonic frequencies of interest.

If one is making low precision harmonic measurements of ± 3 dB and the combination of notch filter and spectrum analyzer errors is ± 2 dB then there is no problem. A reasonable notch filter should have less than 2 dB of insertion loss over the range of interest. Typically notch filters are used in conjunction with analyzers below 1.5 GHz, the reason being that analyzers above 1.5 GHz use a tracking front end filter by virtue of their mixing schemes, improving dynamic range.

Should a higher degree of accuracy be required, the following procedure can be used.

1. Determine insertion loss of the notch filter at each frequency of interest.

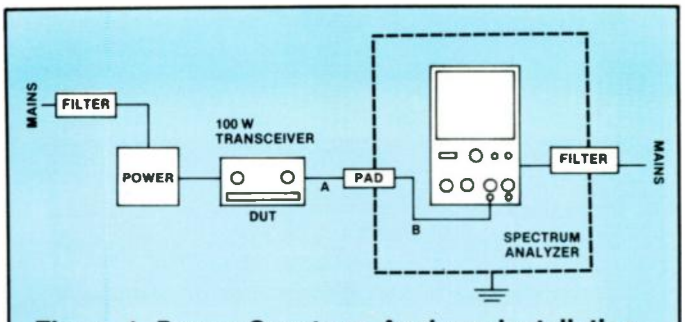

Figure 1. Proper Spectrum Analyzer Installation.

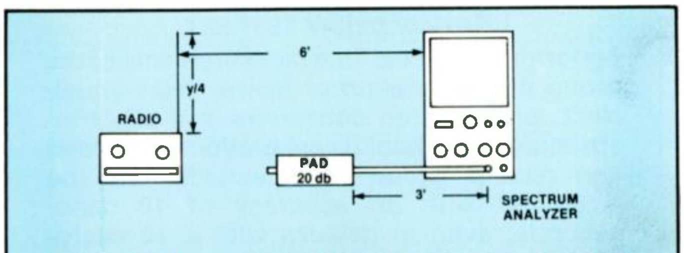

Figure 2. Measurement System Immunity Test.

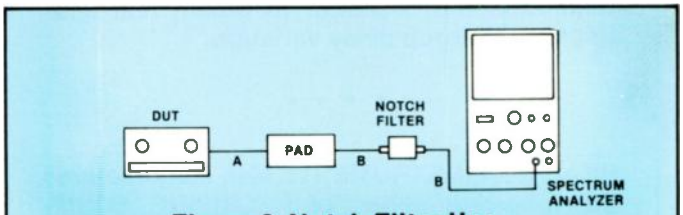

Figure 3. Notch Filter Use.

2. Read spectrum analyzer then subtract the notch filter transmission loss from this reading. Formula for this is as follows:

$P_A = P_r -$ Filter loss

where: P_A = actual power
P_r = SA reading

NOTE: Loss is negative meaning that in absolute value actual power will be higher than the SA reading.

The following is the preferred sequence of events during a harmonic measurement using a notch filter:

1. Connect Pad directly to SA, turn on DUT and set fundamental to 0 dB.
2. Insert notch filter and adjust for maximum attenuation of fundamental.
3. Remove 10 or 20 dB of attenuation.
4. Observe harmonics and subtract 10 or 20 dB as required, e.g.

– 50 dB reading minus 20 dB (attenuation removed) = – 70 dB actual harmonic level.

Enhanced Non-Harmonic Measurements

The notch filter can be useful in extending the dynamic range of non-harmonic measurements. If the frequencies to be measured are outside of the notch disturbance width, the technique is precisely the same as for harmonic measurements described previously. Should the signal occur in the notch disturbance width, dynamic range will be diminished but by carefully determining insertion loss at the frequencies of interest, a useful measurement is still possible.

Here is an example employing an intermodulation test using a 45 MHz and 55 MHz tone. Frequencies of interest are the third order products appearing at 35 MHz and 65 MHz. Additionally these frequencies are expected to be – 80 dB which is lower than the spectrum analyzer's guaranteed dynamic range.

Using a single notch filter to equally reject both carriers is accomplished by tuning the filter to 50 MHz. The passband characteristics of the filter are measured and the attenuation at the frequencies of interest is noted, as contained in Table I.

Frequency	Attenuation
35 MHz	– 12 dB
45 MHz	– 30 dB
55 MHz	– 30 dB
65 MHz	– 10 dB

Table I. Filter Passband Characteristics.

The DUT output is connected directly to the spectrum analyzer input and the tone levels are adjusted so that each tone is at the 0.0 dB level on the display. The notch filter is installed between the DUT and SA input as in Figure 1 and power is applied. As expected the power level of each tone is now at – 30 dB. 30 dB of attenuation is removed from the spectrum analyzer and the 3rd order products are now visible on the screen. The two 3rd order products are measured and found in Table II.

Frequency	Attenuation
35 MHz	– 63 dB indicated
65 MHz	– 61 dB indicated

Table II. 3rd Order Measurements.

To determine the actual level the following formula is used:

$P = P_r - (AR + A_n)$

Where: P = actual tone power
P_r = SA reading (indicated level)
AR = attenuation removed
A_n = gain of notch filter at this frequency.

Therefore: @ 35MHz $P = -63 - (30 + (-12)) = -81$dB
@ 65MHz $P = -61 - (30 + (-10)) = -81$dB

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Frequency Range (MHz)	5 to 2000	5 to 1000	200 to 2000	5 to 2000	5 to 1000	200 to 2000	5 to 2000	5 to 1000	200 to 2000	5 to 2000	5 to 1000	200 to 2000
VSWR	1.1	1.05	1.15	1.1	1.05	1.15	1.1	1.1	1.2	1.1	1.1	1.2
Isolation (dB)	80	60	50	80	60	50	60	60	45	60	60	45
Insertion Loss	1.2	0.5	0.5	1.2	0.5	0.5	1.3	0.6	0.6	1.3	0.6	0.6

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Thus the actual intermodulation components are −81 dB referenced to a single frequency. During the actual measurement the screen displayed a 63 dB difference between the notched fundamental frequencies and distortion products, which is well within the range of a spectrum analyzer rated at 70 dB dynamic range.

Nonstandard Coupling Devices

There are several methods commonly used to couple signals into spectrum analyzers as follows:

1. *Scope probes:* Scope probes are often used to couple signals directly into analyzers because of the convenience and availability of the probes. Probes will not yield proper, or constant attenuation unless designed to supply a 50 ohm port. Additionally, the frequency response of probes is usually much less than that of spectrum analyzers and direct pickup of RF fields will be worse than with a direct coupled system.

2. *Capacitive and resistive pickoffs:* These are used to attenuate the signal to be measured, provide an impedance match to the analyzer, or present a relatively high impedance to the DUT. These devices can be useful in making relative measurements when absolute accuracy is not essential. An example is when tuning a transmitter for best spurious performance or aligning a multiplier stage. For precise absolute measurements these devices are very unreliable and really worthless.

3. *Non-Standard Impedances:* There are two common methods used to match non-standard impedances to a 50 ohm spectrum analyzer input. The resistive match consists of a series resistor in line with the SA input. The second method is to use a broadband matching transformer. Both of these devices are available for spectrum analyzers to match the more common transmission line impedances.

Interface Considerations

Referring to Figure 1, the DUT is a power amplifier with 100 watts power output connected via cable "A" to a nominal 20 dB pad which is connected to cable "B" that is finally connected to the spectrum analyzer RF input. This test set is for the purpose of harmonic measurements to 1.0 GHz with an accuracy of ±0.5 dB relative to the 100 MHz fundamental. Realizing that there are certain losses to deal with, Table III is created.

Element	Gain	
	100 MHz	1.0 GHz
Device Under Test (DUT)	0 dB	unknown
Cable "A" RG-58 U 3 ft.	−0.15 dB	−0.66 dB
Pad-nom 20 dB	−20.00 dB	−20.00 dB
Cable "B" RG-58 U	−0.15 dB	−0.66 dB
TOTALS	−20.30 dB	−21.32 dB

Table III.

For the moment, the other harmonics are ignored since dealing with these uses the same principle as the 1.0 GHz harmonic. Power is applied and the 100 MHz signal is carefully adjusted to 0.0 dB reference on the spectrum analyzer. The 1.0 GHz harmonic is located at −55 dB on the analyzer display. From the above table the test set gain is found to be −1.02 dB relative to 100 MHz; this is subtracted as follows:

$$G_t = G_r - G_d = -55 - (-1.02) = -53.98$$

The system gain is subtracted because it has already affected the reading indicated by the spectrum analyzer.

The DUT specifications indicate the 1.0 GHz harmonic was measured at −53 dB by the equipment manufacturer. Thus the above measurement is in error by 0.98 dB which is not within the allowed measurement tolerance. The reason for this error is the attenuation of the test set is not what had been calculated. The pad was really −20.5 dB at 1.0 GHz and the calculation forgot the connecter losses at 0.1 dB per connection at 1.0 GHz.

As the above example illustrates, attempting to calculate all of the interface losses can be very laborious and prone to error. This step can be eliminated by a simple technique known as substitution. First, replace the DUT with a constant amplitude sweep generator. Assume the generator has 1 watt (−20 dB below 100 watts) and the output level is within ±0.25 dB over the range of 100 MHz to 1.0 GHz. Second, measure the test set gain. This is found to be −40.4 dB at 100 MHz and −42.3 dB at 1.0 GHz. Thus relative to 100 MHz the 1.0 GHz gain is −1.9 dB. The additional 20 dB of loss is due to the fact that the test power is −20 dB related to 100 watts. Using −1.9 dB in the formula to calculate the harmonic level, the result is:

$$G_t = G_r - G_d = -55 - (-1.9) = -53.1$$

The measurement is now in error by 0.1 dB which is well within allowable measurement tolerance.

In the above case the amplitude vs frequency errors were measured at the same power level. Since the 1.0 GHz amplitude is at a much lower power level the amplitude tracking error must be reckoned. For the purpose of the example, this was considered to be 0 dB. In actual practice the error can be as much as ±1.5 dB over the 70 dB display range of the spectrum analyzer. This error can be reduced by decreasing the dynamic range of the measurement using the notch filter. Also, a precision attenuator can be inserted between the sweep oscillator and spectrum analyzer and a second calibration run made. The results of this run can then be compared to the first run and the amplitude tracking accuracy of the analyzer determined.

By using the substitution method, the measurement system is treated as a "black box" whose relative performance is measured which in turn measures the DUT. It is also possible to measure the separate entities of the test set and then treat each one as a separate term in the final equation.

In using the substitution method the following pitfalls must be considered:

1. The calibration reference (sweep oscillator) may have errors of its own. This is the prime item that will limit the accuracy of the measurement.
2. The test set may be unstable which means that the test set gain may vary somewhat over time. This can be due to temperature effects, loose connectors or voltage instability etc.
3. System gain may change at different power levels. This can be due to attenuator heating, dielectric effects or amplitude tracking errors.

Computer Aided Measurements

The previous section discussed substitution as a technique to enhance the amplitude accuracy of the measurements. The total process of determining system errors via

substitution and then calculating them out during the actual measurement is known as *normalization.*

A typical system might consist of a signal generator, one coax cable and a spectrum analyzer. In this system each of these three items will be considered a "black box". In other words, it will not be precisely known where inaccuracies are occuring. For example, in the spectrum analyzer, the attenuator, mixer or the display may cause inaccuracies.

The test system gain would be expressed:

$$G_t = G_{sg} + G_c + G_{sa}$$

where: G_t = overall system gain
G_{sg} = gain of signal generator
G_c = gain of cable
G_{sa} = gain of spectrum analyzer

The gains of these items are known and are as follows:

$$G_{sg} = -10\ \text{dB},\ G_c = -2\ \text{dB},\ G_{sa} = -3\ \text{dB}$$

When the measurement is performed, the signal is displayed at −15 dB as calculated. Not only is the total system gain known but the individual component gain as well. Each part of the measurement is referred to as a *term*. An important conclusion is that if two of the three terms are known and the final result is known, the remaining term can be calculated.

The process of normalization consists of determining the gain of the system at each frequency and then compensating for it during actual measurements. The overall system accuracy may now be expressed as:

Error = Reference Error + Stability Error + Amplitude Tracking Error.

Using the above method with an HP-8350A sweep oscillator as a reference and an HP-8568A spectrum analyzer, the system accuracy at reference amplitude is better than ±0.3 dB. Over the 70 dB dynamic range, the accuracy remains better than ±1.4 dB. The main drawback with the normalization process is the enormous number of calculations required. This is laborious and, quite frankly, boring. Arithmetic errors are also frequent, causing confused data and results.

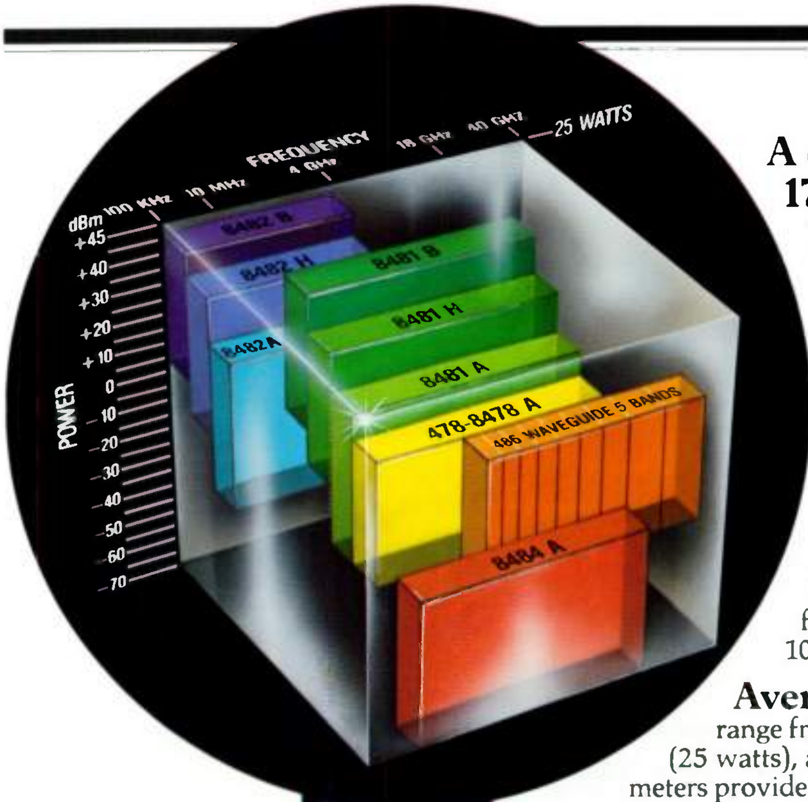
The HP-8568A has a trace memory feature that will make this process very easy. Simply make the calibration run and store this trace in memory. Then make the DUT run and subtract the previous trace from the present trace and display the result. This is now a normalized measurement with no calculations performed by the operator.

For some of the lower cost spectrum analyzers that do not have the nifty trace memory feature, a storage normalizer is available. This device basically contains A to D converters and digital storage arrays. When connected to a spectrum analyzer this device will read a trace, store it and then present the trace on the display. The normalizer has a memory page which can be stored and then subtracted from the present sweep. The normalization is then accomplished as explained above. In addition to normalizing, a normalizer can be used to display a slow sweep at viewable rates once the sweep is in memory. Another benefit in using the memory feature is to store a previous display for comparison purposes with the present display, thus eliminating grease pencils!

Computer Enhancement

A computer-controlled spectrum analysis system utilizes a computer to automatically perform many of the operator

HP's spectrum of power measurements.



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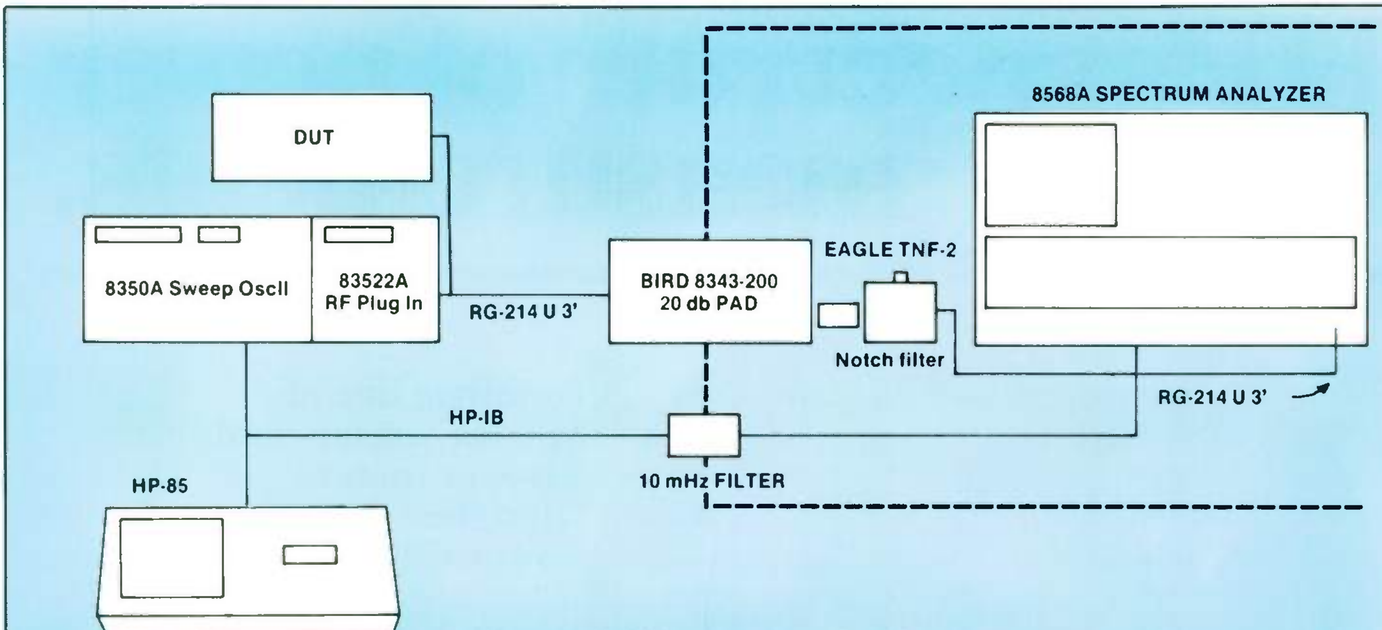

Figure 4. 100-Watt Spectrum Test Set.

functions. Figure 4 illustrates a typical automated system. In this system the computer sets the analyzer functions, controls the signal generator, performs a calibration run using the sweep generator, makes the test run using the DUT, performs normalization calculations and presents final data.

