March/April 1982

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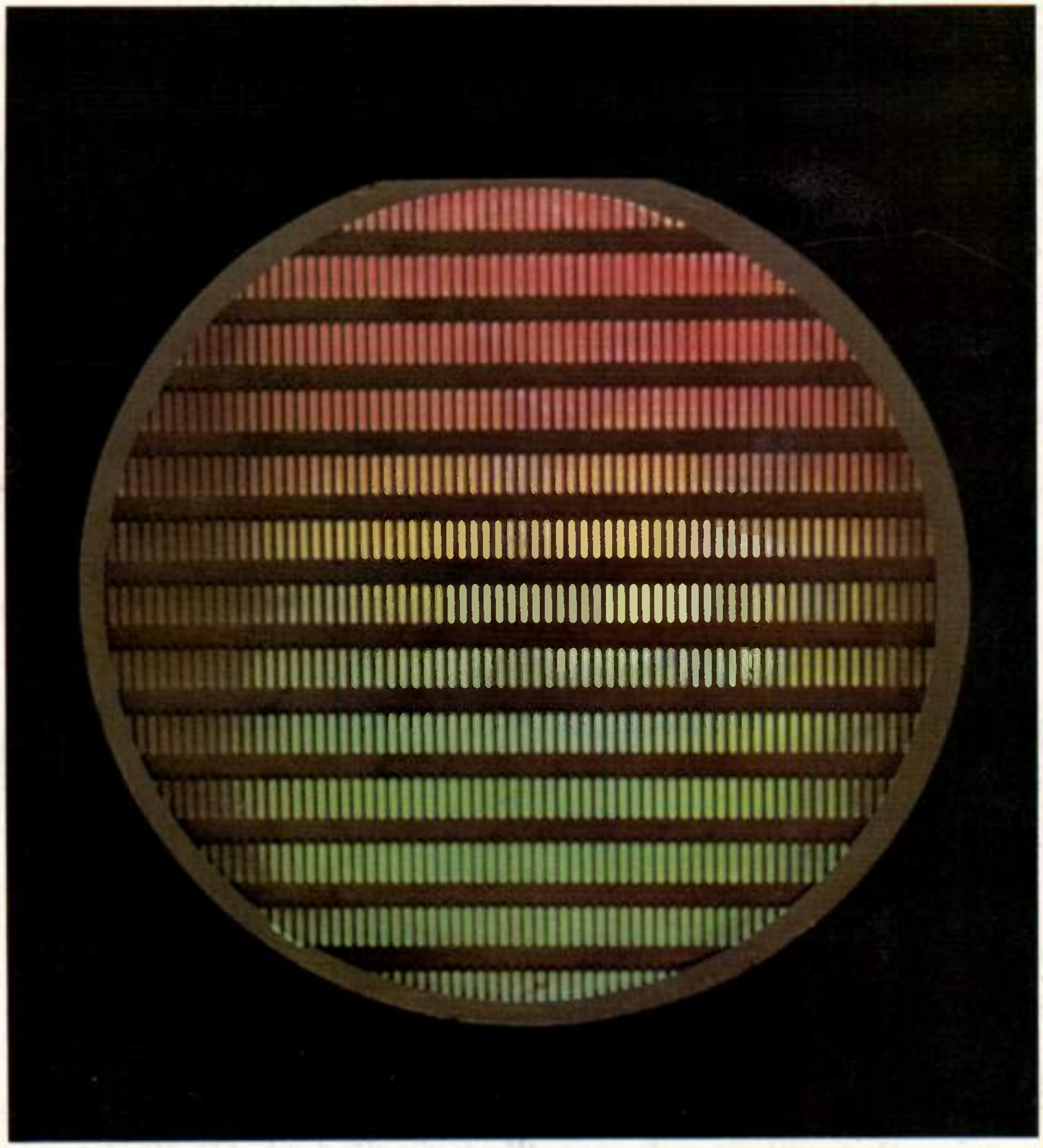

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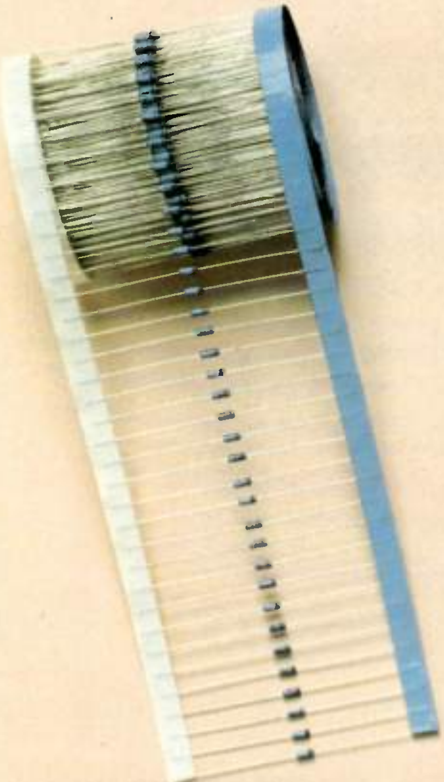

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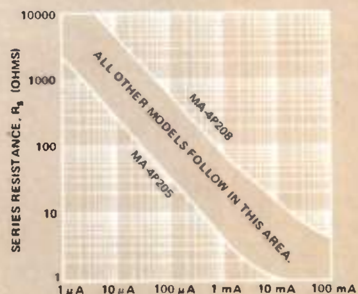

TYPICAL CHARACTERISTICS

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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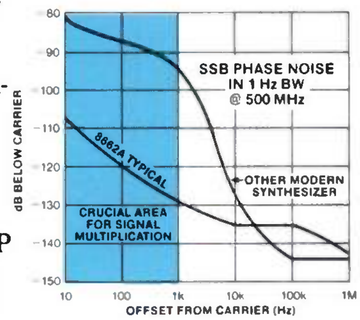

New Application Note Available. To find out how HP achieved this spectral excellence and more information on the 8662A, write for AN 283-1, "Applications and Measurements of Low Phase Noise Signals Using the 8662A." Hewlett-Packard Company, 1820 Embarcadero Rd., Palo Alto, CA 94303.

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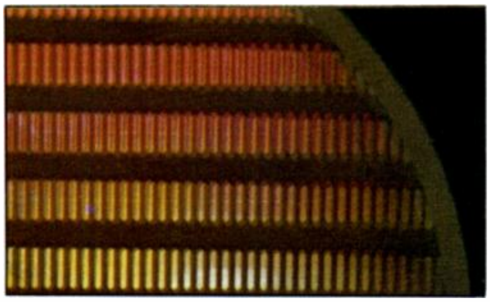
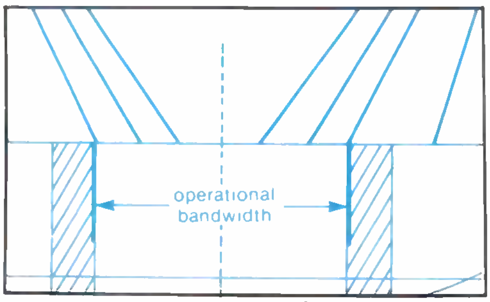


SAWS!



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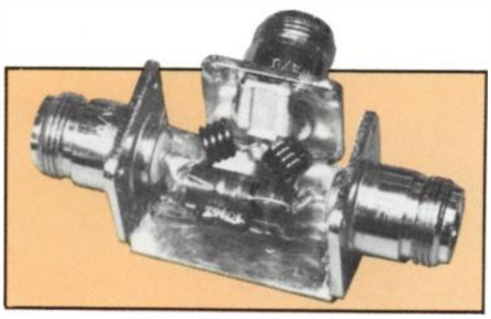

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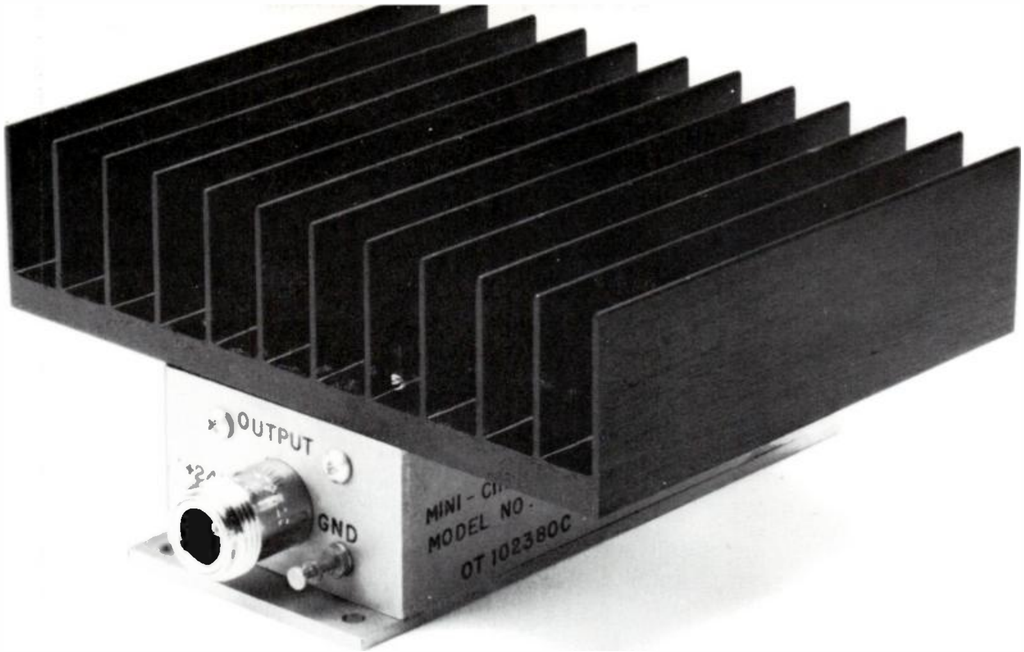

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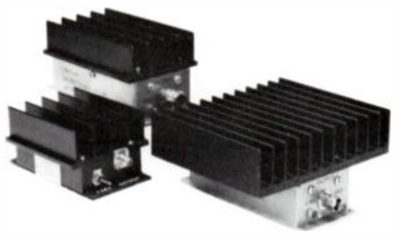

* Model No.	Freq. MHz	Gain dB	Gain Flatness dB	Max. Power Output dBm 1-dB Compression	Noise Figure dB	Intercept Point 3rd Order dBm	DC Power Voltage	DC Power Current	Price $ Ea.	Price Qty.
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ZHL-3A	0.4-150	24 Min.	±1.0 Max	+29.5 Min.	11 Typ.	+38 Typ.	+24V	0.6A	199 00	(1-9)
ZHL-1A	2-500	16 Min	±1 0 Max	+28 Min	11 Typ	+38 Typ.	+24V	0 6A	199 00	(1-9)
ZHL-2	10-1000	15 Min.	±1 0 Max	+29 Min	18 Typ	+38 Typ.	+24V	0 6A	349 00	(1-9)
ZHL-2-8	10-1000	27 Min.	±1 0 Max	+29 Min	10 Typ	+38 Typ.	+24V	0.65A	449 00	(1-9)
ZHL-2-12	10-1200	24 Min.	±1 0 Max	+29 Min *	10 Typ	+38 Typ	+24V	0 75A	524 00	(1-9)
ZHL-1-2W	5-500	29 Min	±1 0 Max	+33 Min	12 Typ	+44 Typ.	+24V	0 9A	495 00	(1-9)

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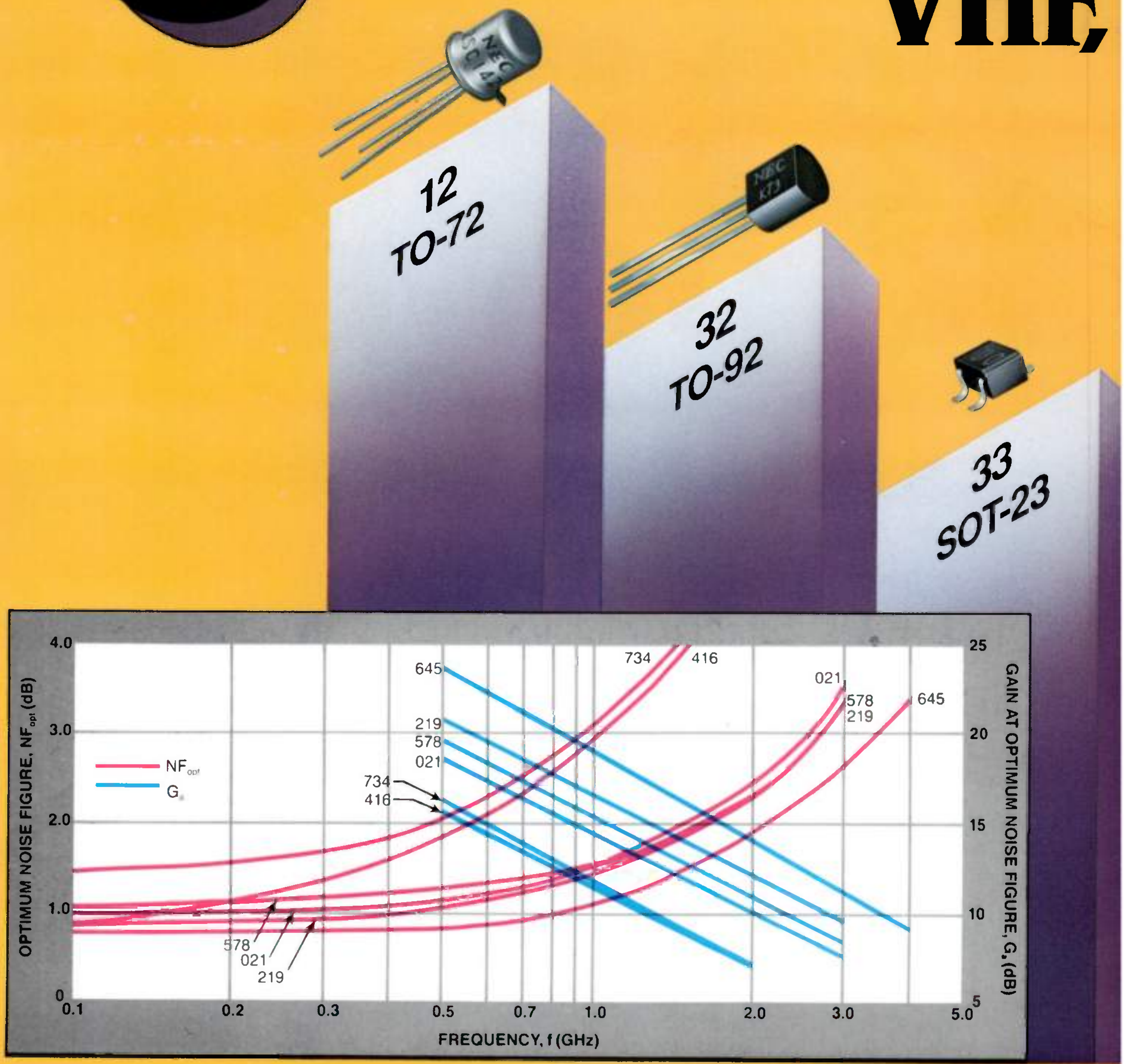

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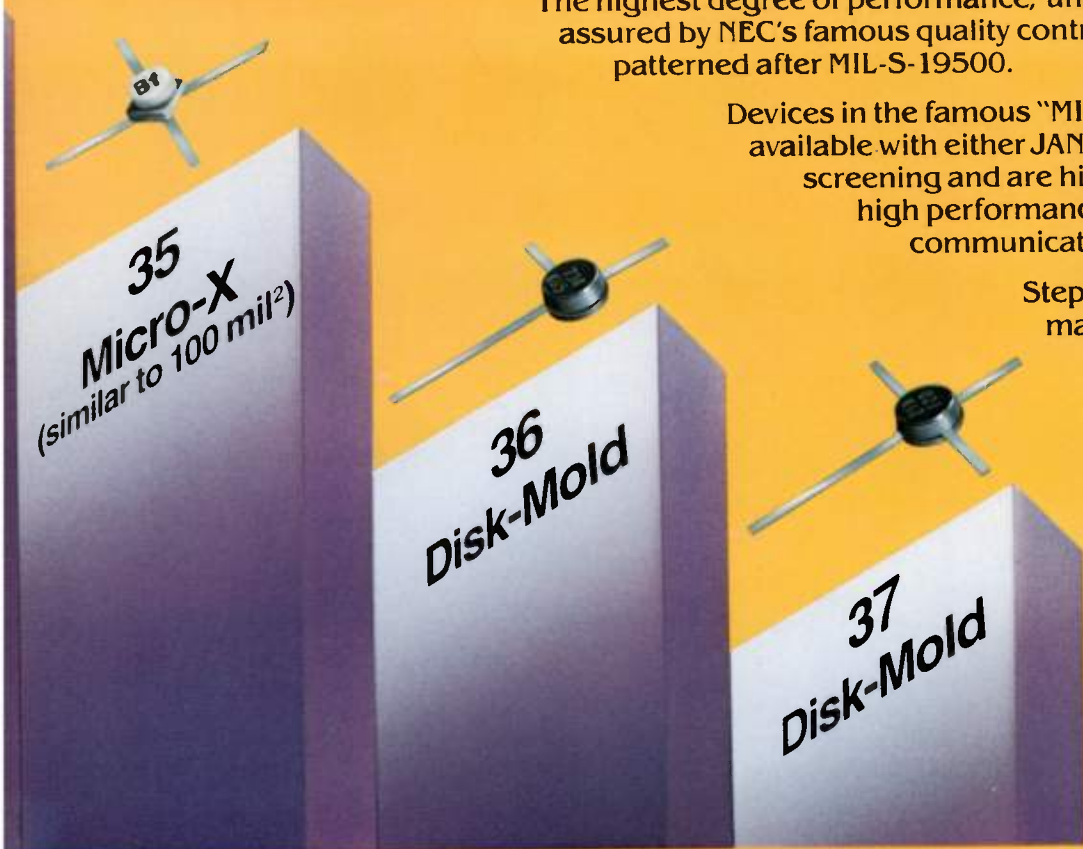

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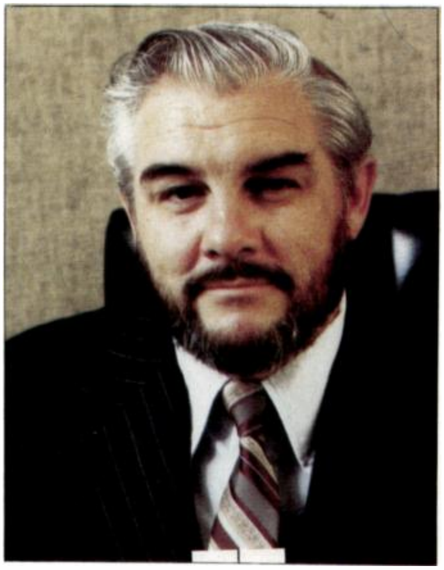
Looking Ahead

Each year, most of us look forward to another year; and some of us initiate plans that will bring us toward certain year-end goals. This year, we have been a few months late in getting our 1982 plans off to a timely start. I am, however, now in the position to share with you some 1982 planning that I am very pleased with.

First of all, with the last issue of the magazine and in all future issues, you will begin to see pleasing changes in the magazine. More color and larger issues are just two of the many goals established for *R.F. Design* magazine. Naturally, we plan to maintain the high quality you have come to expect in the articles throughout the year.



The last three issues of 1982 will be especially interesting. In addition to the usual menu of good RF technical articles, each issue will provide an opportunity for you to win a $1000 U.S. Savings Bond by understanding what makes your fellow engineers read advertisements and technical articles.

- In July/August, a contest to rank the ads in order of readership
- In September/October, a contest ranking technical articles in order of helpfulness
- In November/December, a "super contest" to design a circuit or system using items advertised.

In 1982, I am going to look forward to bringing you better issues of *R.F. Design* magazine. Also, if you are in Boston for Electro 82, please stop by our booth, pick up a fresh issue of the May/June magazine, and meet our new and expanding staff.

Thank you.

Bill W. Childs
Publisher

Interesting Problem

Dear Sir:

Is there any way we can appeal to your readers for assistance in the following problem?

Motorola has discontinued manufacturing the MPS-H83 transistor, a UHF pnp AGC-able device having no second source or equal alternative. Motorola's stock and all distributor stocks are used up. We received no notice of this, and badly need about 500 pieces quickly to tide us over until a redesign can be completed.

Perhaps somewhere on a shelf are 500 such transistors no longer needed by the owner. Do you know of any route through which we might contact such a surplus stock, if it exists? Motorola has not been able to help us.

J.W. Streater
Chief Engineer
Mentor Radio Company
1561 Lost Nation Road
Willoughby, Ohio 44094
(216) 942-2025

Stability Circles

Dear Editor:

In Marty Jones' recent article, "High Frequency Transistor Amplifier Design," Nov./Dec. 81 issue, the author discusses stability circles and states that "an individual impedance point must be evaluated to determine whether the interior of the circle is the unstable region." Actually, the origin of the Smith Chart serves this purpose ideally. Since the stability circles describe a boundary condition, all points inside are stable (unstable) if any point inside is stable (unstable). From the non-unilateral port reflection equations:

$$S'_{11} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \Bigg| \begin{array}{l} = S_{11} \\ \Gamma_L = 0 \text{ (50}\Omega\text{ load)} \end{array}$$

$$S'_{22} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} \Bigg| \begin{array}{l} = S_{22} \\ \Gamma_S = 0 \text{ (50}\Omega\text{ source)} \end{array}$$

Assuming the transistor (or two-port) was stable when characterized in a 50 Ω system, $|S_{11}| < 1$ and $|S_{22}| < 1$. In other words, the origin represents a stable termination. It is then safe to say that, for such a two-port, if the stability circle encompasses the origin, all points within the circle represent a stable termination. If the circle does not enclose the origin, all points inside are unstable.

Hal Hamilton
Sr. Staff Engineer
Motorola Government Electronics Division

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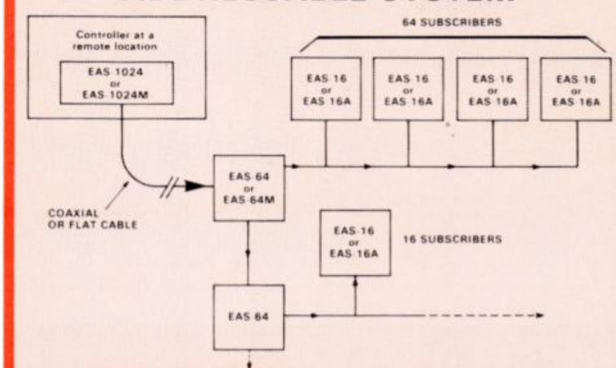

EAS The above system is composed of 3 units — a microprocessor control (EAS-1024); a decoder (EAS-64); and a wide-band, multitap switch assembly (EAS-16). The system can be installed in 2 alternative configurations and is most compatible with other systems.
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SAWS!

By Darrel L. Ash
Vice President Engineering
and Richard H. McLean
Vice President Sales
RF Monolith
Dallas, Texas

When Surface Acoustic Wave devices entered the marketplace some 15 years ago, first applications were found in certain government-related programs where the initial high cost could be justified by the size/performance advantages they offered.

Tediosly fabricated and produced in small quantities, outstanding performance was achieved in applications such as radar signal processing and high resolution countermeasures receivers. Further applications have been successful in major U.S. Defense Programs, making use of SAW delay line oscillators and SAW correlators for spread-spectrum communications.

During this period, low-frequency SAW filters found use in television receivers in large quantities and at very low cost.

These "First-Generation" series of designs generally resulted in size-efficient filters, generally below 100 MHz

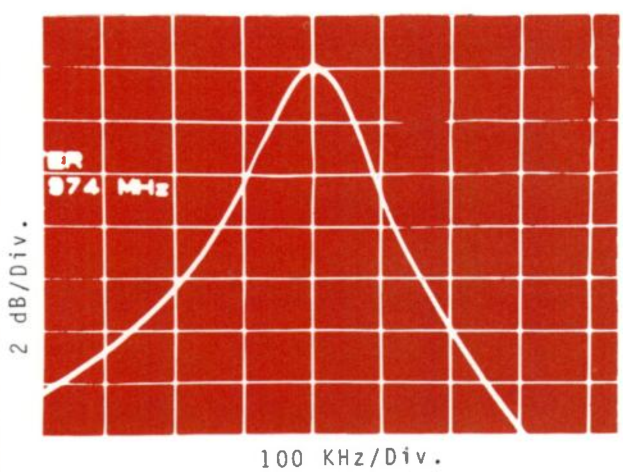


Figure 1A. SAW Resonator Frequency Response.

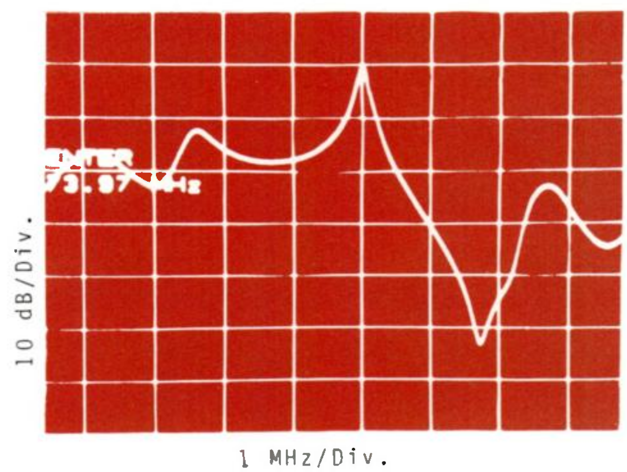


Figure 1B. SAW Resonator Frequency Response.

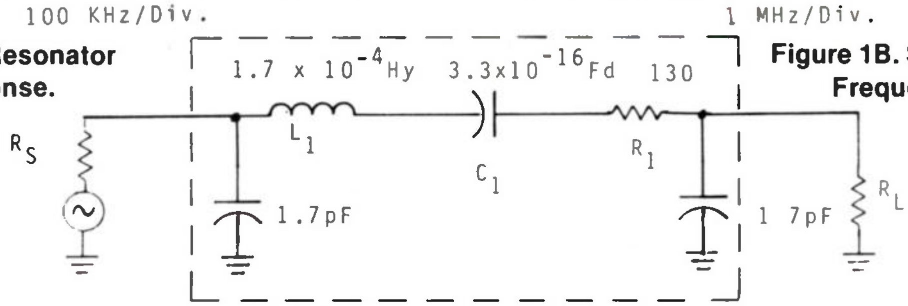


Figure 1C. SAW Resonator Equivalent Circuit.

POWERFUL PASSIVES

Second generation SAW filters are providing powerful new alternatives in circuit design.

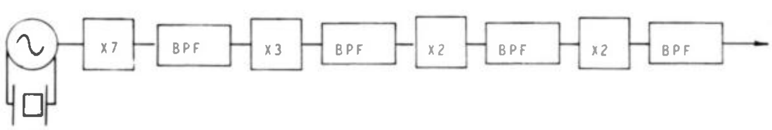


Figure 2A. Conventional Crystal Oscillator & Multiplier Chain.

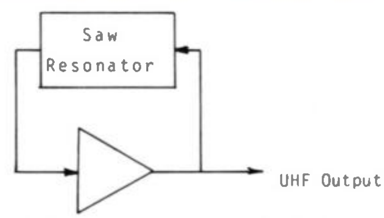


Figure 2B. SAW Resonator Oscillator.

and usually exhibiting insertion loss figures in the 25 dB area. Given this background, a new awareness began among communications engineers of the benefits in using the technology, requiring less than the exotic designs necessary for reflective array correlators but more powerful devices than the mass-produced video IF TV filters.

This discussion, then, will address the fastest growing segment of SAW applications, generally defined by the frequency boundaries of 40 MHz to 1000 MHz which are being placed into use by the commercial communications industry.

These newer designs may be termed "Second-Generation," in that filters may be produced in the mid-100's frequency range, smaller in size and with much reduced signal loss than their first generation counterparts. Also defined by the second generation label are SAW resonators for use as the frequency-control element for size-and-cost-efficient oscillators, operating directly at frequencies to 1 GHz, while at the same time eliminating crystals, multiplier chains and amplifier stages.

SAW Resonator Applications

A SAW Resonator consists of two reflective gratings forming a resonant cavity with transducer coupling of energy into and out of the cavity. The typical unloaded Q range is from 50,000 to 5,000 depending on center frequency, reflector array size, and other factors. Center frequencies up to 1000 MHz are attainable within that Q range. Figure 1 includes photographs of the frequency responses of a two port 674 MHz resonator, presently being mass produced, as well as its equivalent circuit. The equivalent circuit (within the dash lines) consists of a series resonant circuit with a series Q-determining resistor and shunt capacitances to ground at each port.

The high Q and low spurious modes exhibited by SAW resonators make them ideal frequency references for oscillator circuits. Figure 2A is a block diagram of a crystal oscillator and frequency multiplier chain which is typical of the approach presently used to obtain a stable, fixed UHF output for local oscillator or transmitter applications. Figure 2B is a block diagram of a SAW resonator oscillator with the same UHF output. Figure 3 is a schematic diagram of a SAW oscillator circuit presently in mass production

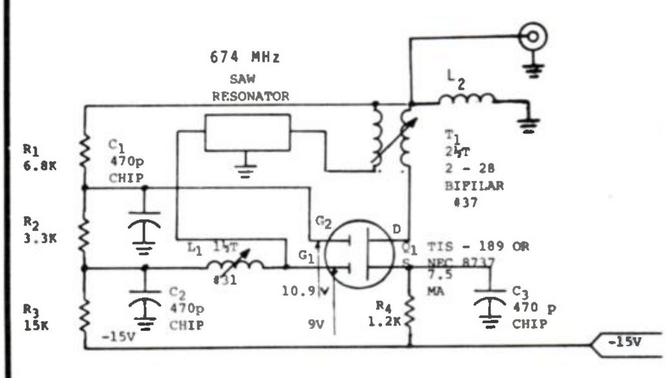


Figure 3. Present SAW Resonator Oscillator.

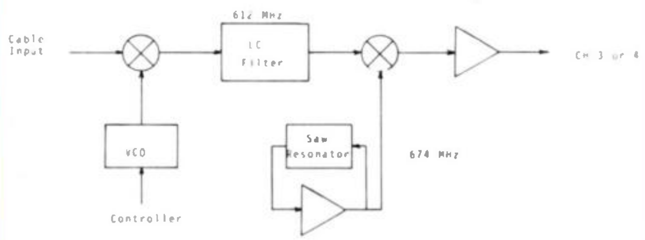


Figure 4. Cable TV Converter.

which accepts resonators with center frequencies from 674 MHz to 680 MHz. The simplicity of the SAW resonator oscillator compared to the crystal oscillator and multiplier chain is obvious. A single transistor amplifier with simple tuned input and output circuits is all that is required in addition to the SAW device to realize a stable UHF oscillator.

Use in Cable TV

The circuit of Figure 3 is presently being used to provide a stable second local oscillator for cable TV converters as illustrated in the block diagram of Figure 4. A stable second local oscillator is needed in cable converters to maintain an output within the AFC pull-range of television receivers using frequency synthesizers and also to interface with descrambler circuits. A 674 MHz L.O. at the second mixer produces a converter output on Channel 3.

Referring to Figure 3 again, the SAW resonator used in this circuit has a phase shift of 0° at the resonator center frequency. The transformer, T1, and inductor, L1, are adjusted to yield a 0° phase shift around the oscillator loop. T1 consists of an air core bifilar-wound coil which couples a portion of the output power back to the amplifier input through the SAW device. A dual gate FET device is used for the amplifier to present a large Q-reducing impedance to the resonator. This particular circuit can be pulled more than 200 kHz by adjusting T1. One disadvantage of this circuit is that it will oscillate only on the high side of the resonator response. Excess phase shift in the transistor (approximatey 240° rather than 180°) must be compensated for by using part of the phase slope of the resonator device on the high side of resonance to obtain 0° phase shift around the oscillator loop, since the external tuning elements have too much attenuation at the required phase shift.

The more recent oscillator circuit of Figure 5 can be pulled through the resonator center frequency to either side of resonance. This oscillator uses a resonator device which exhibits a 180° phase shift at its center frequency. The series inductors, L1, and L2, in conjunction with the transistor input and output capacitances and the input and output capacitances of the resonator yield the necessary phase shift aroundthe oscillator loop to compensate for the excess phase shift in the transistor. This oscillator can also be pulled more than 200 kHz by adjusting L2.

Other Uses for the Resonator

The block diagram of Figure 6 illustrates another use for such a SAW stabilized oscillator. A very low cost, single transistor, low power (>0 dBm), stable, FSK transmitter can be realized by adding a varactor diode to pull the oscillator in response to a modulation source.

SAW resonators can also be used as narrow band band-pass filters. The block diagram of Figure 7 illustrates how a narrow band receiver can be greatly simplified by using a SAW resonator filter. Most narrow band receivers cannot take advantage of low cost 455 kHz IF filters for the first IF because the resulting image frequency at 910 kHz is too close to the desired frequency to be filtered out in the RF filter in front of the first mixer. Thus, double conversion is usually used in such receivers (10.7 MHz for the first IF and 455 kHz for the second IF). Referring to Figure 1B, the notch on the high side of the resonator response can

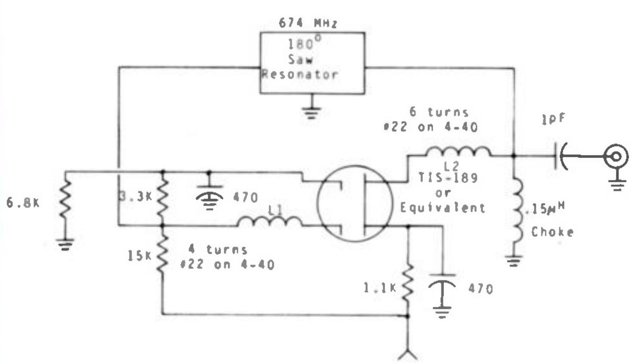


Figure 5. New SAW Resonator Oscillator.