In this system a notch filter is used to extend the dynamic range of the system to 90 dB for harmonic measurements. Before the calibration run, the notch filter is adjusted to the DUT carrier frequency. During the calibration run, the disturbances caused by the notch filter are stored. After the test run these errors are automatically factored out which means that the final data will show the exact relationship between the various frequencies monitored. All operator calculations are thus eliminated reducing the chance for errors in the final data. □

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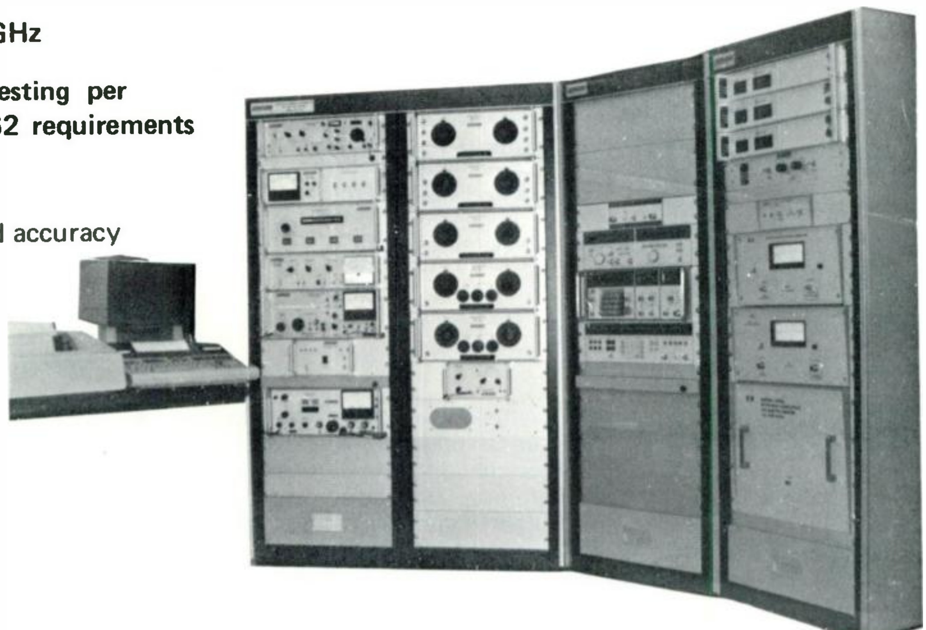

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Designing a High-Stability VHF Oscillator

This article describes the author's experience in developing a high-stability (± 5 ppm) VHF overtone oscillator and presents a novel circuit configuration that has proven superior in achieving the required stability.

By Donald K. Belcher, PE
Harris Corp., GESD
Melbourne, FL

Background

Since fundamental model crystals cannot reliably be fabricated above 30 MHz, due to the fragile dimension (in the range of 2 mils (0.002") thick at 30 MHz) of the blank, VHF crystal oscillators are normally implemented using overtone crystals. To achieve high frequencies, odd overtone modes are used; 3rd, 5th, 7th and sometimes 9th.

Overtone oscillators generally require more complicated circuitry because the oscillator must be "fooled" into operating on the desired overtone and not the fundamental. This is normally accomplished by the use of frequency selective feedback; i.e., a form of zonal filtering in the feedback loop. Figure 1 illustrates this concept.

Notice that the crystal exhibits many responses (approximately harmonic related) and the relatively broad zonal filter selects the appropriate overtone.

Above 100 MHz, a common base configuration is normally used. It has several advantages:

- It operates the crystal in a series resonant mode.
- It is relatively easy to design.
- Tuning procedures are straightforward.

Stability considerations involve the normal temperature characteristics of the crystal plus the temperature effects of the oscillator circuitry. It is these "temperature effects of the oscillator circuitry" that present the major difficulty when attempting high stability oscillators, particularly with common base.

A typical requirement might be to achieve ± 5 ppm stability over temperature. It has been determined by analysis and verified by lab results that the common base circuitry can contribute as much as 1.5 ppm to the frequency instability of the oscillator. This can be intuitively understood by refering to Figure 2, which is a simplified model of a crystal and zonal filter as described in Figure 1.

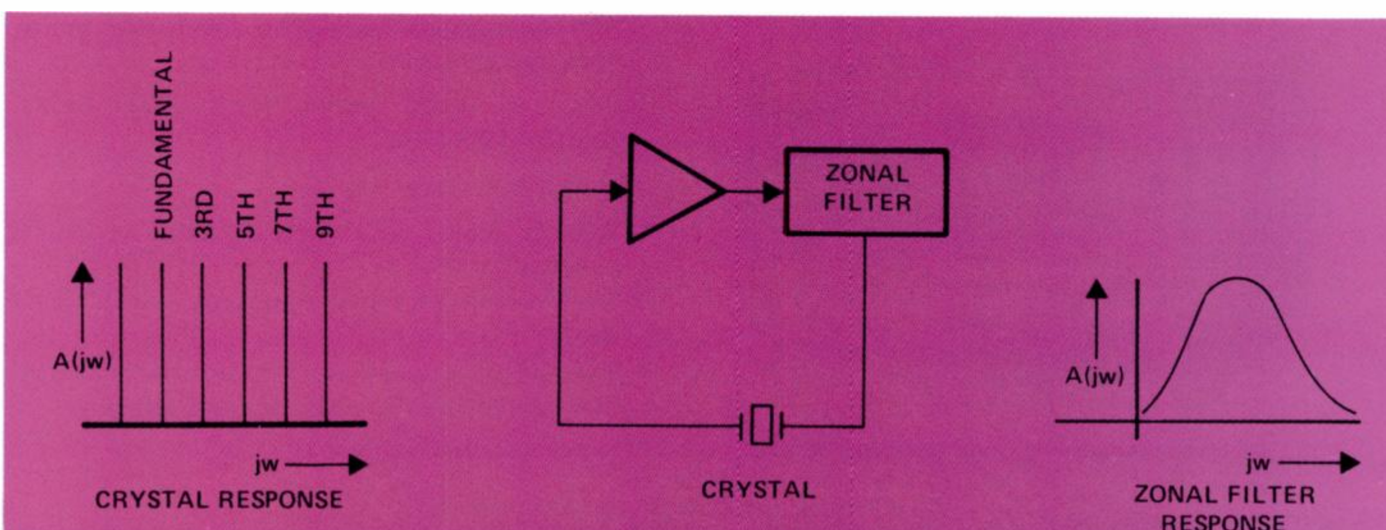

Figure 1. Crystal and Zonal Filter Responses.

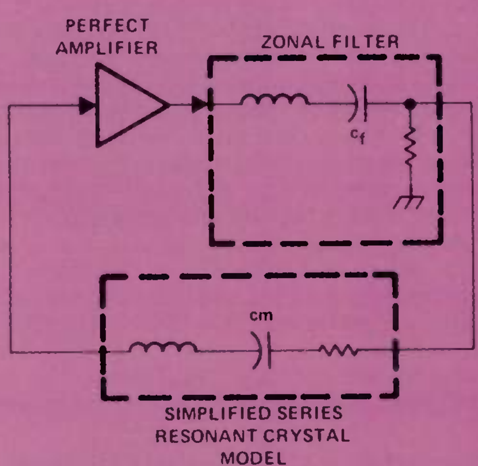

Figure 2. Simplified Oscillator Mode, Common Base.

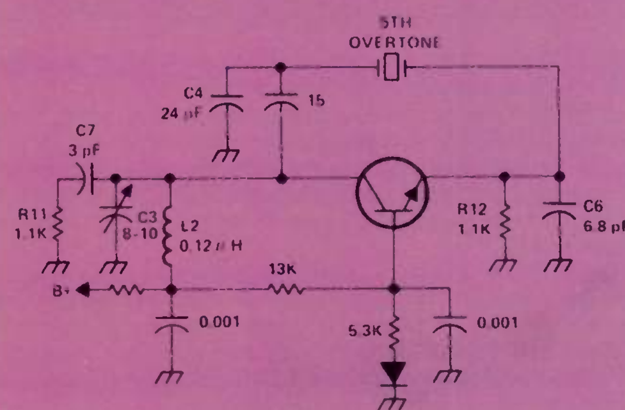

Figure 3. Typical 100 MHz Common Base Oscillator.

Assuming a 5th overtone crystal is used in the 100 MHz range, the maximum crystal Q achievable is in the range of 70,000. Further, assuming all variations in the oscillator are lumped into the zonal filter, variations in phase of the zonal filter transfer function will shift the operating frequency of the overall amount as determined by the crystal Q (or phase slope). As a rough approximation, the filter phase will shift ± 45° over the 3 dB bandwidth. The same assumption can be made regarding the crystal transfer function. This yields the following relationship:

$$\Delta\Phi \cong \frac{90\,\Delta f_c\,Q_x}{10^6} \qquad (1)$$

where:

$\Delta\Phi$ is the allowable phase shift in the oscillating circuitry.

Δf_c is the budgeted frequency error in ppm due to the oscillator circuitry.

Q_x is the crystal Q

As an example, with the crystal mentioned above and a budgeted circuitry error of 0.5 ppm, the allowable phase

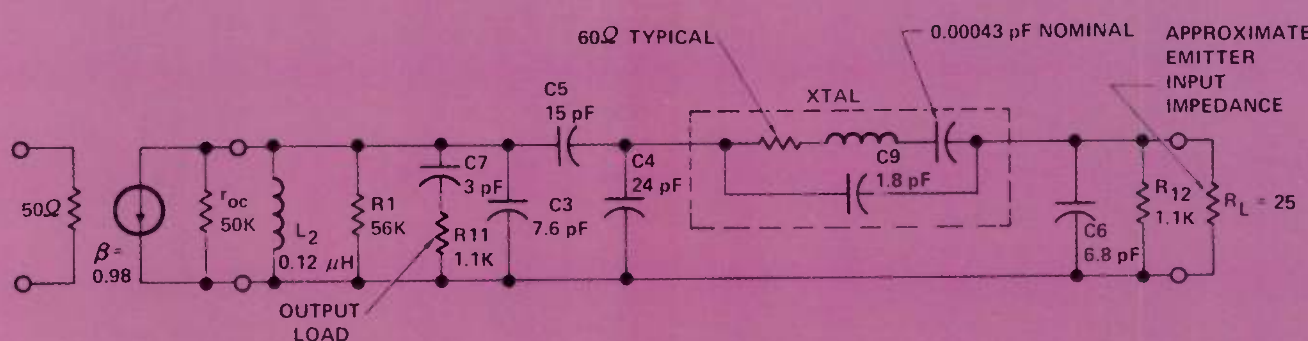

Figure 4. Computer Model of Common Base Oscillator.

Component		Original Value	Sensitivity
roc	(Transistor Output Z)	50.0K	0.005 ppm/%
R1	(Inductor Loss)	56.0K	0.009 ppm/%
L2	(Tuning Inductor)	0.12 μH	4.93 ppm/%
C3	(Tuning Capacitor)	7.6 pF	4.917 ppm/%
C4	(Feedback Capacitor)	24.0 pF	0.897 ppm/%
C5	(Feedback Capacitor)	15.0 pF	1.79 ppm/%
C6	(Phasing Capacitor)	6.8 pF	0.090 ppm/%
C7	(Output Capacitor)	3.0 pF	0.009 ppm/%
C9	(Crystal Case Capacitance)	1.8 pF	0.004 ppm/%
R11	(Output Load)	1.1K	0.009 ppm/%
R12	(Emitter Bias Resistor)	1.1K	0.009 ppm/%

Table I. Analysis Results

shift of the oscillator loop would be only ≅ 3°. This is obviously hard to achieve.

Another more rigorous analysis done by computer modeling via COMPACT[1] will yield a set of sensitivities; i.e., the partial derivatives of frequency with respect to the various parameters. Figure 3 shows the common base oscillator schematic, followed by Figure 4 depicting the computer model used.

Table I lists the results of that analysis. Note that the various sensitivities are listed in parts-per-million per-percent change in component value. Notice in particular that L2 and C3 have the highest sensitivities and that a fractional percentage change would cause significant frequency error.

This analysis is further substantiated by observing laboratory data. Figure 5 shows the crystal oscillator performance over temperature. Note the additional frequency error introduced by the oscillator circuitry on the order of 1.5 ppm.

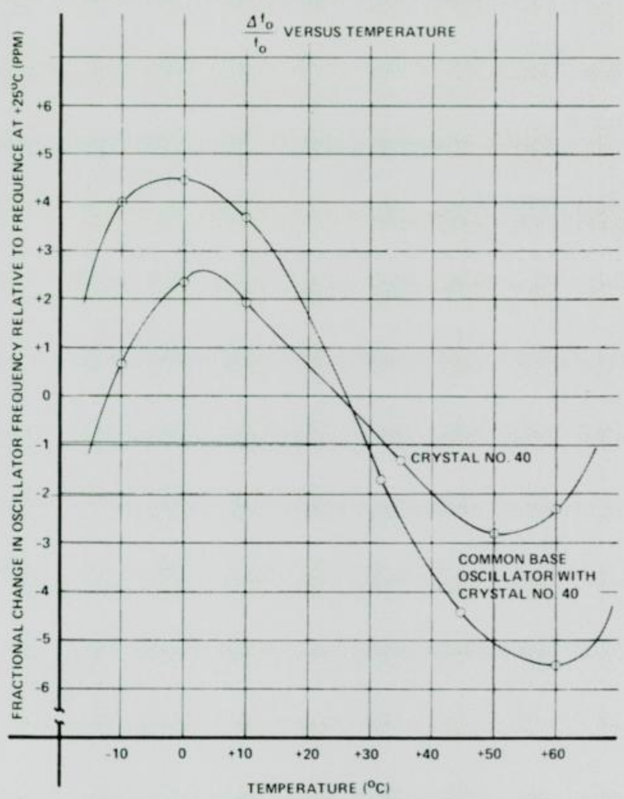

Figure 5. Crystal Oscillator Performance Over Temperature.

An additional problem common to many oscillator configurations is that of output buffering, where the output load change can "pull" frequency. The common base oscillator also has this problem.

Alternatives

There are basically three configurations potentially usefull as VHF overtone oscillators:

- Grounded base
- Impedance inverting Pierce
- Complex oscillator

The grounded base has just been discussed. Frerking[2] gives a good trade-off chart (Chapter 7) relating the various performance properties of common oscillator circuits, and for highest stability recommends the impedance inverting Pierce below 75 MHz. It must be noted that at the time this trade-off was performed, bipolar transistors could not achieve the f_T available today. For that reason, the impedance inverting Pierce was further investigated as a potential candidate using high f_T transistors.

Complex oscillators refer to configurations where the gain and phase of a feedback amplifier (generally multistage) are controlled by the use of the complex circuitry. This is generally prohibitive in miniature equipment due to the required real estate; however, it could be a viable approach on large ground-based equipment. For typical applications, the Pierce appears to be the best choice.

The Impedance Inverting Pierce

Since the design equations for the Pierce (as presented by Frerking) indicate the basic feedback scheme is in-

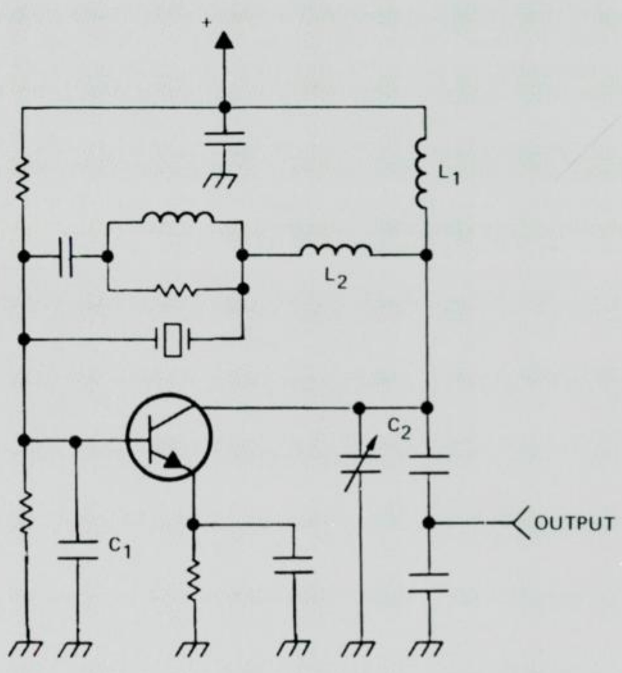

Figure 6. Typical Impedance Inverting Overtone Pierce.

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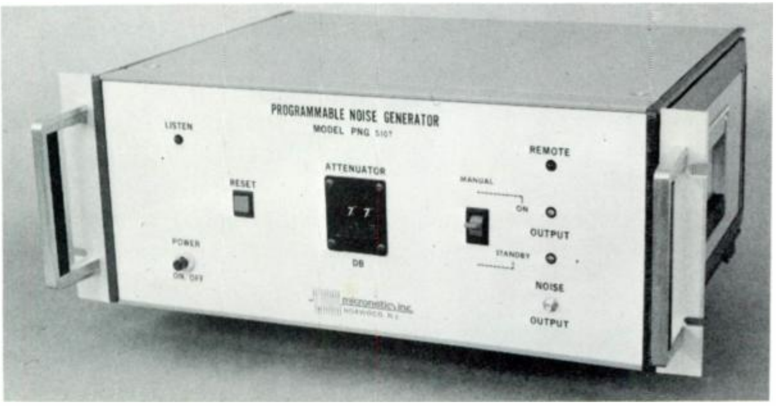

SPECIFICATION

MODEL NO.	FREQUENCY RANGE	FLATNESS	VSWR	RF OUTPUT DBM/Hz
PNG 5101	10Hz-20KHz	±0.5 DB	1.5:1	−33
PNG 5102	10Hz-100KHz	±0.5 DB	1.5:1	−40
PNG 5103	10Hz-500KHz	±0.5 DB	1.5:1	−47
PNG 5104	100Hz-3MHz	±0.75 DB	1.5:1	−55
PNG 5105	100Hz-10MHz	±1.00 DB	1.5:1	−60
PNG 5106	100Hz-25MHz	±1.25 DB	1.5:1	−64
PNG 5107	100Hz-100MHz	±1.50 DB	1.5:1	−70
PNG 5108	1MHz-300MHz	±2.0 DB	1.5:1	−75
PNG 5109	30MHz-500MHz	±2.5 DB	1.5:1	−77
PNG 5110	300MHz-1GHz	±2.5 DB	1.5:1	−79
PNG 5111	1GHz-2GHz	±2.5 DB	2.0:1	−80
PNG 5200	100Hz-1000MHz	±2.5 DB	2.0:1	−80

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NOD 5000 SERIES MANUALLY OPERATED

MODEL NO.	FREQ. RANGE	FLATNESS	VSWR	RF OUTPUT*
NOD 5101	10 Hz-20 KHz	±0.5 DB	1.5:1	-33 DBM/Hz
NOD 5102	10 Hz-100 KHz	±0.5 DB	1.5:1	-40 DBM/Hz
NOD 5103	10 HZ-500 KHz	±0.5 DB	1.5:1	-47 DBM/Hz
NOD 5104	100 Hz-3 MHz	±0.75 DB	1.5:1	-55 DBM/Hz
NOD 5105	100 Hz-10 MHz	±1.00 DB	1.5:1	-60 DBM/Hz
NOD 5106	100 Hz-25 MHz	±1.25 DB	1.5:1	-64 DBM/Hz
NOD 5107	100 Hz-100 MHz	±1.50 DB	1.5:1	-70 DBM/Hz
NOD 5108	1 MHz-300 MHz	±2.0 DB	1.5:1	-75 DBM/Hz
NOD 5109	30 MHz-500 MHz	±2.5 DB	1.5:1	-77 DBM/Hz
NOD 5110	300 MHz-1 GHz	±2.5 DB	1.5:1	-79 DBM/Hz
NOD 5200	100 Hz-1GHz	±2.5 DB	1.5:1	-80 DBM/Hz

COMMON SPECIFICATIONS

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OPERATING TEMPERATURE:	+0°C to +65°C
TEMPERATURE SENSITIVITY:	LESS THAN 0.01 DB/°C
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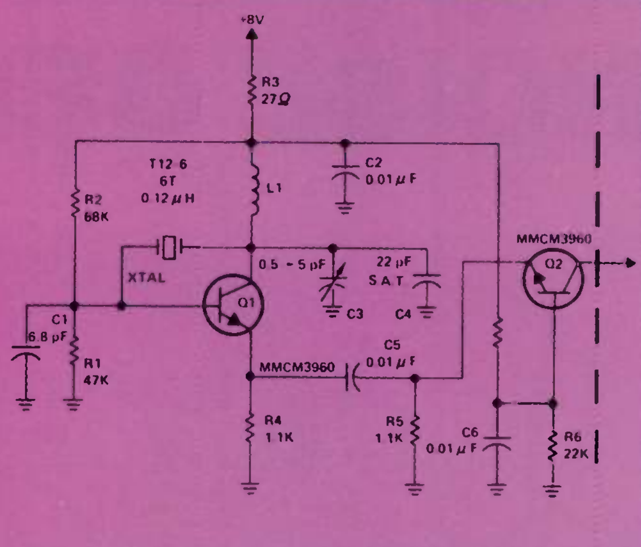


Figure 7. Modified Pierce Oscillator With Buffer.

herently stable, a new circuit has been configured to eliminate the known shortcomings of the Pierce. These shortcomings are:

- Feedback requires a complex crystal phasing circuit (see Figure 6).
- The output is not well-buffered.
- The circuit has a tendency to free run.

All of these shortcomings have been eliminated with a new circuit configuration as shown in Figure 7. Notice that the output network is not shown, but may take any form desired by the user. In this configuration, the crystal is operated above series resonance, i.e., it is inductive. Computer and laboratory results indicate that a 5th overtone crystal should operate approximately 25 ppm above its series resonance frequency.

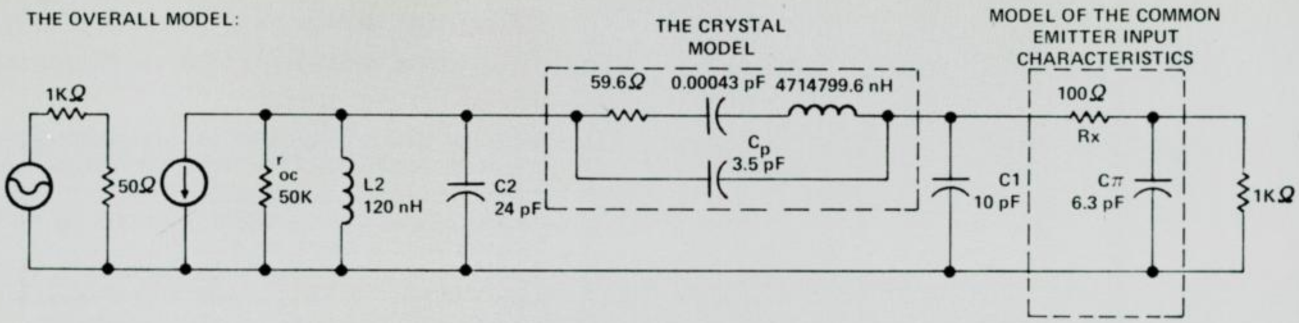
The collector current is approximately 1.5 mA; 1 mA is about the lower limit, since the transistor gm will degrade below this value. However, the supply voltage can be decreased to about 4 volts with no major degradations.

Notice that there is no emitter bypass, as would normally be the case. The emitter is bypassed through the low impedance of the common base buffer stage. In this way, the oscillator AC current appears at the buffer collector and provides almost perfect isolation. No frequency determining circuitry is associated with the output.

Figure 8 shows the computer model used to predict sensitivities. Note the drastic reduction in sensitivities to component changes as shown in Table II. This improvement also manifests itself in the temperature performance shown

Table II. Component Sensitivities in Impedance Inverting Pierce

Component	Nominal Value	f / Value ppm / %
C2	24.0 pF	-0.370
C1	10.0 pF	-0.045
C	6.3 pF	-0.038
C	3.5 pF	-0.073
L2	120.0 nH	-0.250



THE COMMON EMITTER INPUT CHARACTERISTIC:

Rx REPRESENTS THE BASE SPREADING RESISTANCE WHICH IS APPROXIMATELY 100Ω

Cπ REPRESENTS THE BASE-TO-EMITTER CAPACITANCE

$$C\pi = \frac{1}{2\pi f\beta r\pi} - C\mu = \frac{gm}{2\pi f_T} - C\mu \text{ , SINCE } f_\beta = \frac{f_T}{\beta o} \text{ AND } r\pi = \frac{\beta o}{m}$$

$$= \frac{38.9}{2\pi \times 10^9 \times 10^3} \text{ SINCE } gm = 38.9\, I_c \qquad C\mu \ll C\pi \text{ AND } I_c \cong 1 \text{ mA}$$

$$C\pi \cong 6.3 \text{ pF}$$

Figure 8. Modified Pierce Sensitivity Analysis.

in Figure 9. This close tracking was observed in all of the oscillators tested and does not require temperature compensation.

In order to use this configuration, the crystal series resonance must be specified approximately 25 ppm lower than the desired operating frequency. This will vary somewhat as the motional capacitance associated with the crystal model varies, however, 25 ppm has been found to be typical.

In the oscillators tested, no intentional negative temperature coefficient components were used, however, good quality components such as NPO capacitors and low permeability inductors were used.

(Continued on page 36.)

6 Solutions
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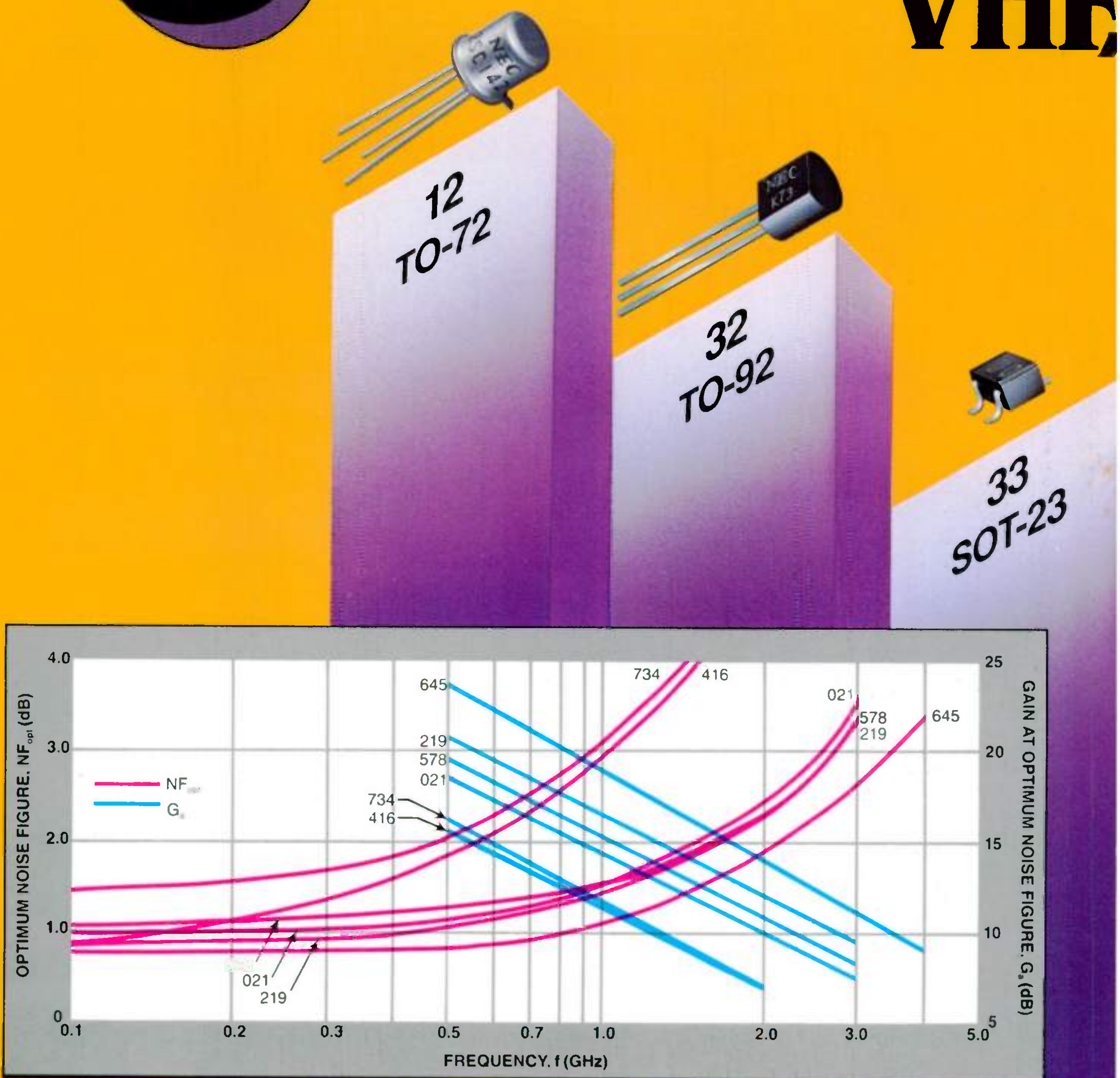

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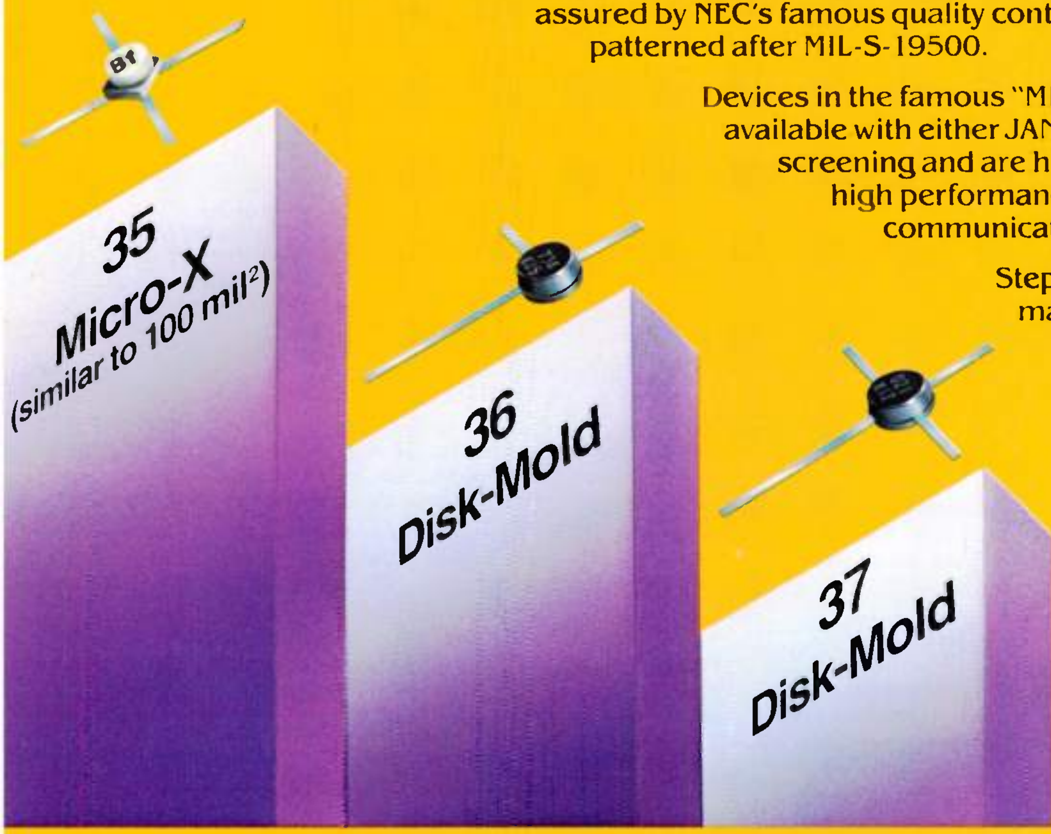

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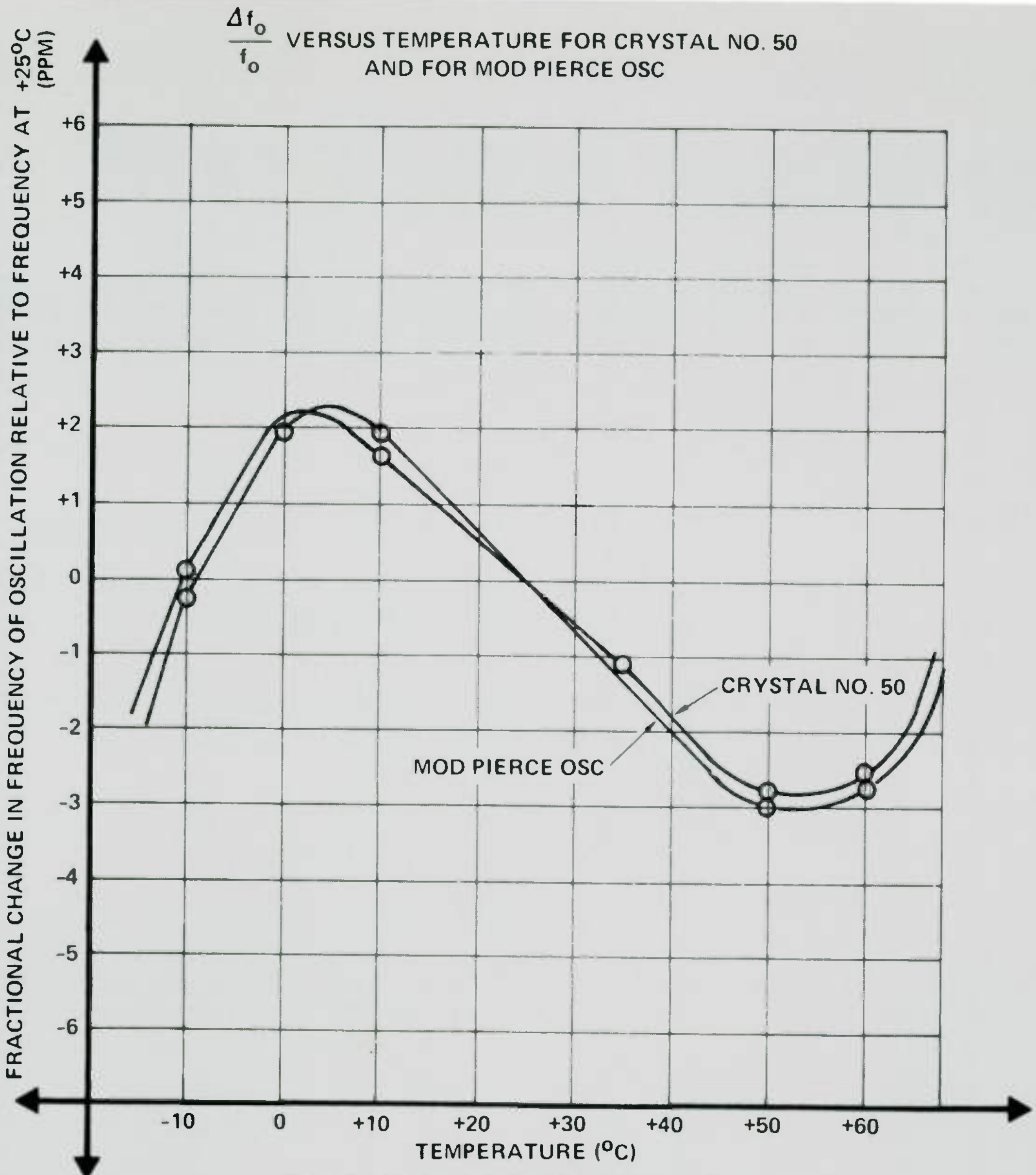

Figure 9. Temperature Performance

Conclusion

In conclusion, a modification of the basic Pierce oscillator has been developed, which provides stable performance, is free from spurious oscillations, and requires simple circuitry.

Acknowledgements

The author wishes to acknowledge helpful conversations with Marvin Frerking of Rockwell International, Cedar Rapids, Iowa and the computer programming assistance of Donna Wilt and Don Stickels of Harris Government Systems Group. The author also expresses appreciation to Harris Corp. for supporting this work.

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Author's Background

Mr. Belcher is on the Technical Staff of the Signal Processing Section of Harris Government Electronic Systems Division in Melbourne, Florida. His principal responsibility is to provide technical guidance to a variety of RF communication projects. Mr. Belcher has over fourteen years of design experience in a broad category of products ranging from high rel to commercial. He has contributed to various publications and has authorized six patents.

Mr. Belcher has received the BSEE from the University of Kentucky and MSE from the University of Florida. He is a member of HKN, holds a first class radio telephone and amateur advanced license and is a Registered Professional Engineer.

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

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Multiple-Tuned Amplifiers

By Jack Porter
Sr. Technical Specialist
Defense Systems Division
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San Diego, CA

Double-tuned amplifiers were described by Aiken[1, 2] in terms of coupling index S and the Q_1/Q_2 ratio—ρ. In these terms, the Butterworth response corresponds to S = 1 and ρ = 1. Chebyshev responses with various degrees of ripple result from over-coupling (S > 1) with ρ = 1. S = 1 and ρ = 3.732 produces the Bessell filter response.

Among the many possible triple-tuned response shapes described by Mather[3], only those corresponding to conventional Butterworth, Chebyshev, etc., responses are of practical significance.

For purposes of analysis and design, a tuned amplifier is a network like that shown in Figure 1. A transistor, described by Y-parameters Y_a is connected thru a coupling network to a load which may be the input admittance of another amplifier stage. The coupling network is a bandpass filter containing N tuned circuits. This filter may be of any type; that shown in Figure 1 was chosen because it is commonly used and has been frequently described in the literature[4-6]. Since the filter can contain any reasonable number of tuned circuits, the amplifier is best described as multiple-tuned.