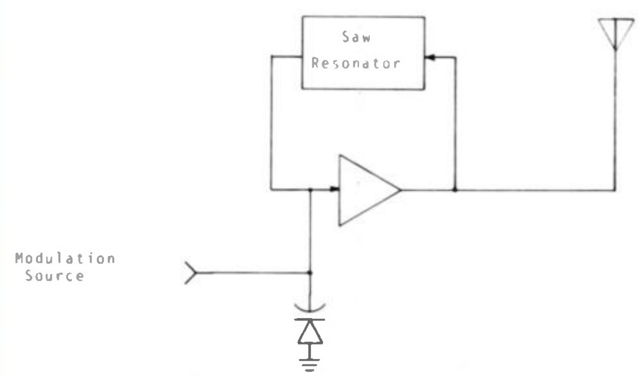


Figure 6. Low Cost Transmitter.

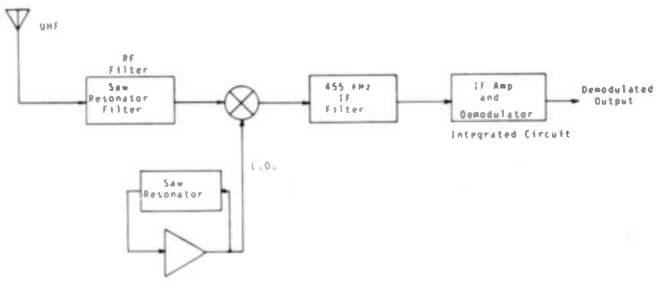


Figure 7. SAW Resonators in Simple Narrowband Receivers.

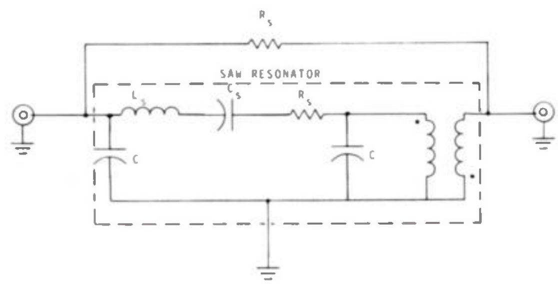


Figure 8. SAW Resonator Notch Filter.

be placed 910 kHz above the response by a simple design change. Thus, an image rejection of ~50 dB would be obtained at RF with a single resonator filter. Of course, a SAW resonator could also be used for the local oscillator rather than a crystal oscillator and multiplier chain which would further simplify the receiver. The result is a very low cost, high performance, narrow band receiver.

Notch Filter

Figure 8 schematically illustrates how a SAW resonator with 180° phase shift at center frequency can be used in a narrow notch filter configuration. A feedback resistor is added whose resistance is equal to the internal resistance of the resonator. This produces a null at the resonator center frequency. The pass-band insertion loss is the loss of the feedback resistor. A lowpass filter network making use of the resonator shunt input and output capacitance can be added to the simple circuit of Figure 8 to further decrease the passband insertion loss. The response of such a notch filter at 674 MHz is shown in the photographs of Figure 9. The notch depth was 65 dB and the pass-band insertion loss was 3 dB.

Determining Resonator Bandwidth

Roughly stated, loaded Q (Q_L) will be approximately equal to resonant frequency divided by bandwidth, and can be calculated from the RLC relationship shown in Figure 1C. Since the equivalent circuit for the resonator is a series resonant circuit, the bandwidth can be increased by increasing the source and/or load resistance, R_s and R_L in Figure 1C.

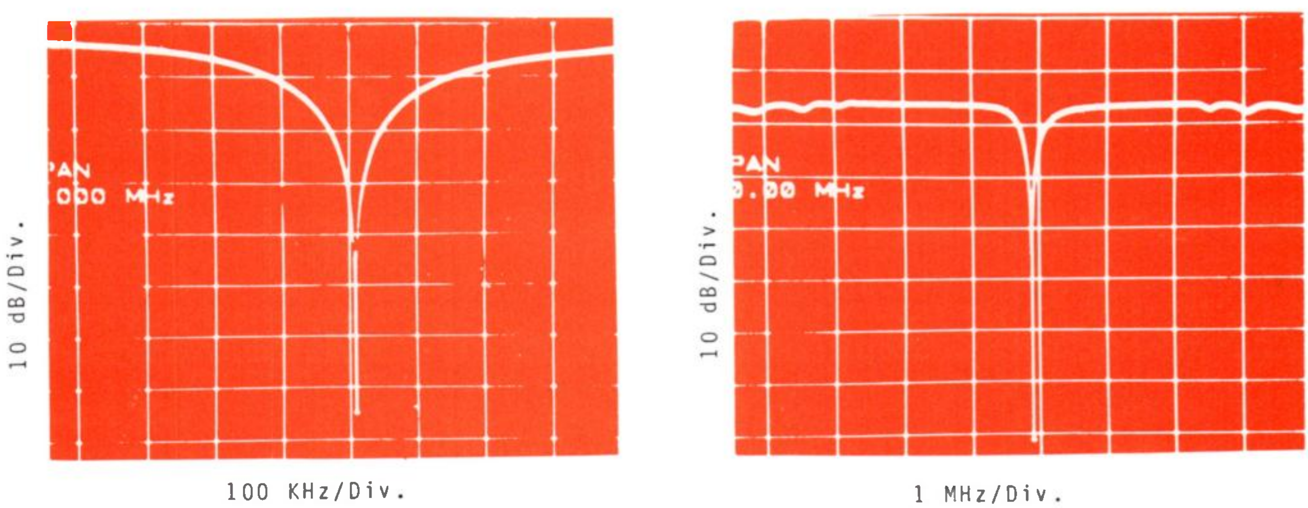
$$Q_u = \mu L/R_1 \text{ where } \mu = 2\pi fo$$

Loaded Q (Q_L) appears as:

$$Q_L = \mu L/(R_1 + R_s + R_L)$$

Saw Filters

The SAW filters presently in use in TV IF's are simple transducer based filters with high insertion loss (20 dB typical). Such a high loss requires that additional gain blocks beadded in front of the filter to avoid noise figure degradation. However, gain blocks in front of the IF filter are not protected from large interfering signals and are potential sources of cross modulation and intermodulation distortion.



(9A)

(9B)

Insertion Loss = 3 dB

Center Frequency = 674 MHz

Figure 9. SAW Resonator Notch Filter.

Table I
Broadband Band Pass Filters

	Demonstrated	Production	Projected Practical Limits
Center Frequency	1.0 MHz-2.75 GHz	10 MHz-1.0 GHz	10 MHz-2.0 GHz
Minimum Insertion Loss	0.65 dB	2.0 dB	1.0 dB
Maximum Fractional Bandwidth	100%	50%	100%
Sidelobe Rejection	70 dB	60 dB	90 dB
Minimum Bandwidth	100 kHz	100 kHz	50 kHz
Minimum Transition Bandwidth	100 kHz	100 kHz	50 kHz
Minimum Shape Factor	1.15	1.2	1.1
Triple Transit Suppression	55 dB	45 dB	60 dB
Amplitude Ripple	±0.02 dB	±0.05 dB	±0.01 dB
Phase Deviation from Linear	±0.1°	±2°	±0.1°

New Designs — New Applications

A second generation of mass producible SAW filters is now available. These filters offer low insertion loss, low triple transit distortion, independent tailoring of magnitude and phase response, excellent out-of-band rejection, and center frequencies approaching 1 GHz. Table I is a summary of absolute minimums and maximums both demonstrated and projected for this second generation of filters.

These devices are three-phase transducer based filters with three electrodes per wavelength on the substrate driven 120° out of phase. As a result, the matching network is more complex than that for first generation high loss filters, but the resulting improvement in system performance and the elimination of gain blocks is more than a fair trade-off. Figure 10B includes a schematic diagram of the equivalent circuit of a three-phase unidirectional transducer within the dashed lines. The necessary transformation from a single phase driving source to a three phase driving source is accomplished using a 60° Pi phase shifting network schematically illustrated in Figure 10A. The design of the network is straight-forward, as shown in Figure 10. An L-network is used to obtain the desired impedance match. The resulting matching network consists of three coils and one capacitor for each of the two three-phase transducers.

Wide-band Receivers

Given the low loss, excellent out-of-band rejection, and independent magnitude and phase response exhibited by the unidirectional low-loss filter, as well as a center frequency in the UHF band, a means of simplifying wide band receivers becomes evident. A common problem in broadband systems such as television receivers is spurious

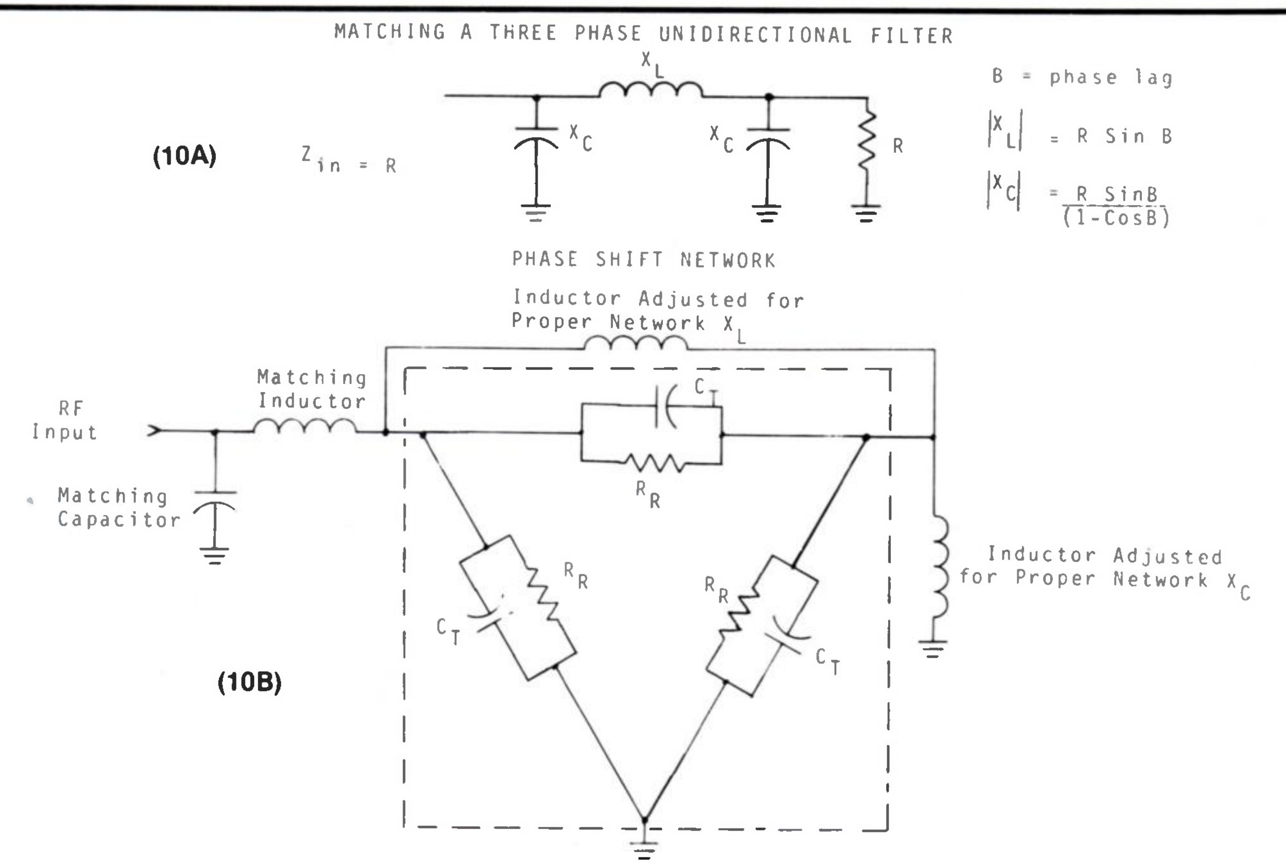


Figure 10. Matching and Phasing for a Three-phase Unidirectional Transducer.

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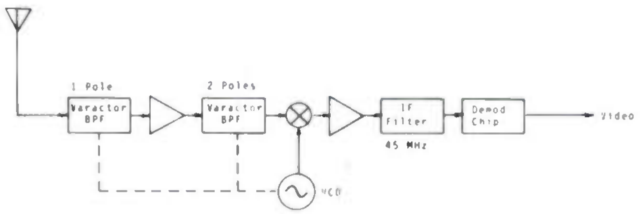


Figure 11. Varactor Tuned TV Receiver.

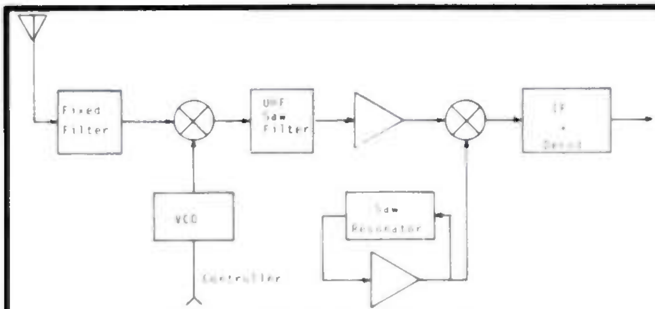
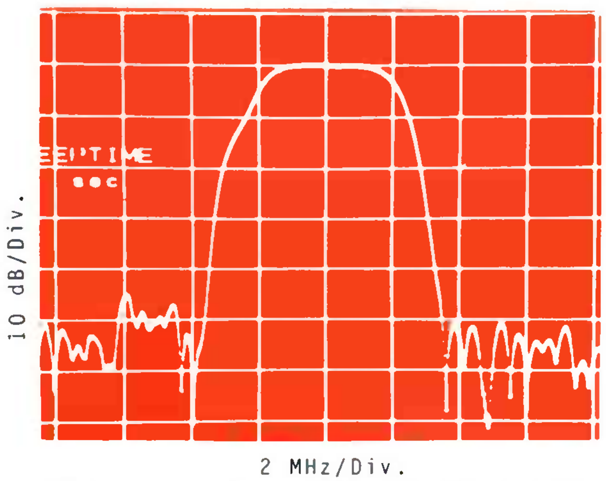


Figure 12. High IF Receiver.



Picture Carrier = 448.725 MHz (-4dB)
Insertion Loss = 6.5dB
Vestigial Weighting
Predistortion Phase Compensation
Sound Carrier = 13dB

Figure 13. FCC Low Loss SAW Filter.

receiver responses to IF related signals. Examples are IF beat, half-IF beat, image response, and local oscillator radiation. This problem is presently addressed using tunable RF filters such as the three pole varactor filters in TV front ends illustrated in Figure 11. However, UHF varactor filters are necessarily broad and have insufficient rejection to eliminate these responses in a system with a 45 MHz IF.

The low loss SAW filter makes it feasible to use a high first IF such as 450 MHz or higher to move the IF related spurious responses out of the desired frequency band. This makes it possible to use a fixed broadband RF filter to eliminate these responses. For example, a 450 MHz IF would have an image response at 900 MHz above the desired RF signal and an IF beat response at 450 MHz above the desired signal. Figure 12 is a block diagram of such a high first IF system. The low loss, UHF SAW filter is capable of rejecting adjacent channels in excess of 50 dB in a TV receiver with an IF of 450 MHz.

As shown in Figure 12, the high IF system would probably make use of double conversion to take advantage of IF gain and demodulator integrated circuits at IF frequencies such as 45 MHz. A SAW resonator oscillator would provide a low cost, stable second L.O. for such a system. The use of a high dynamic range mixer in conjunction with low RF gain minimizes cross modulation and intermodulation in the front end while providing a low system noise figure due to the low loss of the SAW filter. The SAW filter provides the

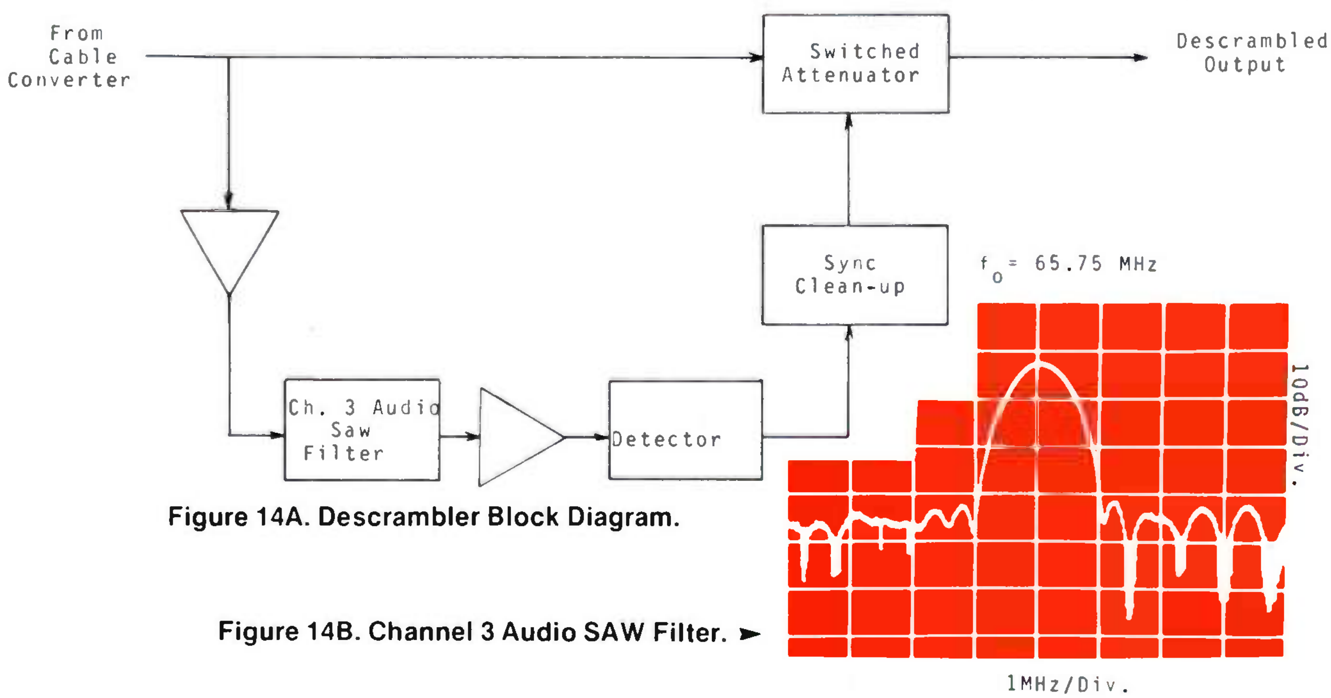


Figure 14A. Descrambler Block Diagram.