The amplifier design consists of four main steps:

1. Select a transistor, choose input/output mismatch ratios, calculate stability factors and input/output admittances.
2. Select a filter type, find the low-pass prototype filter element values, then calculate bandpass filter element values for the specified center frequency and bandwidth.
3. Calculate values for turns ratios n_1 and n_2 and loading resistors R_1 and R_2 which simultaneously provide the chosen transistor mismatch ratios and the calculated filter termination admittances.
4. Calculate the power gain of the amplifier stage.

Transistor Admittances and Stability Factors

The transistor input and output admittances, Stern stability factor, and alignment sensitivity are calculated from the transistor Y-parameters. The filter input admittance and the conductance of the loading resistance are simply admittances added in parallel with y_{11} and y_{22} in these calculations, which makes the calculations much simpler than they would be if scattering parameters or any other two-port network parameters were used. Although the active device is referred to as a "transistor", it may actually be an integrated circuit such as the MC1590, CA3028, or any other suitable device characterized by admittance parameters. The input and output mismatch ratios are defined as

$$m_i = \frac{Re(Y_{Li})}{Re(y_{11})} \text{ and } m_o = \frac{Re(Y_{Lo})}{Re(y_{22})}$$

where Y_{Li} and Y_{Lo} are the total admittances connected across the transistor input and output, respectively. In all of the following calculations it is assumed that the input and output circuits are tuned to resonance, so that $Im(Y_{Li}) \approx -Im(y_{11})$ and $Im(Y_{Lo}) \approx -Im(y_{22})$.

The Stern stability factor — in terms of the transistor parameters and mismatch ratios is

$$K = \frac{2(m_i + 1)[Re(y_{11})](m_o + 1)[Re(y_{22})]}{|y_{12}y_{21}| + Re(y_{12}y_{21})} \qquad (1)$$

An amplifier is unstable if K< 1. For good stability K > 7 is desirable. The alignment sensitivity[8] is

$$\alpha = \frac{|y_{12}y_{21}|}{|(m_i + 1)[Re(y_{11})](m_o + 1)[Re(y_{22})] - y_{12}y_{21}|} \qquad (2)$$

It is approximately the percentage change in total input circuit admittance $Y_{Li} + y_{11}$ which results from a one percent change in Y_{Lo}. The condition for satisfactory alignability is $\alpha < .5$.

A tuned amplifier basically consists of an amplifying device (transistor or integrated circuit) coupled to a load impedance by a bandpass filter. Because the theory of tuned amplifiers was developed a number of years before most of modern filter theory, amplifiers have traditionally been described as single-tuned, double-tuned or triple-tuned and the design of these amplifiers has not been related to modern filter design.

Values of the mismatch ratio m are usually in the range 4-20, but for devices with very high gain or very low output admittance the value may be as large as 100. The value of m may have to be increased to make a high Q circuit physically realizable, i.e. to prevent the transistor from loading it excessively. This condition becomes evident when the calculated value of the loading resistor is negative.

When satisfactory mismatch ratios have been selected, the transistor input and output admittances can be calculated. The input admittance is

$$Y_i = y_{11} - \frac{y_{12}y_{21}}{(m_o + 1)\,Re(y_{22})} \tag{3}$$

and the output admittance is

$$Y_o = y_{22} - \frac{y_{12}y_{21}}{(m_i + 1)\,Re(y_{11})} \tag{4}$$

Designing the Filter

The filter shape and number of poles required to satisfy the attenuation and/or group delay specifications are usually determined from published frequency response curves for low-pass prototype filters[4-6]. For a narrowband filter with center frequency f_o and bandwidth B_o it is usually sufficiently accurate to assume that the filter is symmetrical around f_o and that the attenuation at any frequency f_1 is that shown on the graph at the normalized frequency $\omega = 2|f_1 - f_o|/B_o$.

When a filter has been selected, the prototype element values must be looked up in a table or calculated. Formulas for calculating Butterworth and Chebyshev prototype element values can be found in several of the references[4, 6, 9, 10]; they are summarized in Figure 2. The Chebyshev element values calculated from these formulas are for a low-pass prototype filter with a ripple bandwidth of one radian/second. To scale these values for a 3dB bandwidth filter, multiply each of the elements g_1 thru g_N by ω_H.

Another type of filter which is useful in multiple-tuned amplifiers is the normalized Bessel filter, also called the

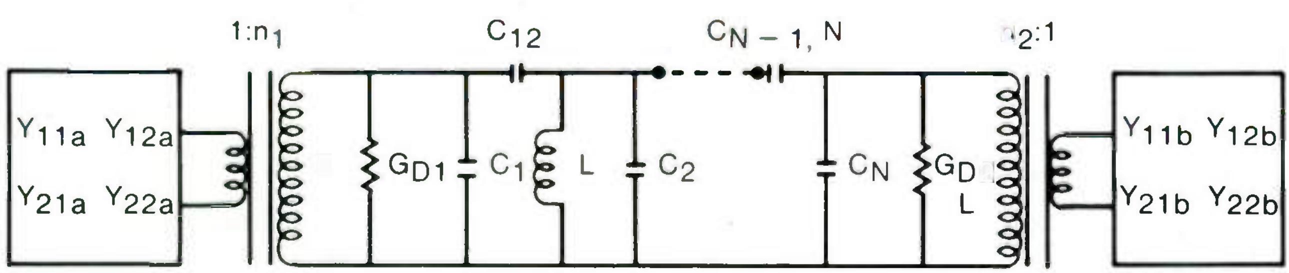
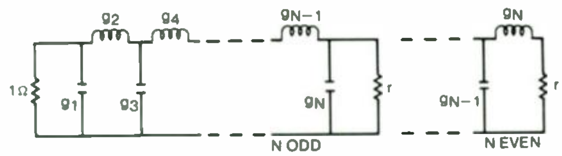


Figure 1



N-POLE LOW-PASS PROTOTYPE FILTER

BUTTERWORTH FILTERS:

$$g_K = 2\sin\left[\frac{(2K-1)\pi}{2N}\right] \qquad K = 1, 2, \ldots, N$$

$$|H(j\omega)|^2 = \frac{1}{1+\omega^{2N}} \qquad r = 1$$

CHEBYSHEV FILTERS: A_m: PASSBAND RIPPLE IN dB

$$|H(j\omega)|^2 = \frac{1}{1+\epsilon^2 T_N^2(\omega)} \qquad \epsilon = \sqrt{10^{A_m/10}-1}$$

$$\beta = 2\sinh^{-1}\left(\frac{1}{\epsilon}\right) = \ln\left(1+\frac{2}{10^{A_m/20}-1}\right) = \ln\coth\left(\frac{A_m \ln 10}{40}\right)$$

$$\gamma = \sinh\left(\frac{\beta}{2N}\right) \qquad a_K = \sin\left[\frac{(2K-1)\pi}{2N}\right] \qquad K = 1, 2, \ldots, N$$

$$b_K = \gamma^2 + \sin^2\left(\frac{K\pi}{N}\right) \qquad K = 1, 2, \ldots, N$$

$$g_1 = \frac{2a_1}{\gamma} \qquad g_K = \frac{4a_{K-1}a_K}{b_{K-1}g_{K-1}} \qquad K = 2, 3, \ldots, N$$

$$r = 1 \text{ FOR N ODD} \qquad r = \tanh^2\left(\frac{\beta}{4}\right) \text{ FOR N EVEN}$$

3dB BANDWIDTH:

$$\omega_H = \cosh\left[\left(\frac{1}{N}\right)\cosh^{-1}\left(\frac{\sqrt{A_H-1}}{\epsilon}\right)\right] \qquad A_H = 2 \text{ FOR N ODD}, \quad A_H = 2(1+\epsilon^2) \text{ FOR N EVEN}$$

Figure 2.

maximally-flat delay or Thomson linear-phase filter. Data on it can be found in Zverev[5]. Bessel filter element values which have been scaled to produce a 1 rad/sec-3dB bandwidth are shown in Table I for N = 1 thru N = 8. The resistive termination is r = 1 for this filter.

When the prototype element values have been found, the bandpass filter values can be calculated. The first step is to find the corrected center frequency f_c. Since the standard low-pass to bandpass transformation results in a geometrically symmetrical passband, the design center frequency f_c should be the geometric center of it, thus

$$f_c = \sqrt{\left(f_0 - \frac{B_0}{2}\right)\left(f_0 + \frac{B_0}{2}\right)} \qquad \text{or}$$

$$f_c = \sqrt{f_0^2 - \left(\frac{B_0}{2}\right)^2} \qquad (5)$$

Then $\omega_0 = 2\pi f_c$ (6),
and the total filter Q, Q_0 can be defined as $Q_0 = f_c/B_0$ (7).
The input circuit Q is
$Q_1 = Q_0 g_1$ (8a),
and the output circuit Q is
$Q_2 = Q_0 g_N/r$ (8b).
Butterworth and Chebyshev filters, including those of even order, have the property $g_N = g_1 r$, thus $Q_1 = Q_2$ for these filters. This isn't true of Bessel filters, nor of most other types of all-pole filters.

The next step is to select the filter inductance L in the bandpass filter shown in Figure 1 and calculate the total node capacitance C_T or vice versa. Since these circuits are usually tuned by variable capcitors, it's more convenient to select C_T and calculate

$$L = \frac{1}{\omega_0^2 C_T} \qquad (9).$$

The filter input and output termination admittances are

$$G_{T1} = \frac{\omega_0 C_T}{Q_1} \qquad (10a)$$

and $$G_{T2} = \frac{\omega_0 C_T}{Q_2} \qquad (10b)$$

The series capacitors are

$$C_{K,K+1} = \frac{C_T}{Q_0\sqrt{g_K g_{K+1}}} \qquad K = 1, 2, \ldots, N-1 \qquad (11)$$

Since the values of C_1 and C_N depend slightly on the turns ratios, the shunt capacitance values aren't calculated until later.

Turns Ratios and Loading Resistors

The condition for proper filter input termination is

$$G_{T1} = G_u + G_{D1} + \frac{Re(Y_{0a})}{n_1^2}$$

and that for achieving the specified transistor output mismatch ratio is

$$m_{0a} Re(y_{22a}) = n_1^2 (G_u + G_{D1} + G_{T1})$$

In these equations G_u is the equivalent conductance due to the finite unloaded Q of the circuit. When good quality capacitors are used Q_u, the unloaded circuit Q, is essentially equal to the inductor Q.

Element Values

1	2.0000			
2	0.57550	2.1478		
3	0.33742	0.97051	2.2034	
4	0.23342	0.67252	1.0815	2.2404
5	0.17432	0.50724	0.80401	1.1110
	2.2582			
6	0.13649	0.40019	0.63916	0.85379
	1.1126	2.2645		
7	0.11056	0.32589	0.52489	0.70201
	0.86903	1.1052	2.2659	
8	9.1905E-02	0.27191	0.44092	0.59357
	0.73026	0.86950	1.0956	2.2656

Table I.

Low-Pass Prototype Element Values For Normalized Bessel Filter.

$$G_u = \frac{\omega_0 C_T}{Q_u} = \left(\frac{Q_1}{Q_u}\right) G_{T1} = \left(\frac{Q_2}{Q_u}\right) G_{T2}$$

Solving these equations for n_1 and G_{D1} results in

$$n_1 = \sqrt{\frac{m_{oa}[Re(y_{22a})] + Re(Y_{oa})}{2G_{T1}}} \qquad (12)$$

and

$$G_{D1} = G_{T1}(1 - \frac{Q_1}{Q_u}) - \frac{Re(Y_{0a})}{n_1^2} \qquad (13)$$

similarly, solving

$$G_{T2} = G_u + G_{D2} + \frac{Re(Y_{ib})}{n_2^2}$$

and

$$m_{ib}Re(y_{11b}) = n_2^2(G_u + G_{D2} + G_{T2})$$

provides the filter output circuit equations

$$n_2 = \sqrt{\frac{m_{ib}[Re(y_{11b})] + Re(Y_{ib})}{2G_{T2}}} \qquad (14)$$

and

$$G_{D2} = G_{T2}(1 - \frac{Q_2}{Q_u}) - \frac{Re(Y_{ib})}{n_2^2} \qquad (15)$$

The values of the loading resistors are

$$R_1 = \frac{1}{G_{D1}} \qquad (16a)$$

and

$$R_2 = \frac{1}{G_{D2}} \qquad (16b)$$

The transformers shown in Figure 1 are assumed to be ideal transformers with unity coefficient of coupling when calculating the turns ratios. This is a good approximation at lower frequencies when powdered iron or ferrite transformer cores are used, but at higher frequencies with very low permeability cores and few turns of wire, the values of n found experimentally are usually lower than the calculated values.

The filter shunt capacitors can now be calculated.

$$C_1 = C_T - C_{12} - \frac{I_m(Y_{0a})}{\omega_0 n_1^2} \qquad (17a)$$

$$C_N = C_T - C_{N-1,N} - \frac{I_m(Y_{ib})}{\omega_0 n_2^2} \qquad (17b)$$

$$C_K = C_T - C_{K-1,K} - C_{K,K+1} \quad K = 2,3,\ldots,N-1 \qquad (17C)$$

The above equations 12 thru 17 assume that the filter is used to couple two amplifier stages. However, the filter may also be used as an input or output network which couples the amplifier to a resistive load.

For an input network, m_{oa} in equation 12 becomes m_s, the source mismatch ratio. If the source resistance is stable and well defined it is desirable to make m_s as nearly equal to one as possible, but if it is made too small the source resistance loads the tuned circuit so much that the specified Q_1 can't be attained. In this case the calculated value of R_1 is negative. If the source resistance isn't stable, the actual value of Q_1 can be quite different from the design value. In this case m_s should be increased, which causes the circuit Q to be more dependent on the value of R_1.

For an input filter the admittances Re (y_{22a}) and Re (Y_{oa}) in equations 12 and 13 become the source conductance G_s. Thus

$$n_1 = \sqrt{\frac{(m_s + 1)G_s}{2G_{T1}}} \qquad (12a)$$

$$G_{D1} = G_{T1}(1 - \frac{Q_1}{Q_u}) - \frac{G_s}{n_1^2} \qquad (13a)$$

and $Im(Y_{oa}) = 0$ in equation 17a.

If the filter is used to couple the final state of an amplifier to a resistive load conductance G_L with a mismatch ratio m_L,

$$n_2 = \sqrt{\frac{(m_L + 1)G_L}{2G_{T2}}} \qquad (14a)$$

$$G_{D2} = G_{T2}(1 - \frac{Q_2}{Q_u}) - \frac{G_L}{n_2^2} \qquad (15a)$$

and $Im(Y_{1b}) = 0$.

Power gain

The power gain in dB of the amplifier stage is

$$A_p = A'_p - A_L \qquad (18).$$

A'_p is the gain of the amplifier if the unloaded tuned circuits (all except the first and last) have infinite Q. The loss of these circuits caused by finite Q is approximated by A_L.

$$A'_p = 10\log\left\{\left|\frac{y_{21a}}{(m_{oa}+1)[Re(y_{22a})]}\right|^2\left(\frac{n_1^2}{n_2^2}\right)\left(\frac{G_{T1}}{G_{T2}}\right)\left[\frac{Re(Y_{ib})}{Re(Y_{ia})}\right]\right\} \qquad (19)$$

$A_L = 0$ for $N = 1$ and $N = 2$, since the loading conductances G_{D1} and G_{D2} are reduced to compensate for the effective resistance caused by finite circuit Q_u. It becomes part of the required loading resistance.

The effect of finite circuit Q in filters with $N > 2$ is to increase the filter loss and produce a more rounded passband than an ideal filter has. This passband rounding also decreases the bandwidth. The magnitude of these effects is obtained from pseudo-exact filter design theory[11, 12].

From an approximation by Cohn[11],

$$A_L = \frac{10}{\ln 10}\left(\frac{Q_o}{Q_u}\right)\sum_{K=2}^{N-1} g_k \qquad (20)$$

The articles by Sleven[12] contain a great deal of information about the effects of finite circuit Q on the filter passband and stopband. To be strictly accurate, in the case of an even order Chebyshev filter the ripple amplitude A_m should be subtracted from the gain calculated by equations 18 thru 20, but this small correction can usually be ignored.

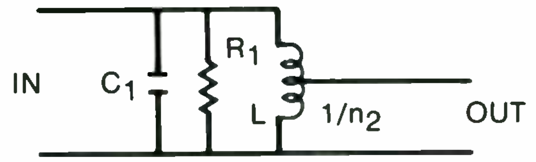


Figure 3a

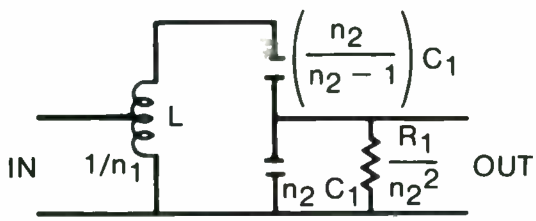
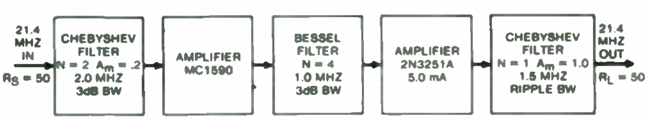


Figure 3b

Equation 19 in the form shown above applies to an interstage network. For a final amplifier stage which is terminated in a load conductance G_L, the term Re (Y_{ib}) in (19) should be replaced by G_L.

For an input filter which couples a source conductance G_s to the first stage of an amplifier, it's more useful to calculate the transducer gain A_T. The transducer gain is the ratio of the power which the input circuit delivers to the transistor input admittance when the power is supplied from a generator with source admittance G_s, divided by the power which the generator could deliver to a matched load G_s. The transducer gain is



EXAMPLE TUNED AMPLIFIER Figure 4

$$A_T = A'_T - A_L \qquad (21),$$

and

$$A'_T = 10\log\left\{\left(\frac{2}{m_{s+1}}\right)^2\left(\frac{n_1^2}{n_2^2}\right)\left(\frac{G_{T1}}{G_{T2}}\right)\left[\frac{Re(Y_{ib})}{G_s}\right]\right\} \qquad (22)$$

Since this is the gain of a passive network, $A_T < 0$ in all cases.

If the first stage transistor with input admittance Y_i is driven directly from a source with admittance G_s, with no matching network,

$$A_T = 10\log\left\{\frac{4\,G_s[Re(Y_i)]}{|G_s + Y_i|^2}\right\} \qquad (23)$$

If the power gains of all amplifier stages are summed and added to the transducer gain of the input network, the result is the transducer gain of the entire amplifier.

Single-Tuned Amplifiers

Equations (5) thru (20) are also valid for N = 1 with some minor changes. From equations 8a and 8b, $Q_2 = Q_1$ since $g_N = g_1$. Equation 11 isn't applicable since there are no series capacitors. The loading resistors are combined into one, and equation 16 becomes

$$\frac{1}{R_1} = G_{D1} + G_{D2} + \frac{\omega_o C_T}{Q_u} \qquad (16c).$$

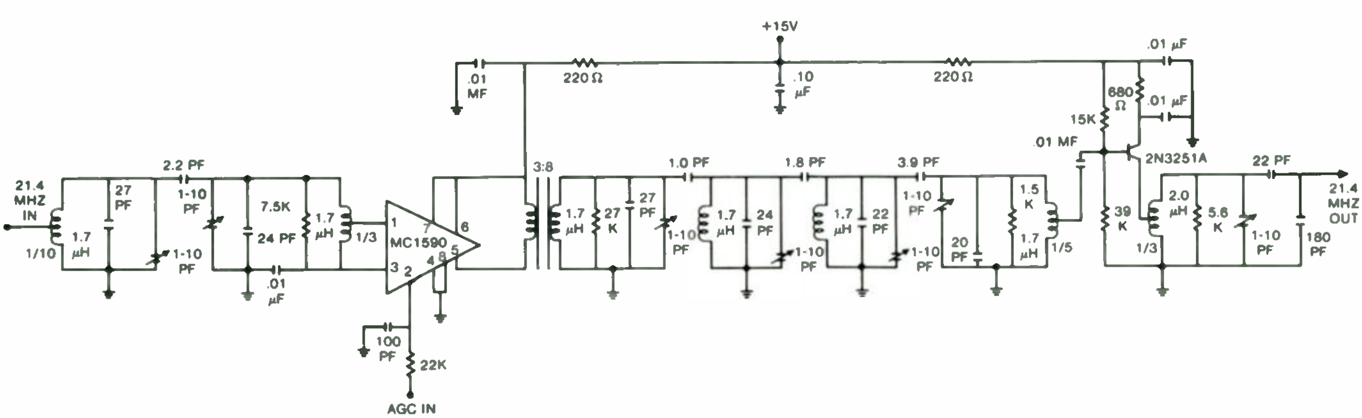
The actual loaded Q of the single-tuned circuit is

$$Q_L = \frac{Q_1}{2} \qquad (24).$$

The tuning capacitor value is

$$C_1 = C_T - \frac{Im(Y_{oa})}{\omega_o n_1^2} - \frac{Im(Y_{ib})}{\omega_o n_2^2} \qquad (17d)$$

Connecting the input and output to the same inductor usually produces unsatisfactory results unless $n_1 = 1$ as shown in Figure 3a.



21.4 MHZ TUNED AMPLIFIER Figure 5

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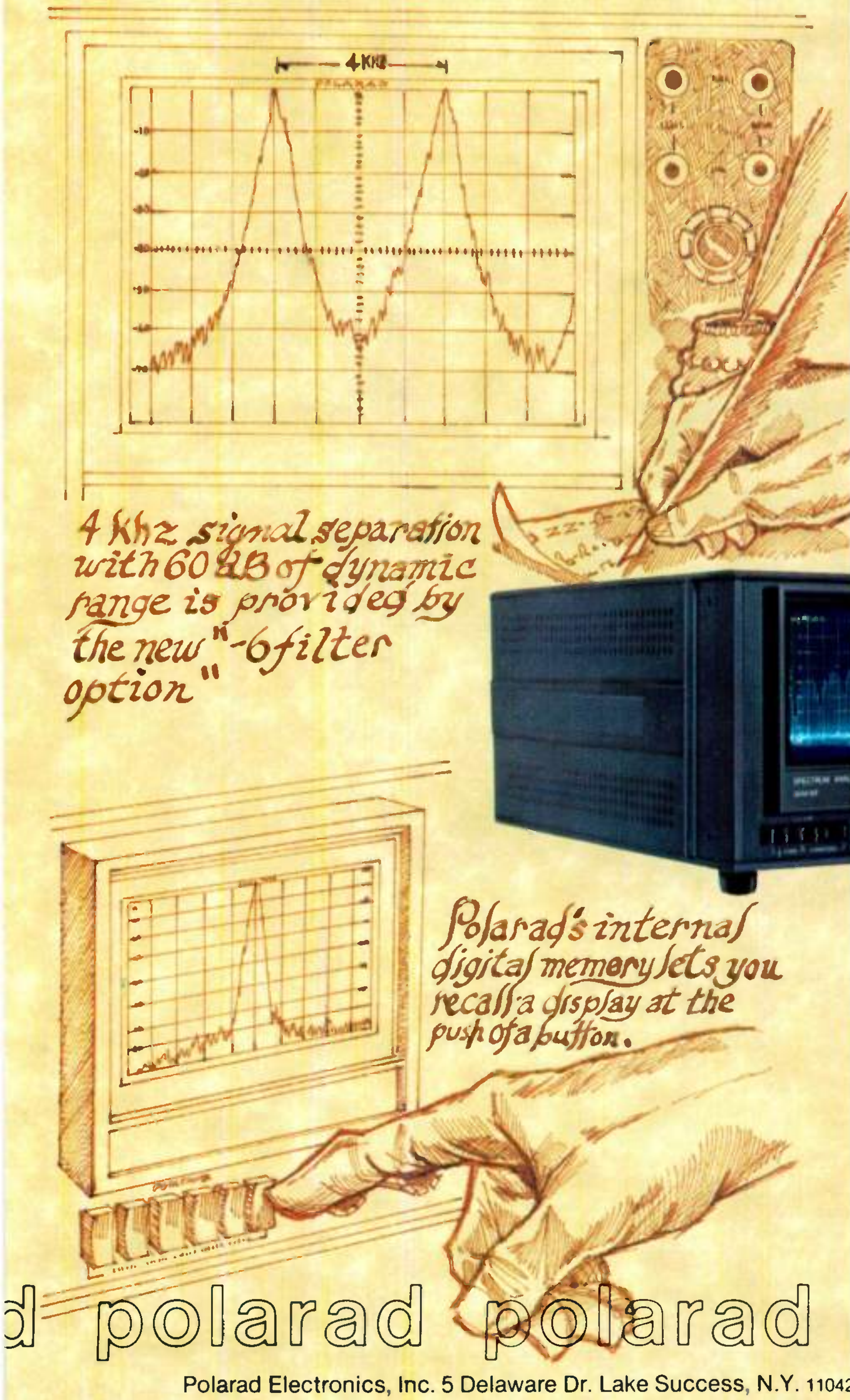

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Symbol Table

Ap	Power gain (dB).
A_L	Loss due to finite circuit Q.
A_T	Transducer gain (dB).
Bo	Bandwidth (Hz).
C_k	Tuning capcitance of k'th tuned circuit.
$C_{k}, K + 1$	Mutual capacitance between circuits k and k + 1.
C_T	Total capacitance at each node.
f_o	Design center frequency (Hz).
f_c	Corrected center frequency used in calculations.
G_{D1}, G_{D2}	Input and output circuit loading conductances.
G_s, G_L	Source and load admittances.
G_{T1}, G_{T2}	Filter termination conductances.
g_k	k'th prototype element value.
Im(y)	Imaginary part of admittance y.
K	Stern stability factor.
$K_i, i + 1$	Coupling coefficient between circuits i and i + 1.
L	Tuned circuit inductance.
m_{ia}, m_{oa}	Input and output mismatch ratios for transistor a.
m_{ib}, m_{ob}	Mismatch ratios for transistor b.
N	Number of tuned circuits.
n_1, n_2	Input and output turns ratios.
Q_o	Total filter Q. $Q_o = f_c/B_o$.
Q_1, Q_2	Input and output tuned circuit Q's.
Q_L	Loaded Q of single-tuned circuit.
Q_u	Unloaded circuit Q; approximately equal to inductor Q.
r	Low-pass prototype filter termination. r is a resistance value for N even, a conductance value for N odd.
R_1, R_2	Input and output loading resistor values.
Re(y)	Real part of admittance y.
$y_{11}, y_{12}, y_{21}, y_{22}$	Transistor short-circuit admittance parameters.
Y_i, Y_o	Transistor input and output admittances.
α	Alignment sensitivity (S in the computer program).
ω_o	Frequency used in calculations (rad/sec). $\omega_o = 2\pi f_c$.

Table II.

N = 4
PROTOTYPE ELEMENT VALUES:
1 2.2404
2 1.0815
3 .87252
4 .923342
r = 1
TRANSISTORS:
1. MC1590 21.4 MHz
Y11 = +3.66E-04 +8.13E-04j
Y12 = −1.38E-08 −3.58E-06j
Y21 = +1.43E-01 −8.77E-02j
Y22 = +5.89E-05 +3.99E-04j

2.2N3251A 5 mA 21.4 MHz
Y11 = +5.34E-03 +4.52E-03j
Y12 = −7.33E-07 −2.86E-04j
Y21 = +9.04E-02 −7.05E-04j
Y22 = +4.13E-04 +6.49E-04j
Mia = 10 Moa = 20
Mib = 7 Mob = 10
Ka = 34.9932 Sa = .11288
Kb = 15.1697 Sb = .131876
Yia = +6.21E-04 +1.23E-03j
Yog = +1.37E-04 + 5.26E-04j
Yib = +5.40E-03 +1.02E-02j
Yob = +4.19E-04 +1.25E-03j

Initial		Final	
N = 4		N = 4	
F0 = 21.4 MHz		F0 = 21.4 MHz	
Fc = 21.3942 MHz		Fc = 21.3942 MHz	
BW = 1 MHz		BW = 1 MHz	
AP = 35.5559 dB		AP = 35.5559 dB	
Q1 = 47.9315 Q2 • 4.99382		Q1 = 47.9315 Q2 = 4.99382	
Qu = 120		Qu = 120	
R1 = 30344.2		R1 = 27535.6	
R2 = 1754.08		R2 = 1591.72	
N1 = 2.7959 N2 = 5.14658		N1 = 2.66337 N2 = 4.90262	
L = 1.84471E-06		L = 1.67397E-06	
CT = 3E-11		CT = 3.306E-11	
C 1 = 2.860E-11		C1 = 3.152E-11	
C 2 = 2.745E-11	C12 = 9.008E-13	C 2 = 3.026E-11	C12 = 9.927E-13
C 3 = 2.482E-11	C23 = 1.644E-12	C 3 = 2.735E-11	C23 = 1.812E-12
C 4 = 2.359E-11	C34 = 3.539E-12	C 4 = 2.600E-11	C34 = 3.900E-12

Interstate Network.

Table III.

N = 2	GS = .02 BS = 0	GS = .02 BS = 0
Am = .2 dB	MS = 1	MS = 1.5
3 dB BW	Mib = 10 Mob = 20	Mib = 10 Mob = 20
PROTOTYPE ELEMENT	K = 34.9932 S = .11288	K = 34.9932 S = .11288
VALUES:	Yib = +6.21E-04 +1.23E-03j	Yib = +6.21E-04 +1.23E-03j
1 1.7702	Yob = +1.37E-04 +5.26E-04j	Yob = +1.37E-04 +5.26E-04j
2 1.15056	N = 2	N = 2
r = .649961	AM = .2 dB	Am = .2 dB
TRANSISTOR:	3 dB BW	3 dB BW
MC1590 21.4 MHz	F0 = 21.4 MHz	F0 = 21.4 MHz
Y11 = +3.66E-04 +8.13E-04j	Fc = 21.3766 MHz	Fc = 21.3766 MHz
Y12 = −1.36E-08 −3.58E-06j	BW = 2 MHz	BW = 2 MHz
Y21 = +1.43E-01 −8.77E-02j	AP (Transducer) = −5.57168 dB	AP(Transducer) = −6.54078 dB
Y22 = +5.89E-05 +3.99E-04j	Q1 = 18.9205 Q2 = 18.9204	Q1 = 18.9205 Q2 = 18.9204
	Qu = 120	Qu = 120
	R1 = 58571.6	R1 = 99167.7
	R2 = 16728.9	R2 = 7604.03
	N1 = 13.5904 N2 = 4.44629	N1 = 10.2442 N2 = 2.99769
	L = 3.63402E-06	L = 1.65183E-06
	CT = 1.52537E-11	CT = 3.35581E-11
	C 1 = 1.425E-11	C 1 = 3.136E-11
	C12 = 1.000E-12	C12 = 2.200E-12
	C 2 = 1.379E-11	C 2 = 3.034E-11
	Initial.	Final.

Input Network. Table IV.

For $n_1 = 1$, G_{T1} is found by rearranging equation 12,

$$G_{T1} = \frac{m_{oa}[\mathrm{Re}(y_{22a})] + \mathrm{Re}(Y_{oa})}{2} \tag{25}$$

and from equation 10a

$$C_T = \frac{Q_1 G_{T1}}{\omega_o} \tag{26}$$

If this results in a capacitance value which is too large, n_1 and n_2 can be calculated in the usual way and the output taken from a capacitive tap as shown in Figure 3b. When a capacitive tap is connected to the input of a transistor the circuit usually oscillates, since the transistor input isn't loaded at low frequencies. To avoid this a loading resistor of value R_1/n_2^2 can be placed between the tap and ground instead of in parallel with the inductor.

Other Coupling Methods

The capacitive coupling between tuned circuits shown in Figure 1 is used as an example, but any of the other coupling methods shown in reference (13) can also be used. For $N > 2$, the total capacitance or inductance of the tuned circuit includes the mutual capacitances or inductances on both sides of the tuned circuit. The coupling coefficients used in the circuits in reference 13 can be calculated from

$$K_{i,i+1} = \frac{1}{Q_o\sqrt{g_i g_{i+1}}} \quad i = 1, 2, \ldots, N-1 \tag{27}$$

All of the symbols used in equations 1 thru 27 are described in Table II.

N = 1	N = 1	N = 1
Am = 1 dB	Am = 1 dB	Am = 1 dB
RIPPLE BW	RIPPLE BW	RIPPLE BW
PROTOTYPE ELEMENT VALUES:	F0 = 21.4 MHz	F0 = 21.4 MHz
1 1.01789	Fc = 21.3869 MHz	Fc = 21.3869 MHz
r = 1	BW = 1.5 MHz	BW = 1.5 MHz
TRANSISTOR:	AP = 22.2228 dB	AP = 22.2228 dB
2N3251A 5 mA 21.4 MHz	QL = 7.2551 Qu = 120	QL = 7.2551 Qu = 120
Y11 = +5.34E-03 +4.52E-03j	R1 = 632.783	R1 = 5693.12
Y12 = −7.33E-07 −2.86E-04j	N1 = 1 N2 = 2.96523	N1 = 2.99949 N2 = 8.8942
Y21 = +9.04E-02 −7.05E-04j	L = 2.25473-07	L = 2.02854E-06
Y22 = +4.13E-04 +6.49E-04j	CT = 2.45617E-10	CT = 2.73E-11
GL = .02 BL = 0	C 1 = 2.363E-10	C 1 = 2.626E-11
ML = 1		
Mia = 7 Moa = 10	Initial	Final
K = 15.1697 S = .131876		
Yia = +5.40E-03 +1.02E-02j		
Yoa = +4.19E-04 +1.25E-03j		

Output Network. Table V.

An Example Tuned Amplifier Design

Figure 4 is a block diagram of the amplifier which is used as an example. The active devices used in the amplifier are an MC1590 integrated circuit and a 2N3251A transistor. The assumed y-parameters for these devices are listed in Table III. The interstage coupling network, which is a normalized Bessel filter, is designed first.

The prototype element values for N = 4 are obtained from Table I. Since the low-pass prototype filter is linear and passive, it is also reciprocal and thus the input can be applied to either end. Therefore, in designing a bandpass filter based on this prototype, the element values can be taken in the order given in the table or in reverse order. Since $Q_1 = Q_2$ for Butterworth and Chebyshev filters, the same bandpass filter results in either case, but this isn't true for other types of filters. When a Bessel filter is used for interstage coupling, a lower output turns ratio results when the low Q end of the filter is terminated in the low input impedance of the transistor.

The first step is to select mismatch ratios for both stages. For the MC1590 these are 10 and 20; lower ratios of 7 and 10 are chosen for the 2N3251A. The calculations involved in the design, while not very complicated, are tedious, and are best done with a programmable calculator or small computer, especially since some of them may have to be repeated once or twice for each coupling network. The results shown in Tables III thru V were obtained using a Radio Shack TRS-80 (Table VI).

The stability factor, alignment sensitivity, and input and output admittances are calculated first, using equations 1 thru 4. As shown in Table III the alignment sensitivity (designated "S" in the table) was .11 for the first stage and .13 for the second.

The coupling network component values are then calculated using equations 5 thru 20 in order. An unloaded circuit Q of 120 is assumed and C_T = 30 pF is selected. The values obtained are satisfactory except for somewhat awkward values for C_{12}, C_{23} and C_{34}. The calculation is then repeated using a value $C_T = 30 \times 3.9/3.539 = 33.06$ pF, so that C_{34} = 3.9 pF exactly. This produces the final results shown in Table III.

The input network is double-tuned with a 0.2 dB ripple Chebyshev response. Prototype element values are calculated using the equations in Figure 2. An attempt is made to match the 50 ohm input impedance exactly ($m_s = 1$).

To provide a convenient coupling capacitor value, C_{12} = 1.0 pF is chosen and C_T is calculated by solving equation 11 for it:

$$C_T = C_{12} Q_o \sqrt{g_1 g_2}$$

Equations 12a and 13a are used in place of equations 12 and 13 and transducer gain is calculated using equations 21 and 22 instead of power gain from equations 18 and 19.

The results of this initial calculation shown in Table IV are unsatisfactory. The value of R_1 is negative, indicating that an exact input match isn't possible. (An exact input match, $m_s = 1$, is impossible with finite Q_u except when N = 1.) Also the value of C_T is smaller than is desirable and n_1 is too large. Satisfactory values are obtained when $m_s = 1.5$ and C_{12} = 2.2 pF are used.

The single-tuned output circuit design is based on a prototype element obtained from the equations in Figure 2 for a Chebyshev filter with N = 1, Am = 1.0 dB and a ripple bandwidth of 1 rad/sec, thus the specified 1.5 MHz tuned circuit bandwidth is the 1.0 dB bandwidth.

The value $M_L = 1$ (output matched to 50 ohms) is chosen and equations 14a and 15a are used in the calculations instead of 14 and 15. Also equation 16c is used instead of 16a & b; and 17d in place of 17a. Initially, C_T is calculated from equations 25 and 26 for $n_1 = 1$. The resulting capacitance value, C_T = 246 pF, is much too large. The calculation is repeated using $C_T = 246/9 = 27.3$ pf so that n_1 = 3. Component values are satisfactory after this change.

Calculated transducer gain of the two-stage amplifier is 35.6 − 6.5 + 22.2 = 51.3 dB.