Figure 14B. Channel 3 Audio SAW Filter. ►

total filtering function as well as any needed phase compensation for predistortion and the like.

FCC Sponsored Receiver

Under a contract with the Federal Communications Commission, this author developed a high IF receiver to show that the UHF interfering signal handling capability of a TV receiver could be greatly improved with a simultaneous improvement in sensitivity or noise figure. The picture carrier frequency in the first IF was 448.725 MHz and a 402.975 MHz SAW resonator second L.O. converted the picture carrier to 45.75 MHz in the second IF.

The frequency response of the low loss SAW filter used in the first IF is included in Figure 13. The filter was vestigially weighted on the picture carrier side (upper side), phase compensation was included for transmitter phase pre-distortion, and adjacent channel rejection was in excess of 50 dB. The magnitude and phase weighting used for this filter also took into consideration other tuned element responses in the system. For example, the picture carrier is 4 dB down on the upper filter skirt, rather than the standard 6 dB, to compensate for a 2 dB roll off in the tuned circuit on the output of the second mixer. This new system exhibits a large improvement in performance over that obtained with present receivers. The performance parameters for this system will be released by the FCC at a later date.

The Low-loss Filter in Cable TV

Referring to Figure 14, another potential application for a low loss unidirectional filter is in the first IF of a cable TV converter to replace the four pole 612 MHz LC filters presently used. Such a filter would greatly reduce potential intermodulation and cross modulation distortion in the converter's second mixer and the TV receiver by rejecting adjacent channels. Alternatively, a low loss Channel 3 SAW filter can be used to reject adjacent channels on the cable converter output if distortion in the converter second mixer is not a problem but TV receiver distortion is a problem.

Other potential applications for such low loss SAW filters and/or resonator filters include satellite links and RF data modems.

"Standard" SAW Filters Still Working

First generation high loss filters, due to their simplicity, will probably continue to be used for applications where high loss does not adversely affect system performance. One example of this is the 45 MHz head end modulator filter which is used to carefully shape the spectral characteristics of cable TV signals. Another function which can be satisfied using a high loss SAW filter is that of a Channel 3 sound carrier filter in a TV descrambler circuit as illustrated in Figure 14A. In such a sync suppression system, the horizontal sync information is derived from the sound carrier. The filter is necessary to avoid interference from the video components of the TV signal which are too close to reliably filter out in mass production with LC filter techniques. Figure 14B is a photograph of the frequency response of a SAW filter designed for this purpose.

Many of these new uses for SAW devices were not even under consideration just a short time ago. As the technology matures, designers feel more secure in specifying them and gaining the improvements offered. No longer laboratory curiosities, SAW components have come of age and can be put to work in all of the above applications as well as many yet to be determined. □

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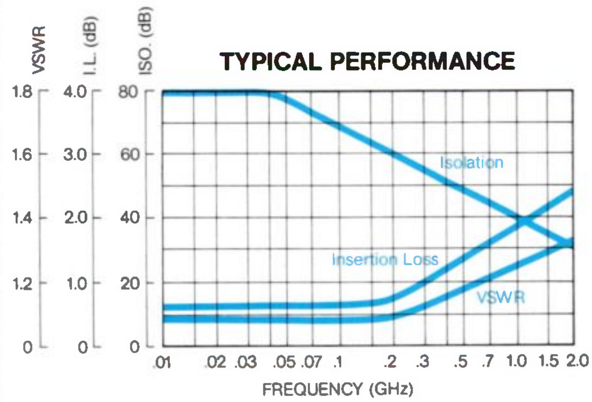

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Size	14 PIN DIP	1-1/2"x 2"x 7/16"
Control	1 line TTL	
Switching Speed	2 microsec. max.	
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R.F. Power	+10 dBm max.	
Impedance	50 ohms	
Unused port terminated in	50 ohms	

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INFO/CARD 8

Design Of RF Amplifiers

Part I: Using Potentially Unstable Devices

Dealing with the "Catch 22" of conjugately matching both the input and output of a potentially unstable device.

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RF designers must often design an amplifier with a potentially unstable device. It is impossible to conjugately match both the input and output of a potentially unstable device.[1] The designer must therefore, calculate Y_S and Y_L for the device, as shown in Figure 1, which insure stability and yield high gain. For an active device whose Y parameters are known, the stability of the device under worst case conditions (both input and output ports open circuited) can be investigated by calculating the Linvill stability factor C.

$$C = \frac{|y_f y_r|}{2g_i g_o - \mathrm{Re}(y_f y_r)} \qquad \text{Eq. (1)}$$

$$y = \begin{bmatrix} y_i & y_r \\ y_f & y_o \end{bmatrix} = \begin{bmatrix} g_i + jb_i & g_r + jb_r \\ g_f + jb_f & g_o + jb_o \end{bmatrix}$$

where || and Re( ) denote the magnitude and the real part of the complex quantity respectively. If $0 \leqslant C < 1$, the device is inherently stable and optimum terminations which yield maximum gain can be directly calculated from formulas.[2] Often, however, C is not between 0 and 1 and the device is potentially unstable.

Stern developed a procedure by which an amplifier can be designed which

TI 59 programs are provided which calculate device terminations and transducer power gain for an RF amplifier with specified stability. The programs use a modified Stern's procedure which can be used with either Y or S parameters.

delivers maximum power gain for a given degree of stability.[3] Stern's procedure was derived using Y parameters, but can easily be adapted for S parameters by performing an S to Y parameter conversion. The Stern stability factor, K is given by

$$K = \frac{2(g_i + G_S)(g_o + G_L)}{|y_f y_r| + Re(y_f y_r)} \qquad \text{Eq. (2)}$$

For K greater than 1, the amplifier is stable. The circuit is potentially unstable for K less than 1.* Design with potentially unstable devices by Stern's procedure basically involves a tradeoff of gain for a higher degree of stability with K generally chosen between 4 and 10.

The values for the real parts of the source and load admittance, G_s and G_L for a specified K can be calculated using the following formulas.

$$G_S = \frac{K[|y_f y_r| + Re(y_f y_r)]g_i^{1/2}}{2g_o} - g_i \qquad \text{Eq. (3)}$$

$$G_L = \frac{K[|y_f y_r| + Re(y_f y_r)]g_o^{1/2}}{2g_i} - g_o \qquad \text{Eq. (4)}$$

---

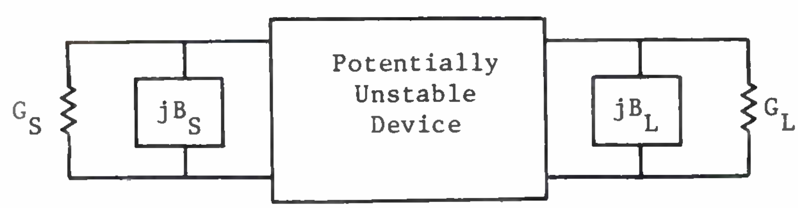
*Note: By setting $G_S = G_L = 0$, one can use Stern's rather than the Linvill stability factor to investigate the stability of the device under worst case conditions.

Solution for the imaginary parts of the source and load admittance B_s and B_L by Stern's procedure requires the solution of a cubic equation and is tedious.

The given program employs an alternative approach which uses an iterative method of successive approximations.[2] B_L is first chosen as $-b_o$ (b_o is the imaginary part of y_o). Next, using the value of G_L calculated from (4) and the first guess for B_L, the input admittance of the amplifier is calculated.

B_s is then chosen to be the negative of the imaginary part of the input admittance. Using the value of G_s calculated from (3) and the first guess for B_s, the output admittance of the amplifier is calculated. The second guess for B_L is then taken to be the negative of the output admittance. This procedure is continued until the new value of B_L is within .1 percent of its value from the previous iteration. The program is very simple to use and yields excellent agreement with the exact Stern's procedure.

If the Y parameters of a potentially unstable device are known, the designer need only enter these parameters and the desired K. The program will then calculate G_s, B_s, G_L, B_L, and the transducer gain in dB. If a printout is desired, label A′ should be pressed before execution. The execution time is a function of K becoming longer as K approaches unity.



$Y_S = G_S + jB_S$ $\qquad$ $Y_L = G_L + jB_L$

Figure 1. Potentially unstable devices with terminations.

Program Listing
RF Amp with Specified K

000	91	R/S	075	42	STO	150	10	10	225	32	X:T	300	13	13	375	43	RCL
001	76	LBL	076	25	25	151	42	STO	226	93	.	301	42	STO	376	26	26
002	11	A	077	61	GTO	152	03	03	227	00	0	302	02	02	377	91	R/S
003	72	ST*	078	23	LNX	153	43	RCL	228	00	0	303	43	RCL	378	43	RCL
004	09	09	079	76	LBL	154	11	11	229	01	1	304	14	14	379	32	32
005	99	PRT	080	33	X^2	155	42	STO	230	22	INV	305	42	STO	380	91	R/S
006	69	OP	081	53	(	156	04	04	231	77	GE	306	03	03	381	76	LBL
007	29	29	082	53	(	157	36	PGM	232	34	√X	307	43	RCL	382	24	CE
008	91	R/S	083	43	RCL	158	04	04	233	43	RCL	308	15	15	383	02	2
009	76	LBL	084	20	20	159	10	E'	234	30	30	309	42	STO	384	06	6
010	16	A'	085	65	×	160	36	PGM	235	94	+/-	310	04	04	385	69	OP
011	86	STF	086	53	(	161	04	04	236	42	STO	311	36	PGM	386	04	04
012	01	01	087	43	RCL	162	17	B'	237	26	26	312	04	04	387	43	RCL
013	91	R/S	088	19	19	163	32	X:T	238	61	GTO	313	13	C	388	20	20
014	76	LBL	089	85	+	164	94	+/-	239	35	1/X	314	94	+/-	389	69	OP
015	12	B	090	43	RCL	165	42	STO	240	76	LBL	315	44	SUM	390	06	06
016	98	ADV	091	18	18	166	24	24	241	34	√X	316	27	27	391	02	2
017	98	ADV	092	54	)	167	43	RCL	242	43	RCL	317	32	X:T	392	02	2
018	42	STO	093	65	×	168	18	18	243	30	30	318	94	+/-	393	03	3
019	20	20	094	43	RCL	169	42	STO	244	94	+/-	319	44	SUM	394	06	6
020	43	RCL	095	21	21	170	01	01	245	42	STO	320	28	28	395	69	OP
021	12	12	096	54	)	171	43	RCL	246	26	26	321	53	(	396	04	04
022	42	STO	097	55	÷	172	29	29	247	61	GTO	322	53	(	397	43	RCL
023	01	01	098	53	(	173	42	STO	248	42	STO	323	43	RCL	398	23	23
024	43	RCL	099	02	2	174	02	02	249	76	LBL	324	27	27	399	69	OP
025	13	13	100	65	×	175	53	(	250	35	1/X	325	33	X^2	400	06	06
026	42	STO	101	43	RCL	176	43	RCL	251	53	(	326	85	+	401	01	1
027	02	02	102	22	22	177	10	10	252	43	RCL	327	43	RCL	402	04	4
028	43	RCL	103	54	)	178	85	+	253	10	10	328	28	28	403	03	3
029	14	14	104	54	)	179	43	RCL	254	85	+	329	33	X^2	404	06	6
030	42	STO	105	34	√X	180	23	23	255	43	RCL	330	54	)	405	69	OP
031	03	03	106	75	-	181	54	)	256	23	23	331	35	1/X	406	04	04
032	43	RCL	107	43	RCL	182	42	STO	257	54	)	332	65	×	407	43	RCL
033	15	15	108	21	21	183	03	03	258	42	STO	333	04	4	408	24	24
034	42	STO	109	54	)	184	53	(	259	01	01	334	65	×	409	69	OP
035	04	04	110	92	RTN	185	43	RCL	260	53	(	335	43	RCL	410	06	06
036	36	PGM	111	76	LBL	186	11	11	261	43	RCL	336	23	23	411	02	2
037	04	04	112	23	LNX	187	85	+	262	11	11	337	65	×	412	02	2
038	13	C	113	43	RCL	188	43	RCL	263	85	+	338	43	RCL	413	02	2
039	42	STO	114	17	17	189	24	24	264	43	RCL	339	25	25	414	07	7
040	18	18	115	94	+/-	190	54	)	265	24	24	340	65	×	415	69	OP
041	53	(	116	42	STO	191	42	STO	266	54	)	341	53	(	416	04	04
042	24	CE	117	26	26	192	04	04	267	42	STO	342	43	RCL	417	43	RCL
043	33	X^2	118	76	LBL	193	36	PGM	268	02	02	343	14	14	418	25	25
044	85	+	119	42	STO	194	04	04	269	53	(	344	33	X^2	419	69	OP
045	32	X:T	120	43	RCL	195	18	C'	270	43	RCL	345	85	+	420	06	06
046	42	STO	121	18	18	196	43	RCL	271	16	16	346	43	RCL	421	01	1
047	29	29	122	42	STO	197	16	16	272	85	+	347	15	15	422	04	4
048	33	X^2	123	01	01	198	42	STO	273	43	RCL	348	33	X^2	423	02	2
049	54	)	124	43	RCL	199	03	03	274	25	25	349	54	)	424	07	7
050	34	√X	125	29	29	200	43	RCL	275	54	)	350	54	)	425	69	OP
051	42	STO	126	42	STO	201	17	17	276	42	STO	351	53	(	426	04	04
052	19	19	127	02	02	202	42	STO	277	03	03	352	24	CE	427	43	RCL
053	43	RCL	128	53	(	203	04	04	278	53	(	353	28	LOG	428	26	26
054	10	10	129	43	RCL	204	36	PGM	279	43	RCL	354	65	×	429	69	OP
055	42	STO	130	16	16	205	04	04	280	17	17	355	01	1	430	06	06
056	21	21	131	85	+	206	10	E'	281	85	+	356	00	0	431	02	2
057	43	RCL	132	43	RCL	207	36	PGM	282	43	RCL	357	54	)	432	02	2
058	16	16	133	25	25	208	04	04	283	26	26	358	42	STO	433	03	3
059	42	STO	134	54	)	209	17	B'	284	54	)	359	32	32	434	07	7
060	22	22	135	42	STO	210	32	X:T	285	42	STO	360	87	IFF	435	69	OP
061	71	SBR	136	03	03	211	42	STO	286	04	04	361	01	01	436	04	04
062	33	X^2	137	53	(	212	30	30	287	36	PGM	362	24	CE	437	43	RCL
063	42	STO	138	43	RCL	213	53	(	288	04	04	363	43	RCL	438	32	32
064	23	23	139	17	17	214	53	(	289	13	C	364	20	20	439	69	OP
065	43	RCL	140	85	+	215	24	CE	290	42	STO	365	91	R/S	440	06	06
066	10	10	141	43	RCL	216	85	+	291	27	27	366	43	RCL	441	91	R/S
067	42	STO	142	26	26	217	43	RCL	292	32	X:T	367	23	23	442	76	LBL
068	22	22	143	54	)	218	26	26	293	42	STO	368	91	R/S	443	15	E
069	43	RCL	144	42	STO	219	54	)	294	28	28	369	43	RCL	444	01	1
070	16	16	145	04	04	220	55	÷	295	43	RCL	370	24	24	445	00	0
071	42	STO	146	36	PGM	221	43	RCL	296	12	12	371	91	R/S	446	42	STO
072	21	21	147	04	04	222	26	26	297	42	STO	372	43	RCL	447	09	09
073	71	SBR	148	18	C'	223	54	)	298	01	01	373	25	25	448	81	RST
074	33	X^2	149	43	RCL	224	50	I×I	299	43	RCL	374	91	R/S			

Example #1

Calculate the terminations which yield maximum gain for a device with the following Y parameters and for K = 10. (C = 4.76)

$$\begin{bmatrix} y_i & y_r \\ y_f & y_o \end{bmatrix} = \begin{bmatrix} 1.5 + j3.5 & 0 - j.3 \\ 56 - j11 & .1 + j.75 \end{bmatrix}$$

all in mmhos

Procedure:

1. Read side 1 and side 2 of the "RF Amp with Specified K" program.
2. Press E to initialize the program.
3. Enter y_i, y_r, y_f, and y_o in order using the following procedure.
   Enter g_i
   Press A (Note: $g_i, b_i, \ldots, g_o, b_o$ are automatically printed as they are entered.)
   Enter b_i
   Press A .
   .
   .
   .
   Enter b_o
   Press A
4. Press A′ if a printout of the output is desired.
5. Enter K
   Press B

This is the resulting printout:

```
       1.5-03
       3.5-03
       0.  00
      -3. -04
       5.6-02
      -1.1-02
       1. -04
       7.5-04


       1.  01      K
 3.0695931-02     GS
-1.1248646-02     BS
 2.0463954-03     GL
-1.2665531-03     BL
 2.2183488 01     GT
```

If the printer is not used, the calculator will stop with K displayed. Successive pressings of R/S will display G_S, B_S, G_L, B_L, and G_T in that order. To repeat the calculation for say K = 4, simply enter 4 and press B. The Y parameters need not be reentered. For K = 4, the following printout is obtained.

```
       4.  00      K
 1.8862495-02     GS
-1.4201746-02     BS
 1.2574996-03     GL
-1.4632185-03     BL
 2.6489989 01     GT
```

Execution time for this example is approximately 45 seconds for K = 10 and approximately 70 seconds for K = 4.