Figure 5 is a schematic diagram of the amplifier. The 99K calculated value for R_1 in the input circuit is so large that this resistor can be omitted. In the output single-tuned circuit the value of the variable capacitor, which is assumed to be 6 pf, is subtracted from C_1 and the two fixed capacitor values are then calculated to be $20.3n_2$ = 181 pF

```
100 REM: PROGRAM TUNEDAMP
110 CLEAR 500
120 DEFINT I-J
130 DIM B(4,2), C(8), G(4,2), M(8), P(8), T$(11),
  Y(8, 10), Y$(8)
140 FOR J1 = 1 TO 8: READ Y$(J1):
NEXT J1
150 DATA "11", "12", "21", "22", "ia",
  "oa", "ib", "ob"
160 PI = 3.141593: DP = 10/LOG(10)
170 REM: READ TRANSISTOR DATA IN
  DATA STATEMENTS BEGINNING AT LI
NE 3000
180 FOR J5 = 1 TO 11
190 READ T$(J5)
200 IF T$(J5) = "****" THEN 240
210 FOR J1 = 1 TO 8
220 READ Y(J1, J5)
230 NEXT J1, J5
240 J5 = J5 - 1
250 F1$ = "TUNED AMPLIFIERS"
260 F2$ = CHR$(13)
270 F3$ = "TRANSISTORS"
280 F5$ = "C## = ##.###↑↑↑↑"
290 F6$ = "PROTOTYPE ELEMENT
  VALUE S"
300 F7$ = "Y%% = + #.##↑↑↑↑ + #.##↑↑↑↑
j"
310 REM: ENTER Am = 0 FOR BUTTER-
  WORTH FILTERS, Am = -1 TO ENTER
  PROTOTYPE VALUES FROM THE KEY-
  BOARD, e.g. FOR BESSEL FILTERS
320 CLS: PRINT F1$: INPUT "N, Am (dB)";
  N%, A1
330 IF N%<1 THEN 3990
340 I1 = 0: I3 = 0: A6 = 0: A7 = 0: R3 = 1
350 IF A1 > = 0 THEN 420
360 PRINT F6$
370 FOR J1 = 1 TO N%
380 PRINT J1;: INPUT P(J1)
390 NEXT J1
400 GOTO 760
410 REM: CALCULATE PROTOTYPE ELE-
  MENT VALUES FOR BUTTERWORTH
  AND CHEBYSHEV FILTERS
420 FOR J1 = 1 TO N%
430 P(J1) = 2*SIN((2*J1 - 1)*PI/(2*N%))
440 NEXT J1
450 IF A1 = 0 THEN 760
460 INPUT "SPECIFY: 1. RIPPLE BW,
  2. 3dB BW";I3
470 A3 = 2
480 A2 = EXP(A1/DP)
490 B2 = LOG(1 + 2/(SQR(A2) - 1))/2
500 X1 = EXP(B2/N%)
510 G1 = (X1 - 1/X1)/2
520 G2 = G1*G1
530 U1 = P(1)
540 P(1) = U1/G1
550 IF N% < 2 THEN 680
560 FOR J1 = 2 TO N%
570 U2 = P(J1)
580 T1 = SIN((J1 - 1)*PI/N%)
590 B1 = G2 + T1*T1
600 P(J1) = U1*U2/(B1*P(J1 - 1))
610 U1 = U2
620 NEXT J1
630 IF N% > 2*CINT(N%/2) THEN 680
640 T1 = 1 - 2/(EXP(B2) + 1)
650 R3 = T1*T1
660 A3 = A3*A2
670 A6 = A1
680 IF I3 < 2 THEN 760
690 T1 = SQR((A3 - 1)/(A2 - 1))
700 T2 = LOG(T1 + SQR(T1*T1 - 1))
710 X1 = EXP(T2/N%)
720 W1 = (X1 + 1/X1)/2
730 FOR J1 = 1 TO N%
740 P(J1) = W1*P(J1)
750 NEXT J1
760 IF N% < 3 THEN 800
770 FOR J1 = 2 TO N% - 1
780 A7 = A7 + P(J1)
790 NEXT J1
800 PRINT F6$; ":"
810 J2 = 0: J6 = 15
820 FOR J1 = 1 TO N%
830 PRINT TAB(J2) P(J1);
840 J2 = J2 + J6
850 IF J2 > 3*J6 THEN J2 = 0: PRINT
860 NEXT J1
870 IF J2 > 0 THEN PRINT
880 PRINT "r = "; R3
890 PRINT: INPUT "PRINT ELEMENT
  VALUES"; I$
900 IF LEFT$(I$,1) = "Y" THEN GOSUB
  2980
910 CLS: PRINT F1$;"  N = ";N%;"  Am
  = ";A1
920 PRINT F3$; ":"
930 J2 = 0: J6 = 30
940 FOR J1 = 1 TO J5
950 PRINT TAB(J2)STR$(J1);". "; T$(J1);
960 J2 = J2 + J6
970 IF J2 > J6 THEN J2 = 0: PRINT
980 NEXT J1
990 IF J2 > 0 THEN PRINT
1000 PRINT "COUPLING NETWORK TYP
  ES: 1. INTERSTAGE 2. INPUT 3. OUTPUT"
1010 INPUT "TYPE"; I1
1020 ON I1 + 2 GOTO 3980, 320, 103, 0, 1050,
  1050
1030 PRINT F3$;: INPUT I5, I6
1040 GOTO 1070
1050 INPUT "TRANSISTOR"; I5
1060 I6 = I5
1070 For J1 = 1 TO 4
1080 IF I1 = 2 THEN 1120
1090 G(J1, 1) = Y(2*J1 - 1,I5)
1100 B(J1, 1) = Y(2*J1,I5)
1110 IF I1 = 3 THEN 1140
1120 G(J1, 2) = Y(2*J1 - 1, I6)
1130 B(J1, 2) = Y(2*J1,I6)
1140 NEXT J1
1150 ON I1 GOTO 1160, 1200, 1260
1160 INPUT "Mia, Moa, Mib, Mob"; M3, M1,
  M2, M4
1170 GOSUB 2070
1180 PRINT "Ka = "; K1; "  Sa = "; S1:
  TAB(31)"Kb = ";K2;" Sb = ";S2
1190 GOTO 1300
1200 INPUT "GS, BS, MS, Mib, Mob";
  G(4, 1), B(4, 1), M1, M2, M4
1210 G1 = G(4, 1): B1 = B(4, 1)
1220 G3 = G1: Y2 = 4
1230 GOSUB 2220
1240 PRINT "K = "; K2, "S = "; S2
1250 GOTO 1300
1260 INPUT "MIA, MOA, ML, GL, BL";
  M3, M1, M2, G(1, 2), B(1, 2)
1270 G2 = G(1,2): B2 = B(1, 2)
1280 GOSUB 2070
1290 PRINT "K = "; K1, "S = "; S1
1300 INPUT "ARE K AND S OK"; I$
1310 IF LEFT$(I$, 1) = "N" THEN 1150
1320 IF I1 = 2 THEN 1340
1330 Y2 = (G(3,1)*G(3,1) + B(3, 1)*B(3,1))/
  (G(4,1)*G(4,1))
1340 PRINT: INPUT "F0(MHz), BW(MHz),
  Qu"; F0, B0, Q3
1350 T1 = B0/2
1360 F1 = SQR(F0*F0 - T1*T1)
1370 W0 = 2E6*PI*F1
1380 Q0 = F1/B0
1390 Q1 = Q0*P(1)
1400 Q2 = Q0*P(N%)/R3
1410 A8 = A6 + DP*A7*Q0/Q3
1420 INPUT "OPTIONS: 1. SELECT
C12 2. SELECT CT 3. N1 = 1 4. N 2 = 2";I2
1430 ON I2 GOTO 1440, 1480, 1500, 1530
1440 IF N% = 1 THEN 1480
1450 INPUT "C12"; X1
1460 C0 = X1*Q0*SQR(P(1)*P(2))
1470 GOTO 1550
1480 INPUT "CT"; C0
1490 GOTO 1550
1500 X1 = (M1*G(4, 1) + G1)/2
1510 C0 = Q1*X1/W0
1520 GOTO 1550
1530 X1 = (M2*G(1, 2) + G2)/2
1540 C0 = Q2*X1/W0
1550 X1 = W0*C0
1560 L1 = 1/(W0*X1)
1570 G5 = X1/Q1
1580 G6 = X1/Q2
1590 N3 = (M1*G(4, 1) + G1)/(2*G5)
1600 N1 = SQR(N3)
1610 N4 = (M2*G(1,2) + G2)/(2*G6)
1620 N2 = SQR(N4)
1630 G7 = G5*(1 - Q1/Q3) - G1/N3
1640 G8 = G6*(1 - Q2/Q3) - G2/N4
1650 IF N% = 1 THEN G7 = G7 + G8 +
  X1/Q3
1660 M(0) = B1/(W0*N3)
1670 M(N%) = B2/(W0*N4)
1680 R1 = 1/G7
1690 IF N% = 1 THEN 1750
1700 R2 = 1/G8
1710 T1 = C0/Q0
1720 FOR J1 = 1 TO N% - 1
1730 M(J1) = T1/SQR(P(J1)*P(J1 + 1))
1740 NEXT J1
1750 FOR J1 = 1 TO N%
1760 C(J1) = C0 - M(J1 - 1) - M(J1)
1770 NEXT J1
1780 X2 = 1 + M1
1790 X1 = Y2/(X2*X2)
1800 X2 = N3*G5*G2/(N4*G6*G3)
1810 A5 = DP*LOG(X1*X2) - A8
1820 PRINT "Fc = "; F1; "MHz"
1830 IF I1 = 2 THEN PRINT "AP(Trans-
  ducer) = "; A5; "dB": ELSE PRINT "AP
  = "; A5; "dB"
1840 IF N% = 1 THEN PRINT "QL = "; Q
  1/2: ELSE PRINT "Q1 = "; Q1;"  Q2 = ";
  Q2
1850 PRINT "Qu = "; Q3
1860 PRINT "R1 = "; R1;
1870 IF N% = 1 THEN PRINT: ELSE
  PRINT"  R2 = "; R2
1880 PRINT "N1 = ";N1;"  N2 = "; N2
1890 PRINT "L = "; L1;"  CT = "; C0
1900 FOR J1 = 1 TO N%
1910 PRINT USING F5$; J1; C(J1);
1920 IF J1 = 3*CINT(J1/3) THEN PRINT:
  ELSE PRINT CHR$(195);
1930 NEXT J1: PRINT
1940 IF N% = 1 THEN 2000
1950 FOR J1 = 1 TO N% - 1
1960 J2 = 11*J1 + 1
1970 PRINT USING F5$; J2; M(J1);
1980 IF J1 = 3*CINT(J1/3) THEN PRINT:
  ELSE PRINT CHR$(195);
1990 NEXT J1
2000 PRINT: INPUT "PRINT RESULTS"; I$
2010 IF LEFT$(I$,1) = "Y" THEN GOSUB
  2380
2020 PRINT: INPUT "NEW C"; I$
2030 IF LEFT$(I$,1) = "Y" THEN 1420
2040 INPUT "NEW M'S"; I$
2050 IF LEFT$(I$,1) = "Y" THEN 1150:
  ELSE 910
2060 REM: SUBROUTINES FOR INPUT &
  OUTPUT IMPEDANCES AND STABILITY
  FACTORS
2070 T1 = G(2,1)*G(3,1) - B(2, 1)*B(3, 1)
2080 T2 = G(2, 1)*B(3, 1) + G(3, 1)*B(2, 1)
2090 T3 = SQR(T1*T1 + T2*T2)
2100 X1 = (1 + M3)*G(1,1)
2110 G1 = G(4, 1) - T1/X1
2120 B1 = B(4, 1) - T2/X1
2130 X2 = (1 + M1)*G(4, 1)
2140 G3 = G(1, 1) - T1/X2
2150 B3 = B(1, 1) - T2/X2
2160 X2 = X2*X1
2170 K1 = 2*X2/(T1 + T3)
2180 X1 = X2 - T1
2190 X2 = X1*X1 + T2*T2
2200 S1 = T3/SQR(X2)
2210 IF I1 > 1 THEN 2360
2220 T4 = G(2, 2)*G(3, 2) - B(2,2)*B(3, 2)
2230 T5 = G(2, 2)*B(3, 2) + G(3, 2)*B(2, 2)
2240 T6 = SQR(T4*T4 + T5*T5)
2250 X1 = (1 + M4)*G(4, 2)
2260 G2 = G(1, 2) - T4/X1
2270 B2 = B(1, 2) - T5/X1
2280 X2 = (1 + M2)*G(1, 2)
2290 G4 = G(4, 2) - T4/X2
2300 B4 = B(4, 2) - T5/X2
2310 X2 = X2*X1
2320 K2 = 2*X2/(T4 + T6)
2330 X1 = X2 - T4
2340 X2 = X1*X1 + T5*T5
2350 S2 = T6/SQR(X2)
2360 RETURN
2370 REM: LINE PRINTER OUTPUT SUB
  ROUTINES
2380 LPRINT
2390 IF I1 = 1 THEN 2430
2400 LPRINT "TRANSISTOR:"; F2$; T$(I5)
2410 GOSUB 2930
2420 ON I1 GOTO 3980, 2570, 2630
2430 LPRINT F3$ + ":"
2440 LPRINT "1. "; T$(I5)
2450 GOSUB 2930
2460 LPRINT "2. "; T$(I6)
2470 GOSUB 2920
2480 LPRINT "Mia = "; M3;"  Moa = "; M1
2490 LPRINT "Mib = ";M2;"  Mob = "; M4
2500 LPRINT "Ka = ";K1;"  Sa = "; S1
2510 LPRINT "Kb = ";K2;"  Sb = "; S2
2520 LPRINT USING F7$; Y$(5), G3, B3
2530 LPRINT USING F7$; Y$(6), G1, B1
2540 LPRINT USING F7$; Y$(7), G2, B2
2550 LPRINT USING F7$; (Y$(8), G4, B4
2560 GOTO 2680
2570 LPRINT "GS = "; G(4, 1); "  BS = ";
  B(4, 1); F2$; "MS = "; M1
2580 LPRINT "Mib = ";M2;"  Mob = "; M4
2590 LPRINT "K = ";K2;"  S = "; S2
2600 LPRINT USING F7$; Y$(7), G2, B2
2610 LPRINT USING F7$1; Y$(8), G4, B4
2620 GOTO 2680
2630 LPRINT "GL = "; G(1, 2); "  BL = ";
  B(1,2); F2$; "  ML = "; M2
2640 LPRINT "Mia = "; M3;"  Moa = "; M1
2650 LPRINT "K = "; K1; "  S = "; S1
2660 LPRINT USING F7$; Y$(5), G3, B3
2670 LPRINT USING F7$; Y$(6), G1, B1
2680 LPRINT F2$; "N = "; N%
2690 IF A1 > = 0 THEN LPRINT "Am = ";
  A1; "dB"
2700 IF I3 = 1 THEN LPRINT "RIPPLE BW"
2710 IF I3 = 2 THEN LPRINT "3 dB BW"
2720 I7 = 1
2730 IF I1 > 0 THEN 2780
2740 LPRINT F6$; ":"
2750 FOR J1 = 1 TO N%: LPRINT J1; P(J1):
  NEXT J1
2760 LPRINT "r = "; R3; F2$
2770 RETURN
2780 LPRINT "F0 = "; F0; "MHz"; F2$; "Fc
  = "; F1; "MHz"; F2$; "BW = "; B0; "MHz"
  "MHz"
2790 IF I1 = 2 THEN LPRINT "AP(Trans-
  ducer) = "; A5; "dB":ELSE LPRINT
  "AP = "; A5; "dB"
2800 IF N% = 1 THEN LPRINT "QL = ";
  Q1/2; "  ";:ELSE PRINT "Q1 = "; Q1;
  " Q2 = "; Q2
2810 LPRINT "Qu = "; Q3; F2$ "R1 = "; R1
2820 IF N% > 1 THEN LPRINT "R2 =
  "; R2
2830 LPRINT "N1 = "; N1;"  N2 = "; N2
2840 LPRINT "L = "; L1; F2$; "CT = "; C0
2850 FOR J1 = 1 TO N%
2860 J2 = 11*J1 + 1
2870 LPRINT USING F5$; J1, C(J1)
2880 IF J1 = N% THEN 2900
2890 LPRINT TAB(15) USING F5$; J2, M(J1)
2900 NEXT J1
2910 LPRINT: RETURN
2920 J2 = I6: GOTO 2940
2930 J2 = I5
2940 FOR J1 = 1 TO 4
2950 LPRINT USING F7$; Y$(J1), Y (2*J1
  - 1, J2), Y(2*J1, J2)
2960 NEXT J1
2970 LPRINT: RETURN
3000 DATA "MC1590 21.4 MHz"
3010 DATA 3.66E-4, 8.13E-4, -1.38E-8,
  -3.58E-8
3020 DATA 1.43E-1, -8.77E-2, 5.89E-5,
  3.99E-4
3030 DATA "2N3251A 5 mA 21.4 MHz"
3040 DATA 5.34E-3, 4.52E-3, -7.33E-7,
  -2.86E-4
3050 DATA 9.04E-2, -7.05E-4, 4.13E-4,
  6.49E-4
3980 DATA "****"
3990 IF I7 = 1 THEN LPRINT STRING$
  (3, 13)
4000 END
```

Table VI. TRS-80 Program.

and $20.3n_2/(n_2 - 1) = 22.9$ pF. Since this circuit drives a 50 ohm resistive load, the loading resistor R_1 can be placed in parallel with the inductor instead of directly in parallel with the output.

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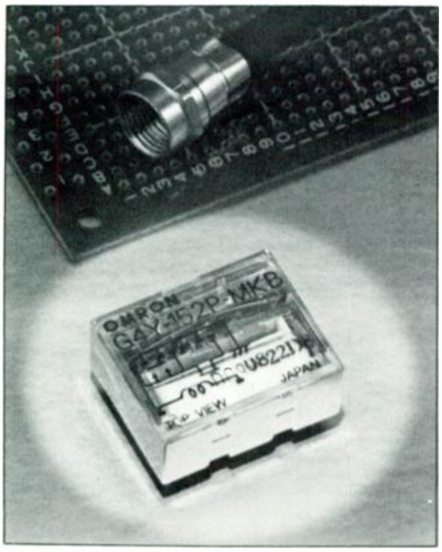
new products

High Performance VHF/UHF Relay

A compact special-purpose PCB relay for switching VHF and UHF signals in a broad range of consumer and industrial electronic products has been introduced by the Control Components Division of Omron Electronics, Inc. The new double-break SPDT relay (Model G4Y) offers a more miniaturized design (about the size of a quarter) and operating characteristics at RF signal levels that match or ex-



ceed those of larger coaxial or RF-switching relays presently being marketed. Performance features for optimal high-frequency switching and signal transmission include high isolation (rated 65 dB min. between switched circuits), low insertion loss (less than 0.5 dB at 900 MHz), and a relatively well-matched VSWR of 1.5 at 900 MHz for maximum power transfer. The G4Y offers unique design in a miniaturized PCB-mountable package, incorporating a metallic RF-protective shield plate and strong metal-plated poly-carbonate resin-molded base while integrating an internal ground contact pin to minimize stray capacitance and signal loss due to capacitative shunting across the relay. Gold-plated bifurcated contacts provide a wide switching capacity, ranging from micro-current/voltage to 1 W loads. The G4Y is light-weight (less than an ounce), compact in size (26 mm W × 21 mm D × 10.5 mm H), has low power consumption (450 mW), and flux-tight construction for auto-flow or dip soldering. A second model series incorporates a special internal connection, reversing the standard model's terminal arrangement for NO and NC contacts. Models include highly-sensitive DIP relays compatible with TTL circuitry, flat-pack power relays and versatile sugar-cube-size families offering a broad selection of contact configurations, coil ratings and switching capacities. List price for the standard G4Y is $4.99 per unit. **Contact Joanne Sullivan, Control Components Division, Omron Electronics, Inc., 650 Woodfield, Schaumburg, IL, 60195, (312) 843-7900 or circle INFO/CARD #140.**

SOT-23 Devices

M/A-COM Silicon Products, Inc., now offers its complete line of semiconductor devices in the SOT-23 package. These reliable, low-cost, compact devices are DC and RF characterized to applicable product specifications. Available devices include MNS (metal-nitride-silicon) capacitors, pin diodes, tuning varactors, Schottky diodes and silicon bipolar transistors. All of these products are readily available in high volume quantities. Typical pricing is as low as 17¢ for 10K quantities for the 4M1010-287 capacitor. **M/A-COM Silicon Products, Inc., Burlington, MA 01803, (617) 272-3000 or please circle INFO/CARD #106.**

Portable Frequency Counter/Wattmeter

Coaxial Dynamics, Inc. is featuring a combination frequency counter/wattmeter, which is both lightweight and totally portable. The frequency counter is a fully-integrated multi-functional counter with maximum versatility provided by placement of all critical controls on the front panel.

BNC inputs are included on both front and back panels, facilitating service bench or in-field use. Wattmeter measurements include forward and reflected power at the touch of a rocker-switch. An additional ×10 switch allows forward power reading increases by a factor of 10 without additional changes or calibration. Six interchangeable pickup modules provide all inputs necessary for each scale designation. Modules are available that may be left in transmission lines of chosen pieces of equipment while the FCW-10 and other modules are taken to separate pieces of equipment. Modules are capable of remote use in transmission lines up to 6 feet away from the FCW-10. Total portability is made possible by use of an optional battery pack which automatically recharges when the unit is connected to an AC line. **Contact Coaxial Dynamics, Inc., 13110 Enterprise Avenue, Cleveland, Ohio, 44135, (216) 267-2233 or circle INFO/CARD #105.**

LCZ Meter

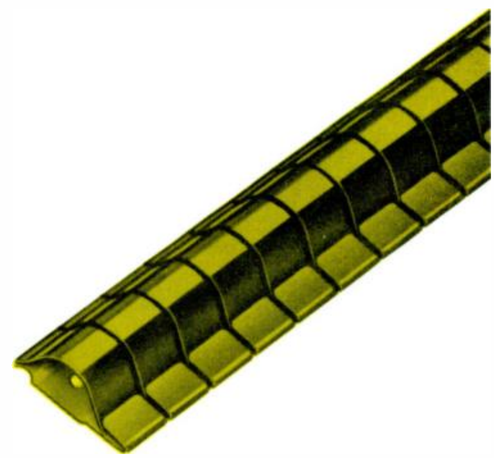
Two new LCZ (capacitance, inductance and impedance) meters, introduced by Hewlett-Packard, are designed for: incoming/outgoing inspection of capacitors, diodes, coils, audio heads and transformers; analysis of electronic materials, sensors, ceramics and ferrites; and evaluation of semiconductor devices and C-V/G-V characteristics of MOS and bipolar devices. The new HP Models 4276A and 4277A, with up to 30-millisecond test speed, have a combined test-frequency range of 100 Hz to 1 MHz. Individual ranges are 100 Hz to 20 kHz (1 Hz maximum resolution) for the HP 4276A and 10 kHz to 1 MHz (100 Hz maximum resolution) for the HP 4277A.

Price for HP 4276A (100 Hz to 20 kHz) is $3,850. The HP 4277A (10 kHz to 1 MHz) is priced at $5,500. **Call your local Hewlett-Packard sales office or INFO/CARD #139.**

Coaxial/Switchable Attenuators

Teledyne Microwave announces its SA-01S50 series of Coaxial Switch/Attenuators. Up to 30 dB attenuation is available with frequencies ranging from DC to 18 GHz. RF performance

This narrow strip...



at 18 GHz is 1.5:1 VSWR with insertion loss of .5 dB in the through path state. The Switchable/Attenuator is available with either latching or failsafe actuators. **Teledyne Microwave, 1290 Terra Bella Avenue, Mountain View, CA 94043, (415) 968-2211, or please circle INFO/CARD #138.**

Integrated Mixer Preamp

An integrated mixer preamp measuring only 1 × 1 × 3/8 inches (3/8 cubic inch) and combining several state-of the-art technologies is available from RHG Electronics Laboratory, Inc. Designated the DML2-18/XXX series, the new mixer preamps are the world's smallest multioctave microwave mixer preamps currently being produced. The units, which meet the requirements of high density military packaging, incorporate built-in limiters to protect the mixers from signal overloads. The device is hermetically sealed, and can be used either with connectors or without ("dropped-in" to the user's circuit). After testing with connectors in place, the connectors are easily removed for the "drop-in" or microstrip mounting.

The high density MIC and IC technology has reduced the new DML series to one-seventh (1/7) the volume of present catalog models. The size reduction has not compromised the unit's electrical performance. In a microstrip line with a high even-mode impedance the stray capacitance and resonance effects from adjacent circuitry and housing are virtually eliminated with the DML series. Three DML models offer 2-18 GHz RF coverage and IF frequency/bandwidth combinations of 30/10, 60/20, and 160/50 MHz. **RHG Electronics Laboratory, Inc., 161 East Industry Ct., Deer Park, N.Y. 11729. (516) 242-1100 or please circle INFO/CARD #137.**

2 GHz Modulation Meter

Marconi Instruments introduces a new full facility Modulation Meter,

Model 2305, which combines the functions of a conventional Modulation Meter, RF Power Meter, Frequency Counter, and Audio Analyzer into a single convenient package. The 2305 is capable of measuring FM deviations up to 500 kHz, phase deviations to 500 radians and AM depths up to 99.9 percent with modulating frequencies up to 300 kHz (50 kHz for AM). Ranges are selected automatically by the instrument to give the best possible resolution. Among the various available detector responses is an 'average-peak' for all routine modulation measurements, avoiding the necessity of making both positive and negative peak measurements to derive average peak. This is particularly important in automatic test systems where test time must be minimized. The model 2305 Modulation Meter is priced at $7,500. **Marconi Instruments, 100 Stonehurst Court, Northvale, N.J. 07647, (201) 767-7250, or please circle INFO/CARD #134.**

Automatic Power Meter Calibrator

Califactor System II by Weinschel Engineering provides automatic calibration of RF power meter/mount combinations operation in the 0.01 to 18 GHz range. The system is IEEE-488 bus controlled and reduces a laborious time consuming task to a quick and accurate procedure. Calibration speed of the system is typically 1.5 seconds per measurement frequency depend-

ing on power meter setting time. System II can be used to calibrate thermistor and thermocouple as well as diode-type RF power meter mounts. Accurate measurement of the output level of signal sources can also be accomplished. **Weinschel Engineering, One Weinschel Lane, Gaithersburg, MD 20760, (301) 948-3434 or please circle INFO/CARD #136.**

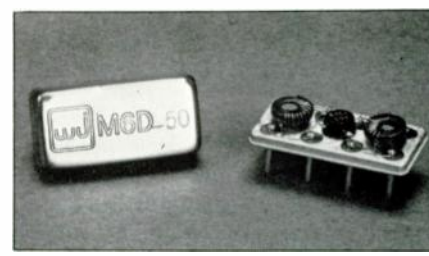
Signal Generator (2-18.0 GHz)

The HP 8672A synthesized signal generator (2-18.0 GHz) now has a frequency-extender unit available to provide high-performance synthesized signals all the way down to 10 MHz. This 10-2000 MHz extension band is important for certain surveillance receiver testing and as a broadband source for automatic test systems. In addition, the extender unit includes a pulse modulator for full-range pulsing with greater than 80 dB ON/OFF ratio and less than 10 ns RISE/FALL times. HP 8672A-E24 generator, 10 MHz to 18.0 GHz $55,000 (includes HP 8672A).

Delivery: 12-16 weeks ARO. **Contact local Hewlett-Packard sales office or INFO/CARD #135.**

Mixers Feature 0 dBm LO Drive

Watkins-Johnson Company has introduced two mixers featuring conversion loss performance as low as 5.5

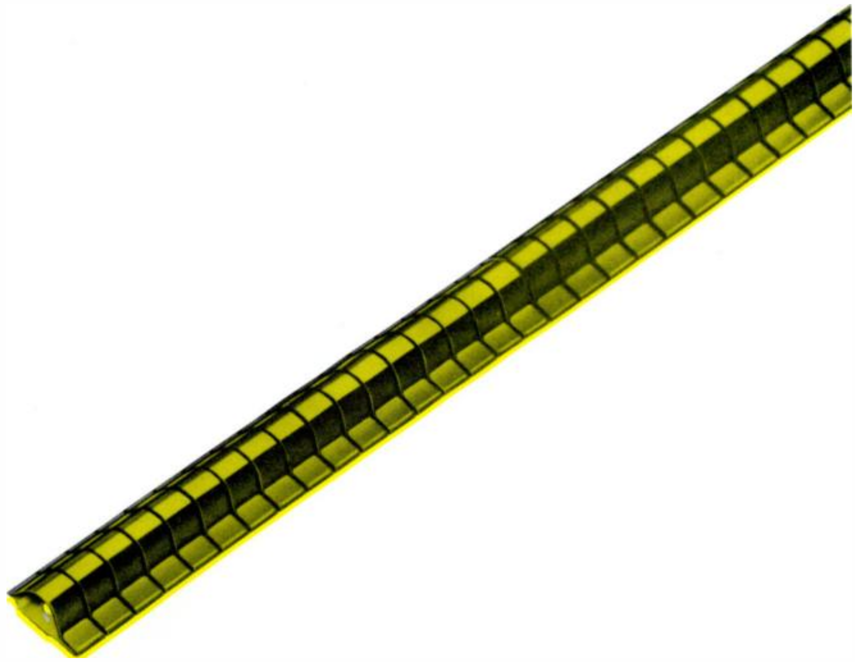


dB with 0 dBm LO drive. The mixers cover the frequency range of 0.05 to 500 MHz. These low-level devices have a typical LO-to-RF isolation of 55 dB and an LO-to-IF isolation of 53 dB below 50 MHz. Above 50 MHz, the LO-to-RF isolation is 43 dB while the LO-to-IF is 36 dB. Designated the M6D-50 and M6E-50, these double-balanced mixers are hermetically sealed, hi-rel tested, RFI shielded and are available in PC relay packages. **Watkins-Johnson Company, Components Applications Engineering, 3333 Hillview Ave., Palo Alto, CA 94304. (415) 494-4141, ext. 2637 or INFO/CARD #132.**

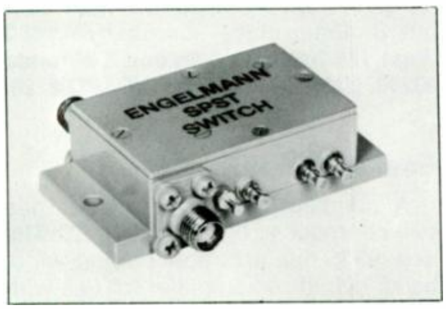
100-18,000 MHz SPST Diode Switch

Engelmann Microwave has announced the availability of a new SPST Diode Switch which operates with a full frequency range of 100 MHz to 18,000 MHz in a single unit. New Model SW 2018A features very low insertion loss with typical isolations rated at 60 dB over most of the frequency range. The guaranteed maximum insertion loss in "L" and "S"

...or this narrower strip...



Band is 0.5 db. The worst case insertion loss up to 18 GHz is 2.5 db maximum. The Model SW 2018A is



supplied with the TTL compatible driver and offers switching speeds of 10 ns maximum, 10-90 percent point. The VSWR in the on position is 1.8. The Model SW 2018A is priced at $250.00 each with deliveries of thirty days maximum. **Engelmann Microwave Co., 662 Myrtle Avenue, Boonton, New Jersey, 07005, (201) 334-5700, or INFO/CARD #131.**

Miniature Flatpack RF Amplifiers

The Alpha Microelectronics Modular RF Amplifiers, AF-401, AF-402, and AF-403, are low cost, single stage, thick-film hybrid modules designed for insertion into microstrip applications requiring high dynamic range and a flat response over a wide frequency range. The modular RF Amplifiers are cascadable to satisfy any desired gain level without sacrificing bandwidth in a 50 ohm system. Featuring the following typical parameters: Freq: 5-400 MHz, Gain: 14.5 dB, Noise figure: 4-7.5 dB, VSWR: 1.5, the AF series are designed to satisfy the most stringent military and aerospace environments in receiver front ends and test equipment. **Alpha Microelectronics Division, 3015 Advance Lane, Colmar PA 18915, (215) 822-1311, or please circle INFO/CARD #127.**

Miniature Leadless Chip Carrier Oscillator

The smallest clock oscillator yet! Construction includes a thick film hybrid oscillator circuit with a precision McCoy crystal matched for optimum performance. Features are size (.480 × .480 × .085") and stabilities (±50 PPM −20 to +70°C, ±75 PPM −55 to +105°C, and ±100 PPM −55 to +125°C), and the oscillators may be screened per MIL-0-55310 for MIL applications. Other specifications include Input +5 VDC ±.25 V, Current 40 ma maximum 12 MHz to 35 MHz and 60 ma maximum 35 MHz to 60 MHz, and TTL Output. Model MC070 is avail-

able to custom requirements. **McCoy Electronics Company, 100 Watt Street, P.O. Box B. Mt. Holly Springs, PA 17065, (717) 486-3411, or please circle INFO/CARD #129.**

Ultra-Miniature Couplers

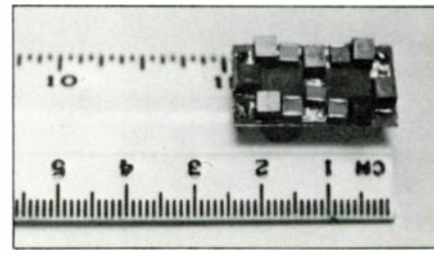
Midisco has introduced the MDC 6200 series of miniature directional couplers. These precision couplers have been designed to achieve high directivity and low VSWR in a miniature stripline configuration. Etched teflon fiberglass stripline circuits are enclosed within a machined housing that eliminates RF moding and provides high RF shielding. Connectors are type SMA stainless steel female complying with Mil-C-39012. The center pins are resistance soldered to the stripline circuit to provide a rugged, reliable contact. **Midisco, 61 Mall Drive, Commack, NY 11725, (516) 543-4774, or INFO/CARD #126.**

Six Watt Power FET

A new six watt power field effect transistor (FET) with what the manufacturer describes as superior guaranteed performance at both band edges (5.9 - 6.4 GHz) has been developed by the Microwave Semiconductor Corp., a division of Siemens Components, Inc. The MPF 5964-6 delivers up to six watts of CW power with 6dB typical gain over the 5.9—6.4 GHz band. The device is internally matched to a 50 ohm system with input and output VSWR less than 2:1 typical. Performance of +37 dBm output power with 5 dB minimum gain is guaranteed in 50 ohms at test points of 5.9 and 6.4 GHz. Delivery times for the devices are being quoted as 8 weeks. Typical pricing is $315 each in 100 piece quantities. **Microwave Semiconductor Corporation, 100 School House St., Somerset, N.J. 08873 or INFO/CARD #130.**

Sub Miniature Chip Filters

Vanguard Electronics has extended its state-of-the-art chip inductor technology to develop new subminiature RF bandstop, bandpass, lowpass and highpass filters with exceptional electrical characteristics. Circuit parasitics have been minimized to such an extent that upper passband frequency to stop frequency ratios may exceed 50 to one on many designs. Besides conventional RF applications, these units can also be used for filtering in data processing systems. Custom designs can be made



to DIP package dimensions. The unit pictured is a balanced-line five-section Chebyschev design with a bandstop center frequency of 1.0 MHz, and a 3 dB bandwidth of 0.85 MHz. Typical rejection at 1.0 MHz is in excess of 50 dB. Passband insertion loss remains less than 2.0 dB to 100 MHz. Size is only 0.5" wide × 1.0" long × .18" high, with total weight of approximately one gram (.04 ounces). Average Price: $42.00 each (1000 lot qty.), Delivery: 8 weeks ARO. **Vanguard Electronics Company, 1480 West 178th Street, Gardena, California 90248, (213) 323-4100, INFO/CARD #128.**

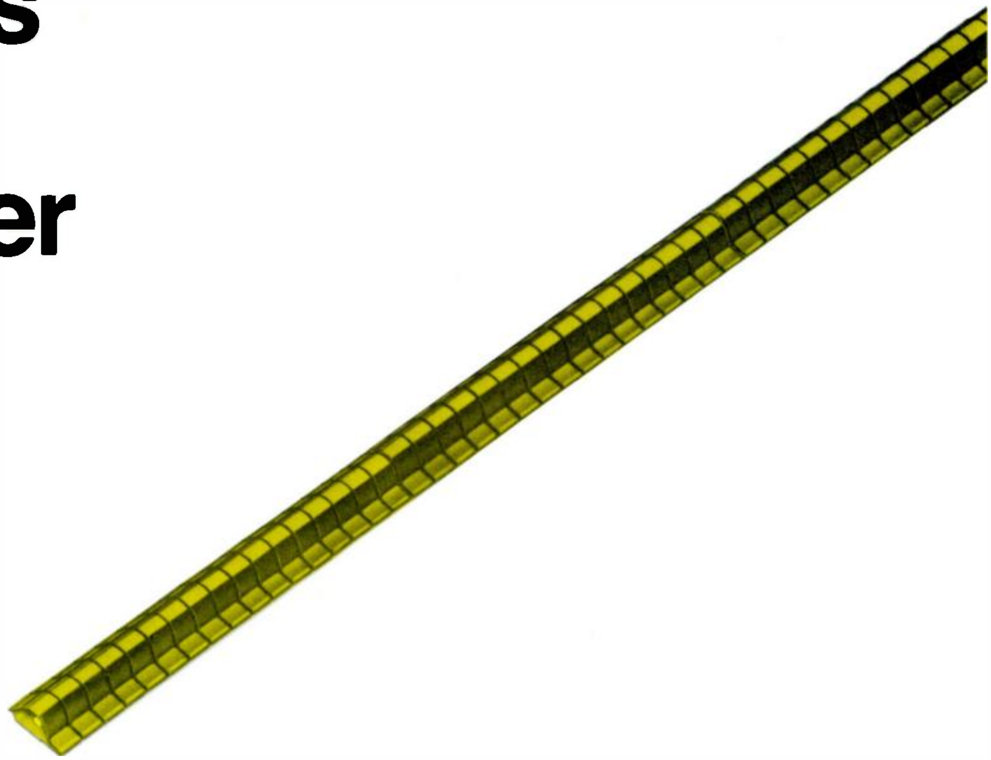
Fast Preamplifier

A compact, fast preamplifier has been introduced by EG&G ORTEC. The new 9305 has a direct-coupled wideband, hybridized amplifier for use with photomultipliers, electron multipliers and other detectors for ion and photon counting or fast timing applications. The detector can be mounted directly to the 9305 to minimize system noise. The 9305 features a fast rise time of <3 ns, a variable voltage gain of 5-10 and can drive ±5 V into 50 Ω load. Overload input protection is included. **EG&G ORTEC, 100 Midland Road, Oak Ridge, TN 37830. (800) 251-9750, or INFO/CARD #125.**

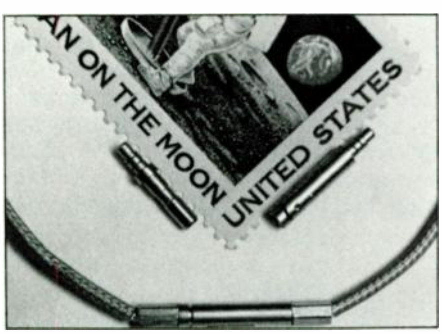
Ultra-Miniature Coaxial Connector

Sabritec has developed a new ultra-miniature coaxial connector designed specifically for coaxial cables .100"

...or this even narrower strip...