Execution time will vary for different Y parameters.

If S parameters are to be used the second program which converts S to Y parameters must be executed first. This program uses the same procedure for entering the device parameters and automatically places the calculated Y parameters in the proper registers for execution of the first program. The S parameters must be entered in polar form.

RF Telemetry Links
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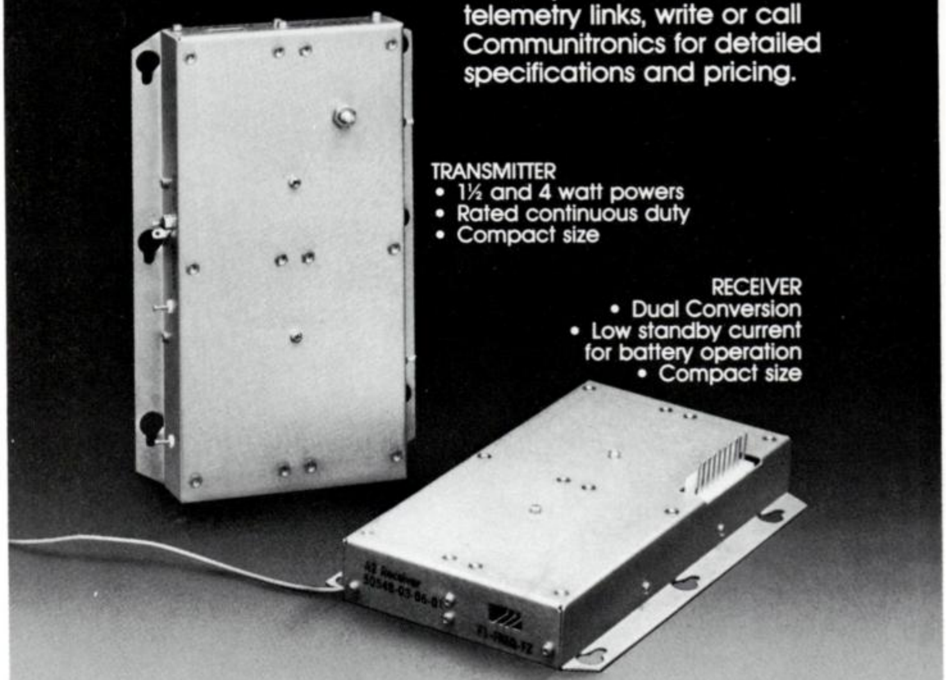

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Program Listing S→Y

000	91	R/S
001	76	LBL
002	11	A
003	72	ST*
004	09	09
005	99	PRT
006	69	OP
007	29	29
008	91	R/S
009	76	LBL
010	12	B
011	60	DEG
012	53	(
013	43	RCL
014	10	10
015	65	×
016	43	RCL
017	16	16
018	54	)
019	32	X:T
020	53	(
021	43	RCL
022	11	11
023	85	+
024	43	RCL
025	17	17
026	54	)
027	37	P/R
028	42	STO
029	28	28
030	32	X:T
031	42	STO
032	27	27
033	53	(
034	43	RCL
035	14	14
036	65	×
037	43	RCL
038	12	12
039	54	)
040	32	X:T
041	53	(
042	43	RCL
043	15	15
044	85	+
045	43	RCL
046	13	13
047	54	)
048	37	P/R
049	94	+/-
050	44	SUM
051	28	28
052	32	X:T
053	94	+/-
054	44	SUM
055	27	27
056	43	RCL
057	27	27
058	42	STO
059	35	35
060	43	RCL
061	28	28
062	42	STO
063	36	36
064	43	RCL
065	10	10
066	32	X:T
067	43	RCL
068	11	11
069	37	P/R
070	42	STO
071	11	11
072	32	X:T
073	42	STO
074	10	10
075	43	RCL
076	16	16
077	32	X:T
078	43	RCL
079	17	17
080	37	P/R
081	42	STO
082	17	17
083	32	X:T
084	42	STO
085	16	16
086	01	1
087	44	SUM
088	35	35
089	43	RCL
090	10	10
091	44	SUM
092	35	35
093	43	RCL
094	16	16
095	44	SUM
096	35	35
097	43	RCL
098	11	11
099	44	SUM
100	36	36
101	43	RCL
102	17	17
103	44	SUM
104	36	36
105	43	RCL
106	35	35
107	32	X:T
108	43	RCL
109	36	36
110	22	INV
111	37	P/R
112	42	STO
113	36	36
114	32	X:T
115	42	STO
116	35	35
117	01	1
118	42	STO
119	37	37
120	43	RCL
121	10	10
122	42	STO
123	39	39
124	94	+/-
125	44	SUM
126	37	37
127	43	RCL
128	11	11
129	42	STO
130	40	40
131	94	+/-
132	42	STO
133	38	38
134	43	RCL
135	16	16
136	44	SUM
137	37	37
138	43	RCL
139	17	17
140	44	SUM
141	38	38
142	43	RCL
143	27	27
144	94	+/-
145	44	SUM
146	37	37
147	43	RCL
148	28	28
149	94	+/-
150	44	SUM
151	38	38
152	43	RCL
153	37	37
154	32	X:T
155	43	RCL
156	38	38
157	22	INV
158	37	P/R
159	32	X:T
160	53	(
161	24	CE
162	55	÷
163	43	RCL
164	35	35
165	54	)
166	32	X:T
167	53	(
168	24	CE
169	75	-
170	43	RCL
171	36	36
172	54	)
173	37	P/R
174	42	STO
175	11	11
176	32	X:T
177	42	STO
178	10	10
179	53	(
180	02	2
181	94	+/-
182	65	×
183	43	RCL
184	12	12
185	55	÷
186	43	RCL
187	35	35
188	54	)
189	32	X:T
190	53	(
191	43	RCL
192	13	13
193	75	-
194	43	RCL
195	36	36
196	54	)
197	37	P/R
198	42	STO
199	13	13
200	32	X:T
201	42	STO
202	12	12
203	53	(
204	02	2
205	94	+/-
206	65	×
207	43	RCL
208	14	14
209	55	÷
210	43	RCL
211	35	35
212	54	)
213	32	X:T
214	53	(
215	43	RCL
216	15	15
217	75	-
218	43	RCL
219	36	36
220	54	)
221	37	P/R
222	42	STO
223	15	15
224	32	X:T
225	42	STO
226	14	14
227	01	1
228	42	STO
229	37	37
230	43	RCL
231	39	39
232	44	SUM
233	37	37
234	43	RCL
235	40	40
236	42	STO
237	38	38
238	43	RCL
239	16	16
240	94	+/-
241	44	SUM
242	37	37
243	43	RCL
244	17	17
245	94	+/-
246	44	SUM
247	38	38
248	43	RCL
249	27	27
250	94	+/-
251	44	SUM
252	37	37
253	43	RCL
254	28	28
255	94	+/-
256	44	SUM
257	38	38
258	43	RCL
259	37	37
260	32	X:T
261	43	RCL
262	38	38
263	22	INV
264	37	P/R
265	32	X:T
266	53	(
267	24	CE
268	55	÷
269	43	RCL
270	35	35
271	54	)
272	32	X:T
273	53	(
274	24	CE
275	75	-
276	43	RCL
277	36	36
278	54	)
279	37	P/R
280	42	STO
281	17	17
282	32	X:T
283	42	STO
284	16	16
285	01	1
286	00	0
287	32	X:T
288	76	LBL
289	22	INV
290	05	5
291	00	0
292	22	INV
293	64	PD*
294	09	09
295	69	OP
296	39	39
297	43	RCL
298	09	09
299	77	GE
300	22	INV
301	91	R/S
302	76	LBL
303	15	E
304	01	1
305	00	0
306	42	STO
307	09	09
308	81	RST

Example #2

Calculate the terminations which yield maximum gain for a device with the following S parameters and for K = 10. (C = 1.56)

$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} .691\angle-141.3° & .043\angle29.1° \\ 3.406\angle92.5° & .781\angle-20.7° \end{bmatrix}$$

Procedure:

1. Read side 1 and 2 of the S→Y program.
2. Enter S_{11}, S_{12}, S_{21}, S_{22} in order.
   S_{11}
   Press A
   $\angle S_{11}$
   Press A
   .
   .
   .
   $\angle S_{22}$
   Press A
3. Press B
4. Read side 1 and 2 of the "Amp with Specified K" program.
5. Press A′ if a printout of the output is desired.
6. Enter the desired K.
   Press B

Note: Neither program uses direct addressing which allows easy additions to the programs should the user desire to add to or modify the program.

```
        0.691
       -141.3
        0.043
         29.1
        3.406
         92.5
        0.781
        -20.7


          10.          K
  0.179353366         GS
 -.1441593196         BS
  .0013361587         GL
 -.0040675364         BL
  20.67207998         GT
```

References

1. Carson, R.S., *High-Frequency Amplifiers*, Wiley Interscience, 1975.
2. Kraus, H.L., *Solid State Radio Engineering*, Wiley, 1980.
3. Stern, A.P., "Stability and Power Gain of Tuned Transistor Amplifiers," Proc. IRE, Vol. 45, March 1957. □

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INFO/CARD 11

A Programmable Calculator Method for Chebyshev Filter Selection

By Marvin Kefer
Hazeltine Corp.
Greenlawn, N.Y.

This article presents a fundamental method of choosing a bandpass filter based on the solution of the Chebyshev Loss Equation. The advantages of this approach are:

- Extension of accuracy
- Selection of *any* integer number of poles
- Selection of *any* bandpass ripple
- Extension of frequency and skirt rejection ranges
- Simplification of computation and rapid results (with no manual normalization required).

Selecting a Chebyshev Filter

An ideal filter provides transparent processing of signals, in the passband, and infinite attenuation of signals outside of the passband. A Chebyshev response is often chosen to approximate the ideal filter. Matched filters, Cauer filters and Bessel function filters are superior filters for many receiver applications, but because of realization and fabrication difficulty the designer often selects a Chebyshev filter instead and accepts a small degradation of performance.

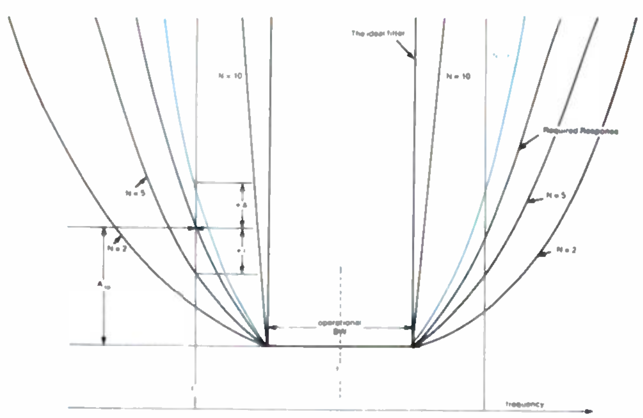

Figure 1. The Number of Poles Determines Filter Selectivity.

Two approaches, for selection of Chebyshev filter parameters, will be illustrated by two examples. In the first example an IF filter is selected by fixing the passband and solving for the required rejection. In the second example an RF filter is selected by fixing the rejection and solving for the required passband.

Example 1 — The passband of an IF filter determines the noise bandwidth of the receiver, defined as the operational bandwidth of Figure 1. To provide adjacent channel rejection the filter must provide an insertion loss (rejection) of A dB at frequency f_1. Figure 1 illustrates that a six pole filter provides the required rejection with Δ dB rejection margin. The five pole filter provides ε dB less than the required rejection.

Because fewest poles is the most practical solution from a cost/alignment viewpoint, the 10 pole solution is rejected in favor of the six pole solution.

Example 2 — The rejection of an RF filter which provides rejection to image, spurious and out of band signals, is set at A dB in Figure 2. The minimum operational bandwidth is the product of the number of channels plus 1 and the channel spacing however, a wider bandwidth may be used to:

1. Allow for filter passband variations for manufacturing tolerances, for non-infinite unloaded "Q_u" and for temperature changes.
2. Move the channels away from the 3 dB bandedge, to where the phase-frequency characteristic is more linear.

In Figure 2 all the filter responses provide the require rejection, but only filters having six or more poles have sufficient operational bandwidth.

The six pole filter solution in example 2 provides bandwidth margin while the example 1 solution provided rejection margin. In each example the solution is not unique. The designer has the option to trade some rejection margin for bandwidth margin.

Numerical Examples

Classically bandpass filter selection involves:

1. Scaling the filter requirements to an equivalent normalized low pass filter.
2. Comparing the scaled low pass response to normalized Chebyshev curve.

The author is not suggesting discarding your Chebyshev filter tables or curves (he's keeping his!), but rather an alternative procedure for selecting a filter based on an exact solution of the Chebyshev loss equation using a programmable calculator.

Our approach solves the Chebyshev loss equation directly, without manual frequency normalization. We use the number crunching ability of the calculator, instead of normalized curves, for precise solution of the loss equation.

The Chebyshev loss equation is:

$$A\,dB = 10 \log \{1 + (10^{AM/10} - 1) \cosh^2 (N \cosh^{-1} W')\} \quad \text{Eq. (1)}$$

The parameters are defined for $W' \geq 1$ in figure 2A.

In the following numerical examples the loss equation solutions will be illustrated by the calculator printout. These solutions will also serve as a program check for the calculator program, which will be explained in the following section.

The examples are intended to demonstrate procedure as well as the insight and benefit achieved by tracking the ripple bandwidth. The numerical examples add numerical values to the previous examples.

Example 1 — How many poles are needed for the following IF bandpass filter requirements?

Characteristic	**Requirement**
Passband ripple (AM)	0.5 dB
Center frequency (f0)	1780 MHz
Operational bandwidth, equals the ripple bandwidth, (BW/A)	140 MHz
Operational bandwidth bandedge rejection (A/DF)	0.5 dB
32 dB rejection bandwidth (BW)	210 MHz

The operational bandwidth was selected as the ripple bandwidth however, the operational bandwidth can be selected as the 3 dB bandwidth or any rejection bandwidth greater than or equal to the ripple bandwidth.

Equation (1) is solved for N = 2,3,4,5,6,8 the results, annotated and printed by the program, are listed in Figure 3.

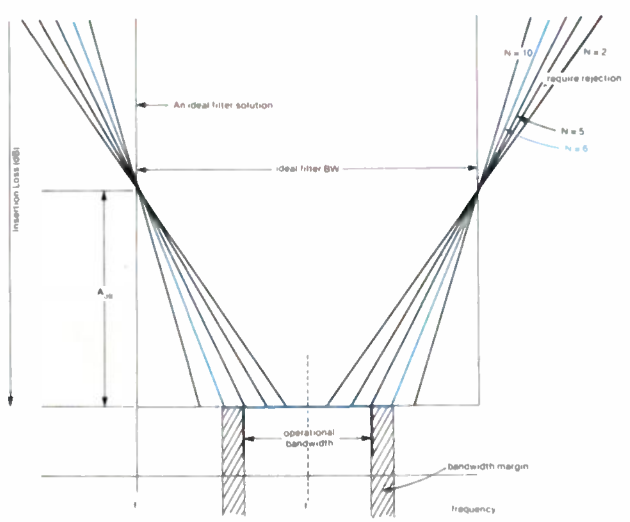

Figure 2. The Number of Poles Determines Filter Bandwidth.