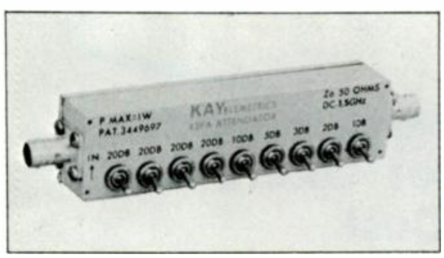
or smaller. Ideal for use in medical and high density applications, the fully mated connectors have a maximum outside diameter of .112" and a maximum length of .950". Mating of the connectors is accomplished by use of a unique quick-connect feature. All connector configurations feature gold plating and the choice of crimp or



solder termination. **Sabritec, 16631 Noyes Avenue, Irvine, CA, (714) 549-7292 or INFO/CARD #124.**

Miniature Attenuator

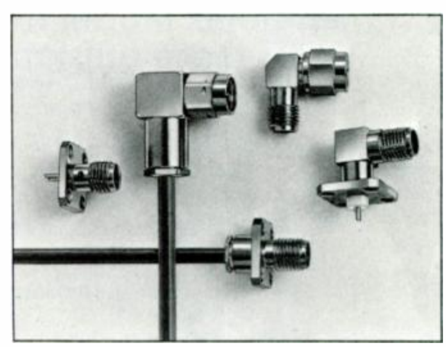
Kay Elemetrics Corp. has redesigned the 439 Miniature Attenuator to provide a higher frequency range (DC-1500 MHz). Attenuation range is 0-101 dB in 1 dB steps. VSWR is 1.2 in the DC-500 MHz range and 1.4 in



the 500-1500 MHz range. Insertion loss is .2dB in the DC-500 MHz range, .5 dB in the 500-1000 MHz range, and .8 dB in the 1000-1500 MHz range. The 439 Miniature Attenuator is designed to meet small (1-¼" x 6") panel space requirements. Available with BNC connectors (optional connectors: SMA or TNC). Price $209.00, Delivery 3-4 weeks. **Kay Elemetrics Corp., 12 Maple Ave., Pine Brook, NJ 07058. (201) 227-2000, or INFO/CARD #123.**

Unitized Body SMA Connector

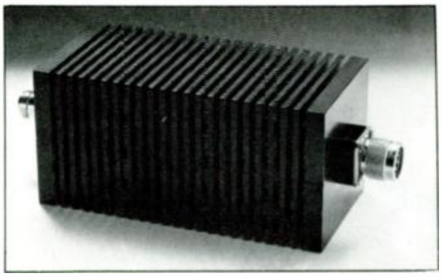
Automatic Connector, Inc., has now developed its unitized body SMA connector series to satisfy requirements in electronic countermeasures as well as other military applications. The unitized body SMA series is a high performance coaxial connector suited for applications where mechanical integrity is paramount. It features a machined stainless steel single piece body, eliminating the potential of breaking or cracking of braze joints at the flange and right angle junctions. This manufacturing technique is a cost-effective process when compared to other procedures required to assure a mechanically superior joint especially when passivated components are used. The right angle



connectors, including receptacles and adapters, exhibit excellent VSWR and insertion loss through 18 GHz due to a unique internal compensation step as well as the use of a single piece machined contact. All other parameters meet or exceed the applicable Mil-C-39012 specification. **Automatic Connectors Inc., 400 Moreland Rd., Commack, N.Y. 11725, (516) 543-5000 or INFO/CARD #122.**

100 Watt Fixed Attenuator

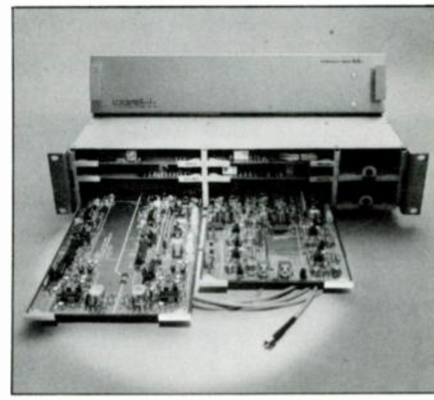
50 Ohm * DC - 1000 MHz * 3, 6, 10, 20, and 30 dB standard * Connectors N standard, BNC or TNC available *.



Price 3, 6, 10 dB, $255.00 (1-9 pc.). Price 20, 30 dB, $275.00 (1-9 pc.) **JFW Industries, Inc., 2719 E. Troy Ave., Indianapolis, IN 46203, (317) 783-9785 or INFO/CARD #120.**

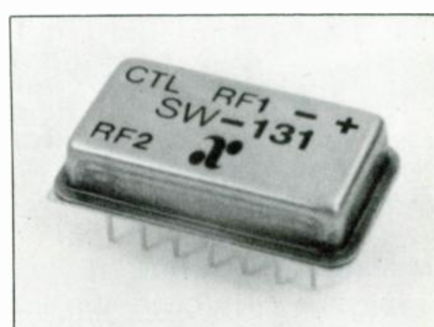
New Audio System

The Grass Valley Group WAVELINK™ family of Fiber Optic Communication products has added two channel audio with video capability, using a unique FM-on-FM modulation scheme. The system modulates two pre-emphasized base band audio signals in 100 kHz oscillators which in turn frequency modulates two oscillators at 8.0 and 9.8 MHz. The signal is then summed with video and transmitted using an LED. The transmitter module can provide for equalization of incoming cable, adjustable gain, and separate output monitor amplifier. The receiver utilizes an avalanche photo diode and features AGC and squelch functions. Delay timing adjustment is available for precise timing of signals. The FM signal is demodulated and passed through a base band filter to restore the original signal. Performance specifications are designed around broadcast requirements and include unweighted signal-to-noise better than 58 dB for video and 65 dB for audio over a 2 km fiber. The maximum output is 20 dBm into 600 Ohms. The transmitted wavelength of the LED is 820 nm. 10 μ of optical power is coupled into the fiber. **The Grass Valley, Inc., P.O. Box 1114, Grass Valley, California 95945, (916) 273-8421 or INFO/CARD #119.**



Broadband RF Switch

Adams-Russell, Anzac Division has developed an SPST switch to cover the 200-2000 MHz frequency range in a single unit. The SW-131 provides 50 dB isolation with only 1.0 dB of insertion loss. The switch has an integral TTL driver and is packaged in an



18 pin dual in-line hermetic case. The SW-131 is designed for military environments and can be screened to MIL-STD-883. Other features of the switch

include a +13 dBm 1 dB compression point, a 1.3:1 VSWR and a 2 μsec switching speed. Delivery is from stock and unit pricing is $235.00 in small quantity. **Adams Russell, Anzac Div., 80 Cambridge St., Burlington, MA 01803, (617) 273-3333 or please circle INFO/CARD #121.**

New Literature

Precision Connector Catalog

Gilbert Engineering has announced a new precision connector catalog describing its line of connectors, terminations, attenuators, and coupling elements. **Gilbert Engineering, 5310 W. Camelback Rd., P.O. Box 23189, Glendale, AZ 85301-7597, (800) 528-5567, (602) 245-1050, or please circle INFO/CARD #118.**

TVRO Products Brochure

This new 20-page brochure describes Merrimac's complete line of products for the TVRO frequencies 3.7-4.2 GHz. The products included are the SR-1 Satellite Video Receiver, Single and Dual-Conversion Down-Convertors, Standard Power Dividers, Isolators, Fixed Attenuators, and

Terminations. **Merrimac Industries Inc., 41 Fairfield Place, West Caldwell, NJ 07006, (201) 575-1300, or INFO/CARD #117.**

Coax/Waveguide Measurement Accessories Catalog

The latest edition of Hewlett-Packard's Coaxial and Waveguide Measurement Accessories Catalog is now available with product information on more than 350 microwave measurement accessories used in coaxial and waveguide measurements to 40 GHz and above. Product sections include attenuators, detectors, couplers, filters, power sensors, slotted lines, network analysis and noise-figure equipment, and a good selection of 75-ohm items. **Contact your local Hewlett-Packard sales office or please circle INFO/CARD #114.**

Filter and Networks Catalog

A catalog covering a complete line of Monolithic Crystal Filters, Discrete Crystal Filters, and LC Filters and Networks is now available. In addition to OPT's custom capabilities for both military (MIL-F-18327) and commercial application, OPT has cataloged thousands of engineering and manufacturing designs covering more than twenty years of know-how. As a result, all products offered in the catalog have been manufactured, accepted and are now operating in many advanced military programs as well as high quality communication systems. The filters are stable over a wide temperature range and withstand shock and vibration. **OPT Industries, Inc., 300 Red School Lane, Phillipsburg, NJ 08865, (201) 454-2600, or INFO/CARD #115.**

Phase Shifters and Attenuators Catalog

A 17 page catalog describes Triangle Microwave Inc. product line including phase shifters, and attenuators. **Triangle Microwave Inc., 60 Okner Parkway, Livingston, N.J. 07039, (201) 740-0100, or INFO/CARD #116.**

Test Equipment Catalog

A new short-form catalog of electrical test equipment is now available from Beckman Instruments, Inc. The four-page bulletin provides principle specifications and operating characteristics for insulation and dielectric breakdown testers, high voltage schering bridges, megohmmeters, liquid power factor testers, and high voltage power supplies. Also featured are dielectric withstand test sets, wheatstone bridges and resistance decade boxes. **Beckman Instruments, Inc., Cedar Grove Operations, 89 Commerce Road, Cedar Grove, N.J. 07009, (201) 239-6200, or INFO/CARD #113.**

Filter Catalog

Corcom, Inc. has just released its new 36-page RFI power line filter catalog. This catalog contains more than 200 different filter models. Each series is comprehensively described and provides complete specifications for every type model with mechanical dimensions given in both the English and metric systems. There are complete electrical schematics, 50 ohm insertion loss data, application suggestions, as well as detailed international safety agency requirements

and approval information. The catalog also contains an extensive technical section covering all phases of filter technology from the basic, "What is radio frequency interference?" to complex information and procedures on RFI measurements. One of the most unique features of this new catalog is a "pull-out" RFI Power Line Filter Sector Chart to assist the user in matching his requirements with the proper filter. **Corcom Inc., 1600 Winchester Rd., Libertyville, IL 60048, (312) 680-7400, or INFO/CARD #112.**

Engineering Bulletin

Sprague Electric Company has introduced a 50 mil × 40 mil multilayer ceramic capacitor chip in its Type 15C high-frequency capacitor series to meet field problems with the use of the more common 50 mil × 50 mil chip. Users requests to have a chip which can be easily oriented on a substrate is met by the new rectangular Sprague chip, which is intended to replace the square 50 mil × 50 mil chip previously used. Sprague Electric has made the same capacitance values available on the smaller rectangular chip as were available on the slightly larger 50 mil square chip. Complete listing of all standard capacitance ratings, as well as performance on Sprague Type 15C multilayer ceramic chip capacitors for high frequency applications is given in Engineering Bulletin 6200. 51A,

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INFO/CARD 39

available from the **Technical Literature Service, Sprague Electric Company, Marshall St., North Adams, Mass., 01247 or INFO/CARD #111.**

Comprehensive New Standard Line Catalog

A 286-page Standard Line Connectors Catalog from ITT Cannon details microminiature, hermetically sealed,

filter, printed circuit, circular, rack/panel connectors and more. The new 1982-1983 edition offers illustrations and full specifications on the complete line of Cannon electrical and fiber optic connectors. **ITT Cannon, Marketing Department, 10550 Talbert Avenue, Fountain Valley, CA 92708, or please circle INFO/CARD #110.**

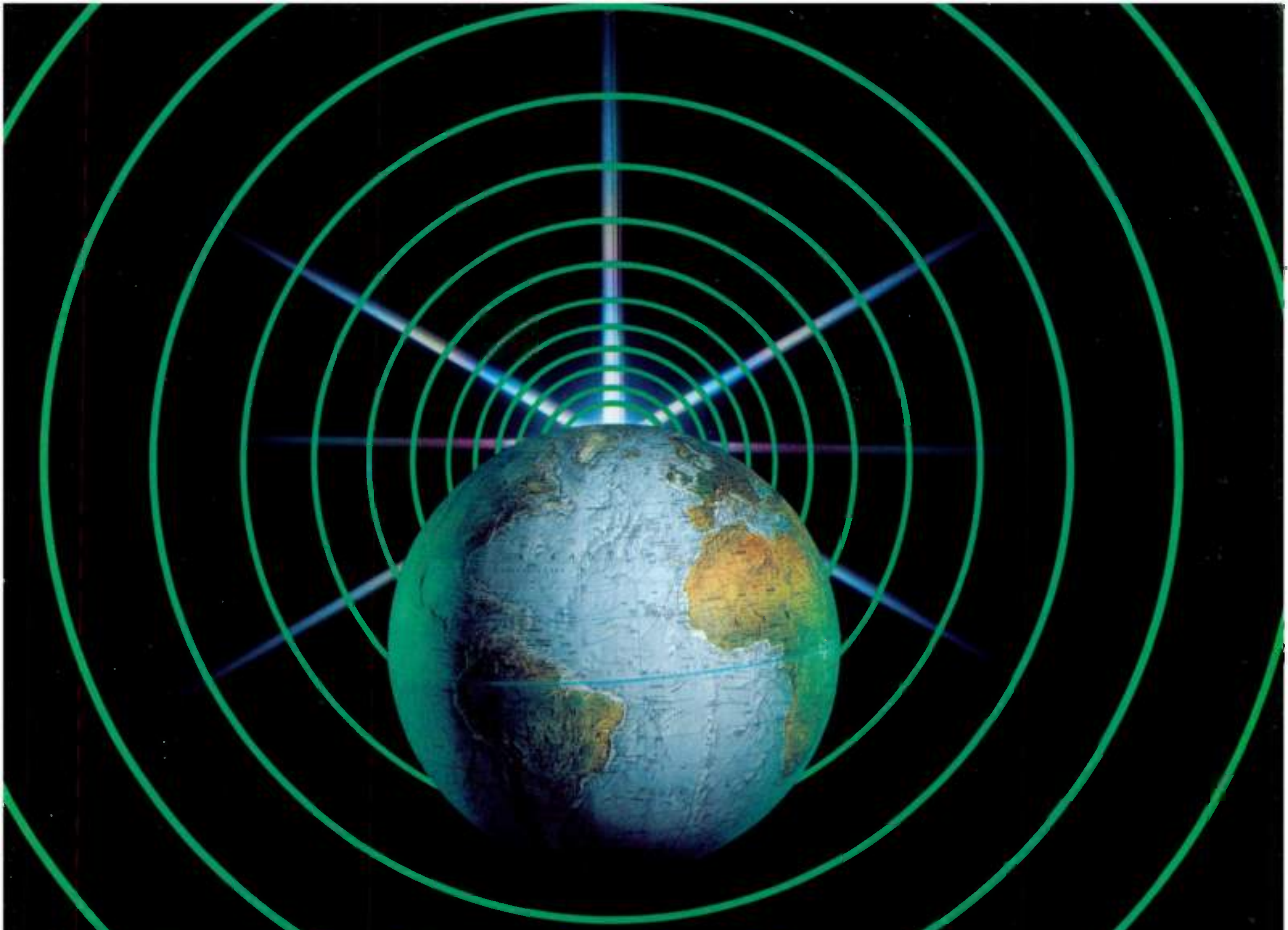
Susceptibility Testing Antennas Booklet

A comprehensive booklet detailing the new products in its susceptibility testing antenna and accessory line is now available from Amplifier Research, Souderton, PA. *Antennas, TEM Cells, and Accessories for RFI susceptibility testing* covers AR antennas (10 kHz to 1 GHz, up to 3500-watt input) for both shielded-room and outdoor testing. It presents performance curves and specifications, including those on a new 30-1000 MHz "cavity exciter." The company's TEM cells, isotropic field-sensor systems, and fiber-optic telemetry systems are thoroughly discussed as well. The booklet aids in establishing equipment needs for a broad range of RF susceptibility testing requirements. Performance curves can be easily matched to those of compatible broad-and narrow-band rf amplifiers. **Amplifier Research, 160**

School House Road, Souderton, PA, 18964, (215) 723-8181, or please circle INFO/CARD #109.

Application Note

Pacific Measurements has just released a new application note (AN-21) and a 16-page, updated technical brochure for its 1038-N10 Swept Measurement System. **Pacific Measurements, Inc., 488 Tasman Drive, Sunnyvale, CA 94086, (408) 734-5780 or please circle INFO/CARD #107.** ☐



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7103	3.7— 4.2 GHz
7114	5.9— 6.5 GHz
7144	6.5— 7.2 GHz
7115	10.5—12.7 GHz

RADAR

Model	Frequency
7123-S1	2.2— 2.7 GHz
7113	2.7— 3.5 GHz
7134	4.0— 4.5 GHz
7124	5.3— 6.0 GHz
7105	8.0—10.0 GHz

ELECTRONIC WARFARE

Model	Frequency
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7206	2.0—18.0 GHz

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