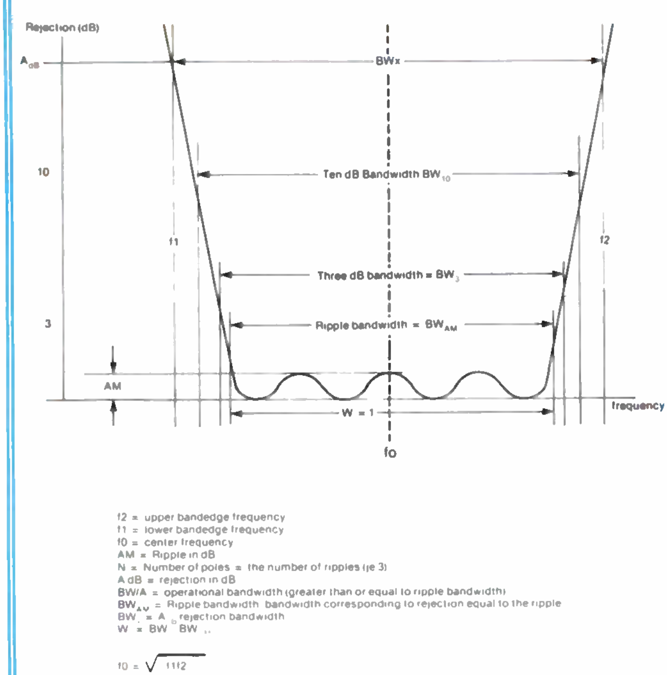

Figure 2A. Definitions of Chebyshev Loss Equation Parameters.

```
FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.5     AM
     2.      N

    210.     BW
3.970228646  IL
1678.094221  F1
1888.094221  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.5     AM
     3.      N

    210.     BW
10.36768374  IL
1678.094221  F1
1888.094221  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.5     AM
     4.      N

    210.     BW
18.34958891  IL
1678.094221  F1
1888.094221  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.5     AM
     5.      N

    210.     BW
26.65115774  IL
1678.094221  F1
1888.094221  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.5     AM
     6.      N

209.9935538  BW
     35.     IL
1678.097254  F1
1888.090808  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.5     AM
     8.      N

    210.     BW
51.71973117  IL
1678.094221  F1
1888.094221  F2
```

Figure 3. Increasing the Number of Poles Linearly Increases the 210 MHz Bandwidth Rejection.

```
FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.125   AM
     6.      N

    210.     BW
30.51243032  IL
1678.094221  F1
1888.094221  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.25    AM
     6.      N

    210.     BW
32.64945431  IL
1678.094221  F1
1888.094221  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     0.5     AM
     6.      N

    210.     BW
35.00214561  IL
1678.094221  F1
1888.094221  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140      BW/A
     1.      AM
     6       N

    210.     BW
38.38511537  IL
1678.094221  F1
1888.094221  F2

FILTER  SKIRTS
    1780.    FD
     0.5     A/DF
    140.     BW/A
     2.      AM
     6.      N

    210.     BW
41.89337226  IL
1678.094221  F1
1888.094221  F2
```

Figure 4. Filter Rejection Increases When Ripple is Increased.

```
FILTER  SKIRTS
    1780.    FD
     32.     A/DF
    210.     BW/A
     0.5     AM
     2.      N

27.70203013  BW
     0.5     IL
1766.202875  F1
1793.904905  F2

FILTER  SKIRTS
    1780.    FD
     32.     A/DF
    210.     BW/A
     0.5     AM
     3.      N

66.96350933  BW
     0.5     IL
 1746.83066  F1
1813.799169  F2

FILTER  SKIRTS
    1780.    FD
     32.     A/DF
    210.     BW/A
     0.5     AM
     4.      N

 101.384986  BW
     0.5     IL
1730.029195  F1
1831.414181  F2

FILTER  SKIRTS
    1780.    FD
     32.     A/DF
    210.     BW/A
     0.5     AM
     5.      N

 127.299598  BW
     0.5     IL
1717.487842  F1
 1844.78744  F2

FILTER  SKIRTS
    1780.    FD
     32.     A/DF
    210.     BW/A
     0.5     AM
     6.      N

146.0341448  BW
     0.5     IL
1708.479908  F1
1854.514053  F2

FILTER  SKIRTS
    1780.    FD
     32.     A/DF
    210.     BW/A
     0.5     AM
     8.      N

169.4619819  BW
     0.5     IL
1697.284537  F1
1866.746519  F2
```

Figure 5. Operational Bandwidth Dependency on the Number of Filter Poles.

Comparing the IL (insertion loss or rejection) in Figure 3 observe that from N = 3 to 8 the increase in IL is approximately 8 dB/Pole. This relationship is a specific example of a general rule which is mathematically proven in Appendix A.

Given constant ripple (AM) and constant bandwidth to ripple bandwidth ratio ($W' = \text{constant}$), rejection will increase linearly for each additional pole added to the Chebyshev loss equation expressed mathematically.

Equation (2) (From Appendix A)

$\Delta A\ dB \cong 8.68\ \phi$

Where

$\Delta A\ dB$ = increase in rejection per additional pole

$\phi = \text{Cosh}^{-1} W'$

given

$A\ dB \geqslant 10\ dB$

AM = fixed

W' = fixed

$N > 2$

Using this relationship reduces the number of trials needed to solve for the required number of poles (see Appendix C).

To gain further insight, about the Chebyshev loss equation parameters, we have varied only the ripple (AM), by doubling it each step, from AM = .125 dB to AM = 2 dB.

As the ripple doubles (In Figure 4) the rejection, at 210 MHz bandwidth increases approximately 3 dB. Another specific case of a general rule proven in Appendix B.

The increase in rejection, caused only by an increase in the passband ripple is approximately 3 dB for each doubling of the ripple. Expressed Mathematically by Equation (3) (From Appendix C).

$$\Delta A\ dB = 10 \log \frac{AM2}{AM1}$$

for AM2 = Ripple Value 2 in dB

AM1 = Ripple Value 1 in dB

BW_{AM} (BW/A) fixed and W' fixed

N fixed

then $A\ dB \geqslant 10\ dB$

Example 2 — A second solution to Example 1 is approached by fixing the rejection bandwidth and solving for the operational bandwidth. This approach corresponds to Figure 2.

Characteristic	Requirement
Passband Ripple (AM)	0.5 dB
Center frequency	1780 MHz
Rejection bandwidth (BW/A)	210 MHz
Rejection bandwidth bandedge rejection (A/DF)	32 dB
1/2 (IL) bandwidth (BW)	140 MHz

Figure 5 shows the dependency of operational bandwidths on the number of filter poles. We meet the 140 MHz requirement with a 6 MHz bandwidth margin, for N = 6.

This approach, while it does not reduce to a linear relationship of bandwidth and poles, can be useful for trading bandpass margin once an approximate solution is known.

Using the Program

The program requires A TI 59 calculator and may be used with or without a PC 100 printer.

The program is divided into four parts sharing common subroutines as follows: (see Figures 6 and 7).

Part 1 — scales the loss equation to the desired frequency and bandwidth

Part 2 — solves for rejection corresponding to the inputted bandwidth

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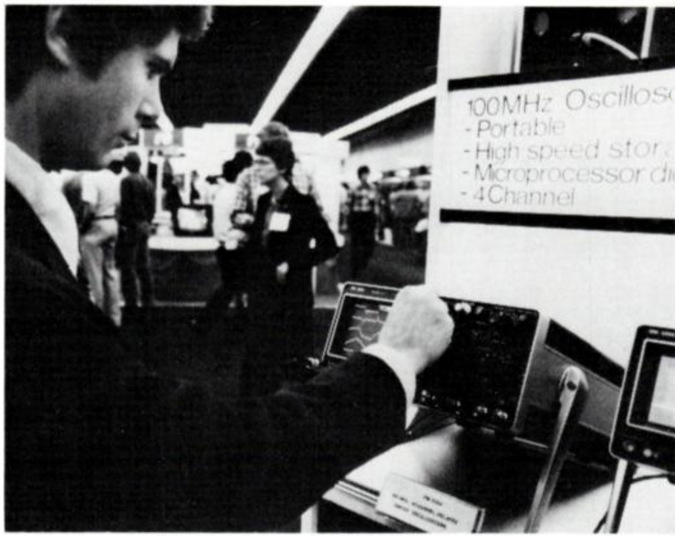

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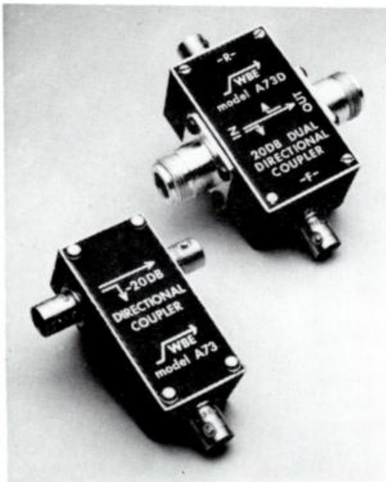

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A73-20GA				30	40			
A73-20GB				40	45			
A73-20P	1-100	single	50W cw (75 ohm limited to 10W cw)	35 dB min		.15	±.1	1.1:1 max
A73D-20P		dual		40 dB min typical		.3		
A73-20PX		single		45 dB min		.15		
A73D-20PX		dual				.3		
A73-20PA	10-200	single		35 dB min		.15		1.04:1 typical
A73D-20PA		dual		40 dB min typical		.3		
A73-20PAX		single		45 dB min		.15		
A73D-20PAX		dual				.3		

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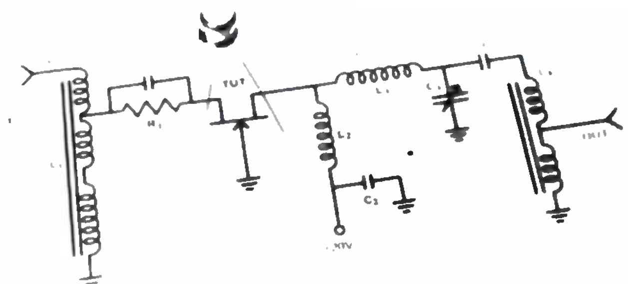

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High dynamic range RF FET now available up to 50V BVDGO for use with 24 & 32V supplies, and where higher drain voltage improves dynamic range. The CP664 (30V), CP665 (40V), and CP666 (50V) have third order intermodulation intercept <+40 dBM, and 50 Ohm VSWR >1.5 to 1 over 0.5 to 50 MHz range.

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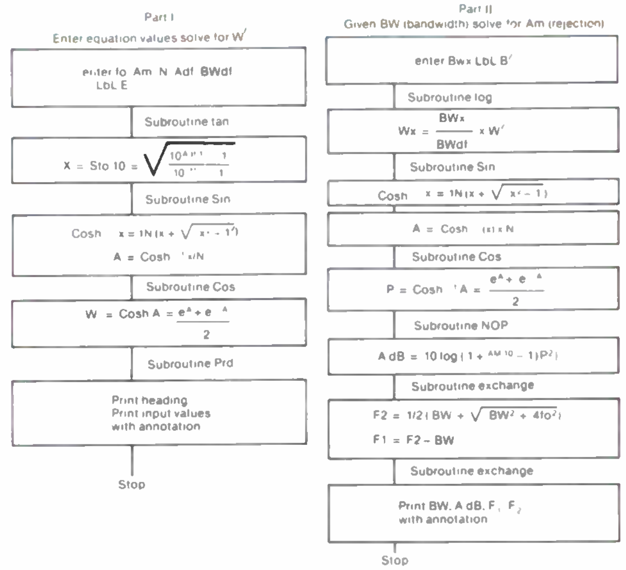

Figure 6. Flow Charts of Parts I and II of Calculator Program.

Part 3 — solves for bandwidth corresponding to the inputted rejection

Part 4 — solves for the bandwidth corresponding to the inputted bandedge frequency

The program is described in Figure 8, with memory locations indicated and is listed in Figure 9.

Comments:

- Care must be exercised when entering a rejection bandwidth (BWx), if the bandwidth entered is smaller than the ripple bandwidth the answer will be incorrect. The $W' \geq 1$ requirement will have been violated and the program has no failsafe solution.
- To use the program for low pass filter selection set $f_0 = 0$ and enter frequency as if it was bandwidth. Band reject and highpass filter responses require program modifications or working with inverse entries.

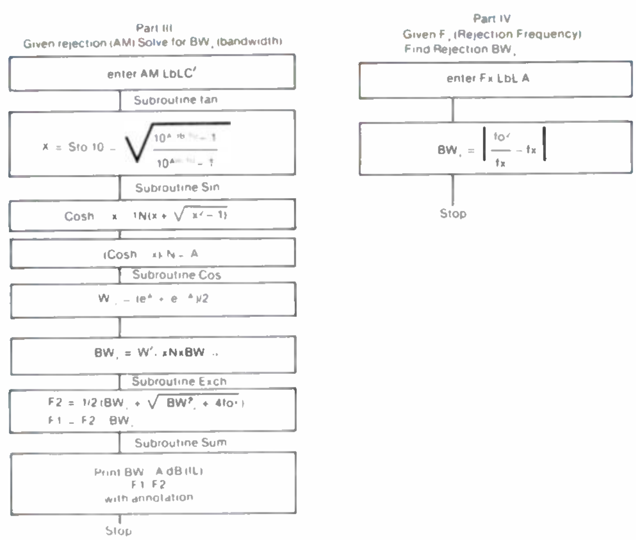

Figure 7. Flow Charts of Parts 3 and 4 of Calculator Program.

TI PROGRAM RECORD

PROGRAM DESCRIPTION

Normalizes the loss equation for center frequency, ripple, number of poles and a particular rejection bandwidth. Solves the normalized equation for bandwidth given rejection or rejection given bandwidth, for low pass and band pass filters. For low pass set F0 = 0 and Bw = frequency.

USER INSTRUCTIONS

STEP	PROCEDURE	ENTER	PRESS			DISPLAY
1	Enter center frequency	FO	A			FO
2	Enter ripple in dB	AM	B			AM
3	Enter number of poles	N	C			N
4	Enter desired rejection	A/DF	D			A/DF
5	Enter corresponding bandwidth Printer lists annotated inputs and calculates w′ (normalizes to input)	BW/A	E			N
6	Enter fx, for calculation of associated rejection bandwidth BWx	fx	A			BWx
7	Enter BWx, for calculation of associated rejection Printer lists annotate outputs BW, IL (rejection) F1 (lower band edge) F2 (upper band edge)	BWx	B			F2
8	Enter IL for calculation of associated BW. Printer lists annotated outputs of BW, IL, F1, F2	IL	C			IL

USER DEFINED KEYS		DATA REGISTERS (INV ■)			
A	FO (center freq)	0	Constant	10	Cosh ϕ
B	AM (dB ripple)	1	Work register	11	ϕ
C	N (# poles)	2	A/DF (input)	12	
D	A/DF (BW reject.)	3	AM (input)	13	F2 (upper freq.)
E	BW/A (reject. BW)	4	F0 (input)	14	F1 (lower freq.)
A′	fx (reject. freq.)	5	BW/A (input)	15	F input
B′	BWx (reject. BW)	6	N (poles input)	16	A/DF
C′	ILx (BW reject.)	7	W′ ripple BW ratio	17	
D′		8	BWx (input or cal.)	18	
E′		9	ILx (A/DF or Cal.)	19	

LABELS (Op 08)

INV _ lnx _ CE _ CLR _ x:t _ x² _
√x _ 1/x _ STO _ RCL _ SUM _ yˣ _
EE _ (_) _ ÷ _ GTO _ × _
SBR _ − _ RST _ + _ R/S _ . _
+/− _ = _ CLR _ INV _

FLAGS	0	1	2	3	4	5	6	7	8	9

Texas Instruments Incorporated 1014966-1

Figure 8. A description of the Chebyshev Loss Equation.

149	76	LBL	202	01	01	254	06	06	306	04	04	358	06	06	410	42	STO
150	30	TAN	203	02	2	255	01	1	307	43	RCL	359	02	2	411	11	11
151	43	RCL	204	07	7	256	03	3	308	06	06	360	01	1	412	71	SBR
152	02	02	205	03	3	257	06	6	309	69	OP	361	00	0	413	39	COS
153	55	÷	206	07	7	258	03	3	310	06	06	362	03	3	414	71	SBR
154	01	1	207	01	1	259	01	1	311	98	ADV	363	00	0	415	68	NOP
155	00	0	208	07	7	260	06	6	312	92	RTN	364	00	0	416	71	SBR
156	95	=	209	03	3	261	02	2	313	76	LBL	365	00	0	417	48	EXC
157	22	INV	210	05	5	262	01	1	314	44	SUM	366	00	0	418	71	SBR
158	28	LOG	211	00	0	263	69	OP	315	25	CLR	367	69	OP	419	44	SUM
159	75	-	212	00	0	264	04	04	316	98	ADV	368	04	04	420	92	RTN
160	01	1	213	69	OP	265	43	RCL	317	01	1	369	43	RCL	421	76	LBL
161	95	=	214	02	02	266	02	02	318	04	4	370	13	13	422	18	C'
162	42	STO	215	00	0	267	69	OP	319	04	4	371	69	OP	423	42	STO
163	12	12	216	00	0	268	06	06	320	03	3	372	06	06	424	09	09
164	43	RCL	217	03	3	269	01	1	321	00	0	373	98	ADV	425	48	EXC
165	03	03	218	06	6	270	04	4	322	00	0	374	92	RTN	426	02	02
166	55	÷	219	02	2	271	04	4	323	00	0	375	76	LBL	427	48	EXC
167	01	1	220	06	6	272	03	3	324	00	0	376	15	E	428	16	16
168	00	0	221	02	2	273	06	6	325	69	OP	377	42	STO	429	71	SBR
169	95	=	222	04	4	274	03	3	326	04	04	378	05	05	430	30	TAN
170	22	INV	223	03	3	275	01	1	327	43	RCL	379	71	SBR	431	71	SBR
171	28	LOG	224	05	5	276	03	3	328	08	08	380	30	TAN	432	38	SIN
172	75	-	225	69	OP	277	69	OP	329	69	OP	381	43	RCL	433	55	÷
173	01	1	226	03	03	278	04	04	330	06	06	382	10	10	434	43	RCL
174	95	=	227	03	3	279	43	RCL	331	02	2	383	71	SBR	435	06	06
175	35	1/X	228	07	7	280	05	05	332	04	4	384	38	SIN	436	95	=
176	65	×	229	03	3	281	69	OP	333	02	2	385	55	÷	437	42	STO
177	43	RCL	230	06	6	282	06	06	334	07	7	386	43	RCL	438	11	11
178	12	12	231	00	0	283	01	1	335	00	0	387	06	06	439	71	SBR
179	95	=	232	00	0	284	03	3	336	00	0	388	95	=	440	39	COS
180	34	√X	233	00	0	285	03	3	337	00	0	389	42	STO	441	65	×
181	42	STO	234	00	0	286	00	0	338	00	0	390	11	11	442	43	RCL
182	10	10	235	00	0	287	00	0	339	69	OP	391	71	SBR	443	05	05
183	92	RTN	236	00	0	288	00	0	340	04	04	392	39	COS	444	55	÷
184	76	LBL	237	69	OP	289	00	0	341	43	RCL	393	42	STO	445	43	RCL
185	49	PRD	238	04	04	290	00	0	342	09	09	394	07	07	446	07	07
186	25	CLR	239	69	OP	291	69	OP	343	69	OP	395	71	SBR	447	95	=
187	69	OP	240	05	05	292	04	04	344	06	06	396	49	PRD	448	42	STO
188	00	00	241	02	2	293	43	RCL	345	02	2	397	92	RTN	449	08	08
189	69	OP	242	01	1	294	03	03	346	01	1	398	76	LBL	450	71	SBR
190	01	01	243	03	3	295	69	OP	347	00	0	399	17	B'	451	48	EXC
191	00	0	244	02	2	296	06	06	348	02	2	400	42	STO	452	71	SBR
192	00	0	245	00	0	297	03	3	349	00	0	401	08	08	453	44	SUM
193	00	0	246	00	0	298	01	1	350	00	0	402	71	SBR	454	25	CLR
194	00	0	247	00	0	299	00	0	351	00	0	403	28	LOG	455	43	RCL
195	00	0	248	00	0	300	00	0	352	00	0	404	71	SBR	456	16	16
196	00	0	249	69	OP	301	00	0	353	69	OP	405	38	SIN	457	42	STO
197	02	2	250	04	04	302	00	0	354	04	04	406	65	×	458	09	09
198	01	1	251	43	RCL	303	00	0	355	43	RCL	407	43	RCL	459	48	EXC
199	02	2	252	04	04	304	00	0	356	14	14	408	06	06	460	02	02
200	04	4	253	69	OP	305	69	OP	357	69	OP	409	95	=	461	92	RTN
201	69	OP															

Figure 9. A Listing of the Loss Equation Program.

- The TI 59 calculator does not have in inverse Cosh function

The relationship:

$\text{Cosh}^{-1} x = 1\text{n}(x + \sqrt{x^2 - 1})$ for $x \geq 1$ was used in the program

Conclusion: A procedure to select the design parameters for a filter using a programmable calculator has been described. The method provides accurate solutions to the Chebyshev loss equation to provide design specifications for:

- center frequency
- operational bandwidth
- bandwidth ripple
- rejection for out of band signals

The procedure also provides an accurate tool for trading rejection margin, ripple and passband margin.

Appendix A

Given constant rejection, ripple bandwidth, ripple and large rejection (≥10 dB). Rejection will increase linearly, for a given rejection bandwidth, for each pole added to the Chebyshev equation.

$$\text{A dB} = 10 \log\{1 + (1^{\text{AM}/10} - 1)\,\text{Cosh}^2(\text{N Cosh}^{-1}\text{W}')\} \quad \text{Eq. (1)}$$

For $\phi = \text{Cosh}^{-1}\text{W}'$ for AM and W'

$$\text{Then A dB} \cong 10 \log(1 + \text{KCosh}^2\text{N}\phi)$$

$$\text{For KCosh}^2\text{N}\phi >> 1 \text{ or A dB} \gtrapprox 10 \text{ dB}$$

$$\text{A dB} \cong 10 \log \text{KCosh}^2\text{N}\phi$$

$$\cong 10 \log \text{K} + 20 \times .434\ 1\text{n Cosh N}\phi$$

$$\text{A dB} \cong 10 \log \text{K} + 8.68\left(\text{N}\phi + \frac{1}{\text{N}\phi}\right)$$

Δ A dB = increase of attenuation for an additional pole

$\Delta \text{A dB} \cong \text{A dB}|(\text{N}+1) - \text{A dB}|_{\text{N}}$

$$= 8.68\left[(\text{N}+1)\phi + \frac{1}{(\text{N}+1)\phi} + \text{N}\phi\frac{-1}{\text{N}\phi}\right]$$

$$\cong 8.68\left[\phi + \left(\frac{1}{\text{N}+1} - \frac{1}{\text{N}}\right)\frac{1}{\phi}\right]$$

$$= 8.68\left[\phi \frac{-1}{\text{N}(\text{N}+1)\phi}\right]$$

For N>2

Then A dB ≅ 8.68 ϕ Equation (2)

Appendix B

Changing only the ripple (AM) (the number of poles (N) and the ratio of bandwidth to ripple bandwidth (W^1) remaining constant). The attenuation will increase as the log of the ratio of the ripple for large rejection (≥ 10 dB) and small ripple (≅3/4 dB).

$$\text{A dB} = 10 \log\{1 + (10^{\text{AM}/10} - 1)\,\text{Cosh}^2(\text{N Cosh}^{-1}\text{W}^1)\} \quad \text{Eq. (1)}$$

for N & W^1 constant

$$\text{A dB} = 10 \log\{1 + (10^{\text{AM}/10} - 1)\,\text{K}\}$$

rearranging terms

$$10^{\text{A dB}/10} - 1 = (10^{\text{AM}/10} - 1)\,\text{K}$$

for $10^{\text{A dB}/10} >> 1$ large rejection (A dB ≥ 10)

$10^{\text{A dB}/10} \cong 1 + \text{AM}/10$ small ripple (≈ 3/4 dB)

then $10^{\text{A dB}/10} \cong (\text{AM}/10)\,\text{K}$

$$\text{or A dB} \cong 10 \log \frac{\text{K}}{10} + 10 \log \text{AM}$$

$$\text{So } \Delta \text{ A dB} = \text{A dB}|_{\text{AM2}} - \text{A dB}|_{\text{AM1}}$$

$$\cong 10 \log \text{AM2} - 10 \log \text{AM1}$$

$$\text{Then } \Delta \text{ A dB} \cong 10 \log \frac{\text{AM2}}{\text{AM1}} \quad \text{Equation (3)}$$

Appendix C

Using the linear relationship, Equation (2), to solve numerical Example 1.

$$\Delta \text{ A dB} \cong 8.68\,\text{Cosh}^{-1}\text{W}' = \text{Constant}$$

where

Δ A dB = increase in attenuation per added pole

and A dB ≥ 10 dB

Instead of trial and error two samples, N = 3 (IL = 10.4) and N = 4 (IL = 18.4), could have been used to solve for the needed number of poles.

For example:

From figure 3

for

N	BW	IL	Δ A dB
3	210	10.37	
4	210	18.35	7.98 ≈ 8

for 32 dB rejection how many additional poles are required?

18.35 + 8x = 32 where x = 1.7 additional poles so that 4 + 1.7 = 5.7 poles are needed. Again six poles are selected, and there is rejection margin. □

AN IMPROVED LUMPED-CONSTANT HYBRID

Here is a lumped-constant "L" matching hybrid which performs better than the conventional lumped-constant Wilkinson Hybrid and is readily constructed at 175 MHz and below. It also readily transforms impedance at the same time as splitting or combining power.

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Lynchburg, Virginia 24502

Designers of medium power RF amplifiers use hybrids to achieve output powers greater than about 75 watts. The power density in a single transistor is so large that two transistors or two modules must be combined to achieve this power . Power combining is accomplished by using a hybrid. Hybrid junctions may be used to combine or to split power. This makes the device reciprocal. An example of using the hybrid in a mobile radio power amplifier for splitting and combining power is shown in Figure 1.

The hybrid must present the proper impedance at all ports. It must also provide isolation between the side ports. Maintaining spectral purity in a mobile radio power amplifier under the conditions of variations in supply voltage, RF drive power, operating temperature and load impedance is difficult enough with a single output transistor. If two transistors are connected in parallel at the output to achieve higher power, the interaction of dynamic impedances makes the job even more difficult. Thus we need the hybrid which offers high isolation between the side ports.

At frequencies below 175 MHz the hybrid is typically formed using lumped-constants. The familiar Wilkinson[1] Hybrid is formed with quarter-wave transmission lines synthesized with two pi sections. A typical example of this lumped-constant Wilkinson Hybrid used in an RF amplifier is given by Ken Dufour.[2] It is suggested that the performance of this amplifier can be improved by replacing the Wilkinson with an "L" network hybrid. This hybrid is based on the familiar "L" matching network. Both of these hybrids are shown in Figure 2 for comparison. The Wilkinson uses quarter-wave or 90° phase-shift lines. The improvement uses "L" matching networks with a phase-shift dependent on any impedance transformation.

Wilkinson Lumped Constant

The Wilkinson Hybrid is composed of two quarter-wave (90° phase shift) lines connected together at the summation port. The artificial transmission line can be synthesized using a pi section as shown in Figure 3a. We could just as easily have used the "T" section.[3] The combined pi sections are also shown forming the Wilkinson Hybrid. Power applied at the sum port 1 will be split equally (3.01 dB) between side port 2 and side port 3. The impedance looking into any port will be 50 ohms, provided that the other ports are properly terminated.

The beauty of the Wilkinson Hybrid is that side port 2 is isolated from port 3. This isolation is achieved by using the balance resistor R1. The transmission path from side port 2 to side port 3 is shown re-drawn as two paths in Figure 3b. Power from a signal generator injected at side port 2 traveling through the two pi sections undergoes a 6 dB loss and a 180° phase shift before exiting at the termination placed at side port 3. The alternate path from side port 2 to port 3 is through the balance resistor R1. This path also introduces a 6 dB loss but zero phase shift. The re-combined signal at port 3 is thus cancelled out as two equal amplitude, 180° out-of-phase signals. This isolation is needed to prevent interaction when combining the power from two transistors or from two modules. A value of 20 dB of isolation is considered adequate. Any reflected power at port 2 will then be reduced by a factor of 100 before appearing at port 3.

The theoretical performance of a 50 ohm Wilkinson Hybrid is shown in Figure 4. The results are plotted on a normalized frequency scale so that the designer can merely multiply by the design center frequency to predict the response at any particular frequency. The input return loss at the sum port is better than 15 dB (SWR $\leq$ 1.4), insertion loss less 3.15 dB, and side port isolation greater than 20 dB over a 24 percent bandwidth. (Percent bandwidth is determined by dividing the measured bandwidth by the center frequency.) This in itself is good but the L network lumped-constant hybrid is better.

"L" Network Hybrid

The L network hybrid is shown dissected into two L matching networks in Figure 5a. We can accomplish this dissection because any components placed between the side ports such as C5 and R2 do not affect equal power splitting. Because the voltage and phase at side port 2 is the same as that at side port 3, no current will flow through any component connected between them. The hybrid is composed of two "L" matching networks connected together at the sum port. Because these two matching networks are in parallel at the sum port, they must individually have a 100 ohm input impedance (R4). The output impedance (R3) is made equal to the side port impedance of 50 ohms. The design of an "L" network[4] indicates that:

$$L = \frac{R3\sqrt{(R4/R3) - 1}}{2\pi f}$$

$$C = \frac{\sqrt{(R4/R3) - 1}}{2\pi f\ R4}$$

where R4 > R3
L = inductor in henries
C = capacitance in farads
f = frequency in hertz
π = pi, 3.14159

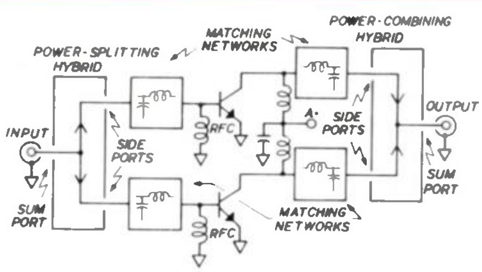

Figure 1. The reciprocity of the hybrid is demonstrated by its use as a power splitter at the input and a combiner at the output.

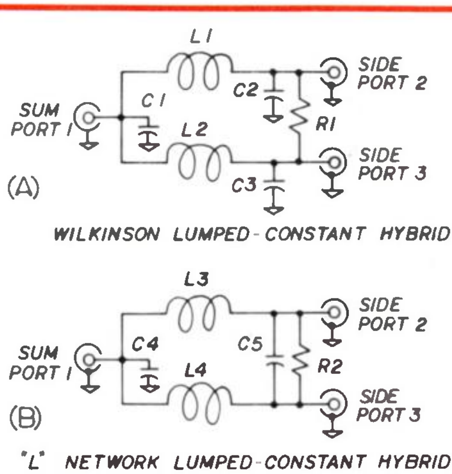

Figure 2. The schematic for the Wilkinson hybrid (a) differs from the "L" Network hybrid (b) only in the capacitors across or in shunt with the side ports.

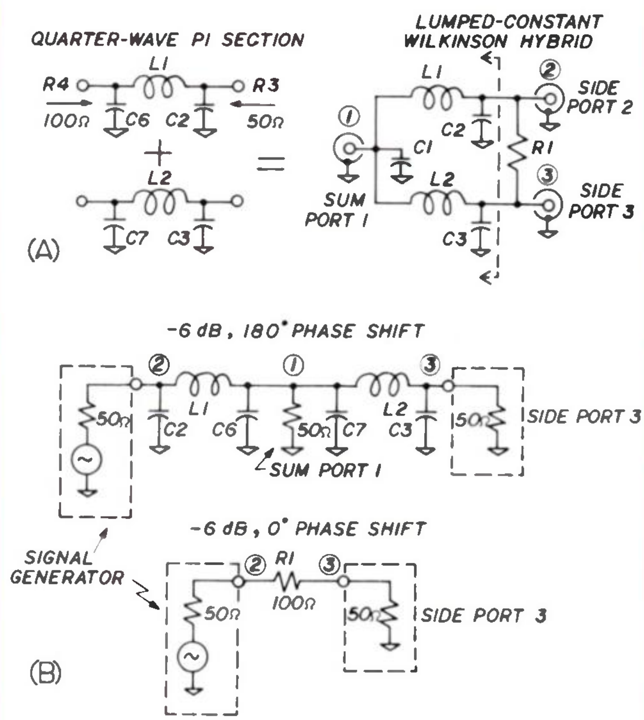

Figure 3. The Wilkinson hybrid uses two 90° artificial line sections (a). The isolation is achieved by two equal loss, 180° out-of-phase paths (b).

Wilkinson Hybrid		Low-pass "L" Network Hybrid		High-pass "L" Network Hybrid	
Component	Formula	Component	Formula	Component	Formula
C1	$1/(\pi f\sqrt{R4R3})$	C4	$[\sqrt{(R4/R3)-1}\,]/\pi f\, R4$	C10, C11	$1/[2\pi f R3\sqrt{(R4/R3)-1}\,]$
C2, C	$1/(2\pi f\sqrt{R4R3})$	C5	$1/[4\pi f R3\sqrt{(R4/R3)-1}\,]$	L5	$R4/[4\pi f\sqrt{(R4/R3-1}\,]$
L1, L2	$(\sqrt{R4R3})/2\pi f$	L3, L4	$[R3\sqrt{(R4/R3)-1}\,]/2\pi f$	L6	$[R3\sqrt{(R4/R3)-1}\,]/\pi f$
R1	2R3	R2	2R3	R5	2R3

where $\pi = 3.14159$
f = frequency in Hertz
R3 = impedance of side port termination
R4 = twice the sum port termination

Table 1. Calculation for the components of a lumped-constant hybrid.

Let's consider the design of a 52 MHz hybrid for example. We substitute R4 = 100Ω, R3 = 50Ω, and f = 52 MHz into the above equations.

$$L3, L4 = \frac{50\sqrt{(100/50)-1}}{(2 \times 3.14)(52 \times 10^6)} = 153 \text{ nH}$$

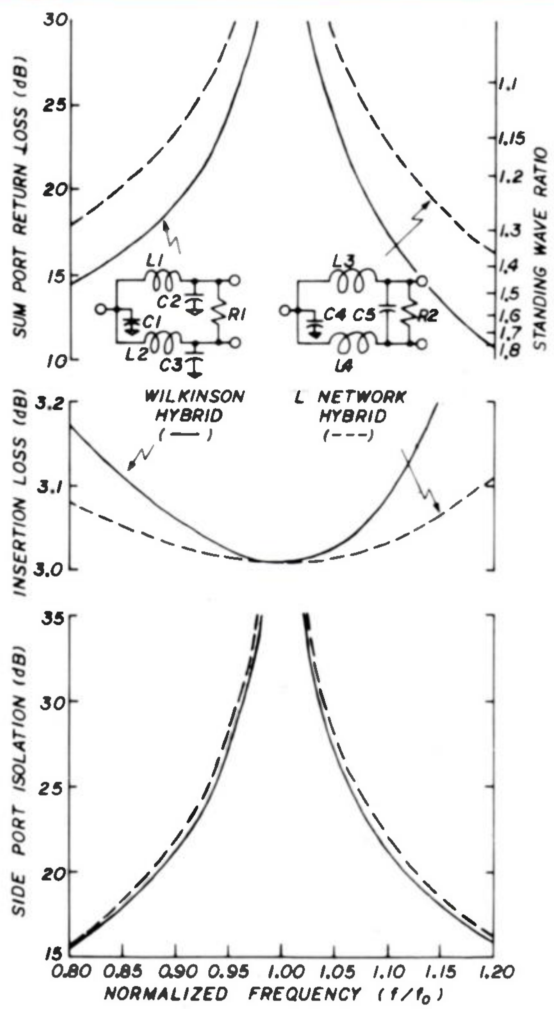

Figure 4. The theoretical response of the Wilkinson (solid line) is less than the response of the "L" Network hybrid (dotted line).

$$C8, C9 = \frac{\sqrt{(100/50)-1}}{(2 \times 3.14)(52 \times 10^6)(100)} = 30.6 \text{ pF}$$

When the two L networks are combined into a hybrid we find that the shunt capacitor C4 at the sum port is equal to the sum of C8 and C9. This forms the "L" network power splitter/combiner. The performance is predicted in Figure 4. The sum port return loss is improved while the insertion loss is reduced as compared to the Wilkinson Hybrid. The isolation for either hybrid is practically equal. The return loss at each side port is not shown because it is always quite good (≥ 20 dB over a 40 percent bandwidth).

The phase shift of an L matching network at the center frequency is defined by the equation:

$$\text{Cos B} = \sqrt{R3/R4}$$

where B = phase shift in degrees

The L matching network needed for this lumped constant network must match the 100 ohm impedance connected to the sum port down to a 50 ohm impedance at either side port, R4 = 100 ohms and R3 = 50 ohms.

$$\text{Cos B} = \sqrt{50/100} = 0.707$$
$$B = 45°$$

The phase shift for a wave travelling between the sum port and either side port is 45°. For a Wilkinson Hybrid this value is always 90°.

The isolation between side ports is only 6 dB at the center frequency if we only use the two L networks to form the hybrid. This means that one-fourth of any mismatched energy at one side port appears at the other side port. The balance network (Figure 2b) consisting of resistor R2 and capacitor C5 increases the isolation to greater than 20 dB over a 20 percent bandwidth. These balance components do not even enter into the power splitting portion of the circuit because they are placed between equal amplitude/phase points. As long as the power divides equally and the phase shift in each leg is equal, there will be no current flow in any device placed between side port 2 and side port 3.

The balance network used to increase isolation consists of two paths. One path is through the "bridged-tee" network as shown in Figure 5b. The 6 dB loss, 180° phase shift in this path must again be cancelled out by the

Frequency (MHz)	C1 (pF)	C2 (pF)	L1, 2 (nH)	Number Turns	ID (Inch)	Wire Size AWG
29	110	55	274	10	0.30	14
52	62	30	153	6	0.30	16
146	22	10	54	4	0.22	16

Table 2. Experimental Models.

6 dB loss, 0° phase shift resistive path through R2. The value of the balance capacitor C5 is set so that its capacitive reactance exactly equals the inductive reactance of L3 plus L4. The value of the balance resistor R2 is still twice the side port impedance.

$$C5 = \frac{1}{2(2\pi f \times R3 \times \sqrt{[R4/R3 - 1]})}$$

$$R2 = 2R3$$

The calculations for constructing either a Wilkinson or an "L" network hybrid are summarized in Table 1.

Experimental Models

Several "L" network lumped-constant models were constructed according to the values shown in Table 2 for 29, 52, and 146 MHz. The measured results are shown in Figure 6, along with a photograph of one of the models. The experimental results agree quite well with the predicted performance. The capacitors were formed by tacking several values together to achieve the calculated values.

The input shunt capacitor should be as close to the calculated values as possible for a good starting point. The value should be selected using a bridge or a digital capacitance meter. The inductors are then adjusted by injecting a signal at the sum port and minimizing the reflected signal into the sum port at the center frequency with the side ports properly terminated. The inductors are optimized by using "diddle" sticks with a core of ferrite on one end and a brass slug on the other end. The presence of brass lowers the inductance of the coils. The isolation is next maximized by injecting a signal into one side port and minimizing the detected energy at the other side port. The balance capacitor C5 is adjusted for maximum attenuation. The coils must be sufficiently spaced from each other so that there is minimal coupling. The largest manageable wire size is used to make self-supporting coils. Figure 7 shows the worse-case theoretical response when all of the components are varied by ± 10 percent. Even if *all* of the parts

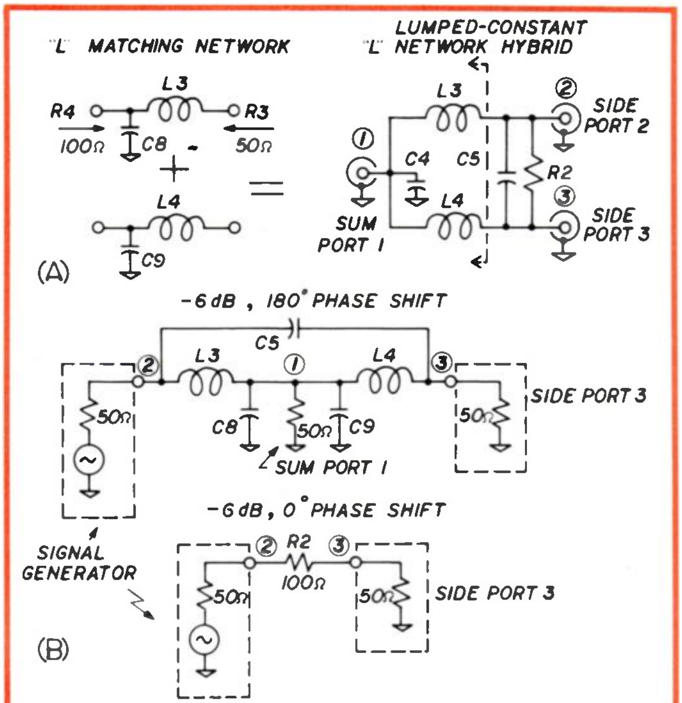

Figure 5. The "L" Network hybrid uses two "L" matching networks (a). The loss and phase delay in the bridged-T path is cancelled by the balance resistor path (b).

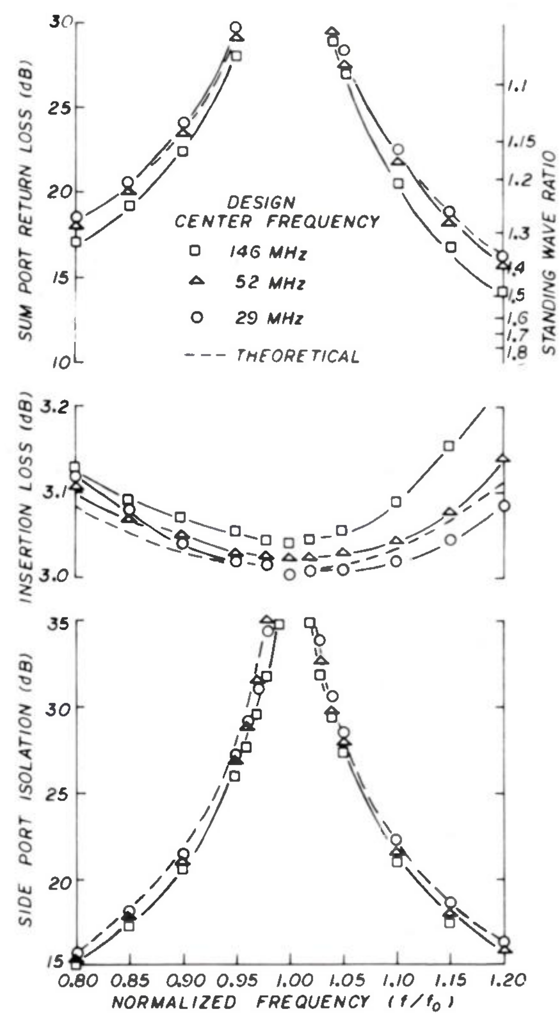

Figure 6. The experimental results for a 10 meter, 6 meter, and 2 meter "L" Network hybrid.

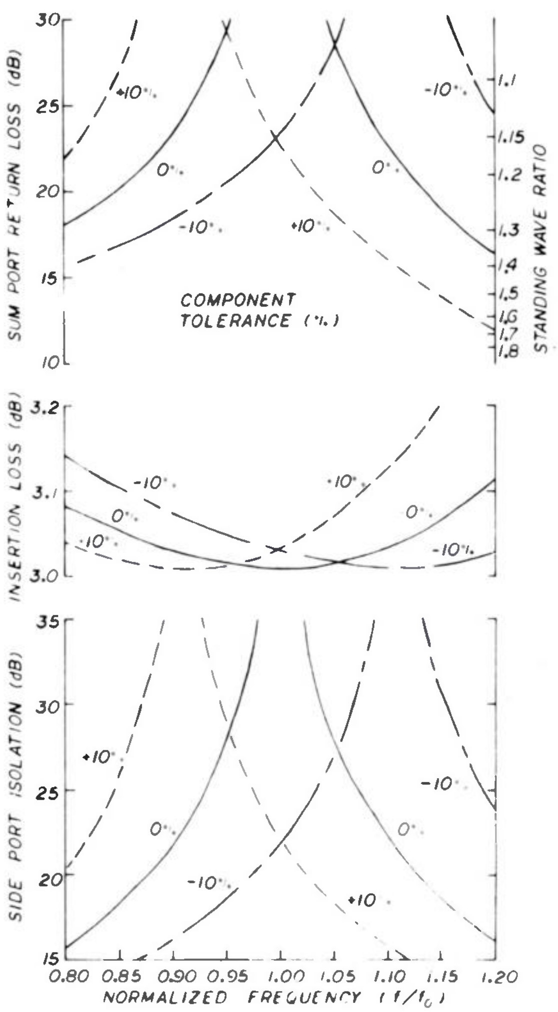

Figure 7. Theoretical response of the "L" Network hybrid when all of the components are varied by ± 10 percent.

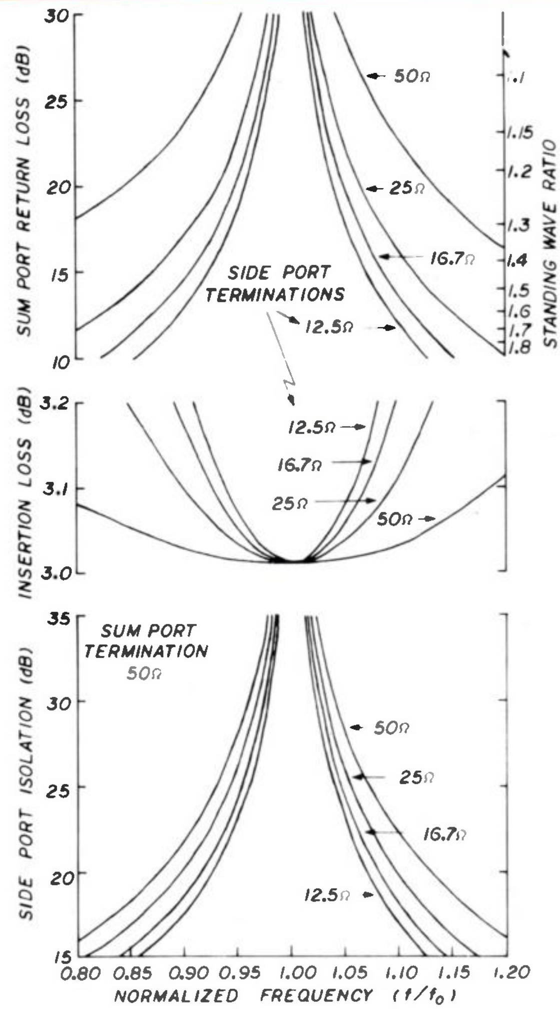

Figure 8. Theoretical response of the "L" network hybrid when used as a hybrid and as an impedance transformer for side port impedances of 50, 25, 16.7, and 12.5 ohms.

vary by 10 percent the sum port return loss is greater than 15 dB, the insertion loss less than 3.15 dB, and the side port isolation is better than 15 dB over a 24 percent bandwidth.

Impedance Transformation

Until now we have only considered hybrids which presented 50 ohms impedance at each port. In practice it is usually desirable to transform impedances at the same time we are combining or splitting power. A good example of this is the hybrid used to combine the power from the collector circuits of two transistors. Instead of using a multi-section network to raise the extremely low transistor output impedance up to 50 ohms we might design a hybrid with a side port of 25 or maybe 12.5 ohms input impedance. A penalty is payed in bandwidth anytime the Q or the impedance transformation ratio is increased, but that penalty must be paid somewhere. As an example, consider two cases for transforming a collector resistance of 2 ohms up to an antenna impedance of 50 ohms where the design is limited to only two "L" matching sections plus the hybrid. The Q of an "L" network is given as

$$Q = \sqrt{Z_{Ratio} - 1}$$

where Z_{Ratio} is the ratio of the input to output resistive component of impedance.

For the first case two "L" sections are placed before the 50 ohm hybrid. The impedance transformation ratio is

$$Z_{ratio} = \left(\frac{50\text{ ohms}}{2\text{ ohms}}\right)^{\frac{1}{\text{Number of sections}}}$$

$$Z_{ratio} = \frac{50^{1/2}}{2} = 5$$

$$Q = \sqrt{5 - 1} = 2$$

Thus, the Q of two for each of the first two sections would do most of the band-limiting. The "L" network within the 50 ohm hybrid has a Q of 1.

For the second case let us still be limited to two "L" matching sections but consider the "L" section within the hybrid as a variable. Thus, effectively there are three sections of matching. The final impedance will be the 100 ohms presented by one-half the hybrid at the sum port.

$$Z_{ratio} = \frac{100^{1/3}}{2} = 3.68$$

$$Q = \sqrt{3.68 - 1} = 1.64$$

Now each matching section is band-limiting equally with individual Q's of 1.64. This yields a wider bandwidth amplifier for the same number of parts. The theoretical results when using the hybrids to transform impedances are shown in Figure 8 using the same equations as before where R4 is always equal to 100 ohms and R3 is equal to the side port impedance. It appears that a ratio of 50 ohms to 25 ohms is the limit for wideband amplifiers. Just as the experimental results matched quite well with the predicted results for the equal impedance case, similar results for the case of impedance transformations can be expected.

The phase shift through each arm of the L network hybrid is no longer 45° as in the 50 ohm hybrid but equal to the arcosine of the square root of the impedance ratio. For the case of a 4:1 impedance transformation, where the sum port is 50 ohms and the side port is 12.5 ohms R4 = 100, R3 = 12.5.

$$\text{Cos B} = \sqrt{\text{R3/R4}} = \sqrt{12.5/100}$$

$$\text{B} = 82.8°$$

High-pass Version

When forming the "L" matching network in the lumped-constant hybrid the low-pass version was naturally chosen. Alternatively, the high-pass version as shown in Figure 9 could have been chosen. The individual "L" sections are combined with the junction shunt inductance L5 equal to the electrical paralleling of L7 and L8. The balance network used for isolation consists of L6 and R5. As before, consider the two possible paths between side ports. The path through the matching networks is made to equal 6 dB loss and 180° phase shift by resonating the inductance L6 with the series sum of the capacitors C10 and C11. The alternate path through R3 is again made equal to twice the side port impedance to achieve the necessary 6 dB loss, zero degree phase shift.

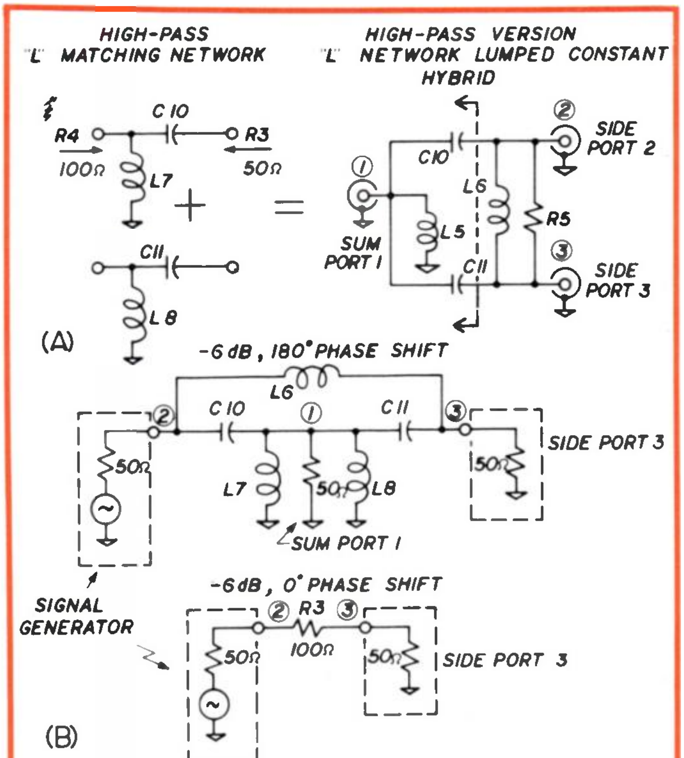

Figure 9. The High-Pass "L" Network hybrid uses two high-pass "L" matching networks (a). The loss and phase delay in the bridged-tee path are cancelled by the balance resistor path (b).

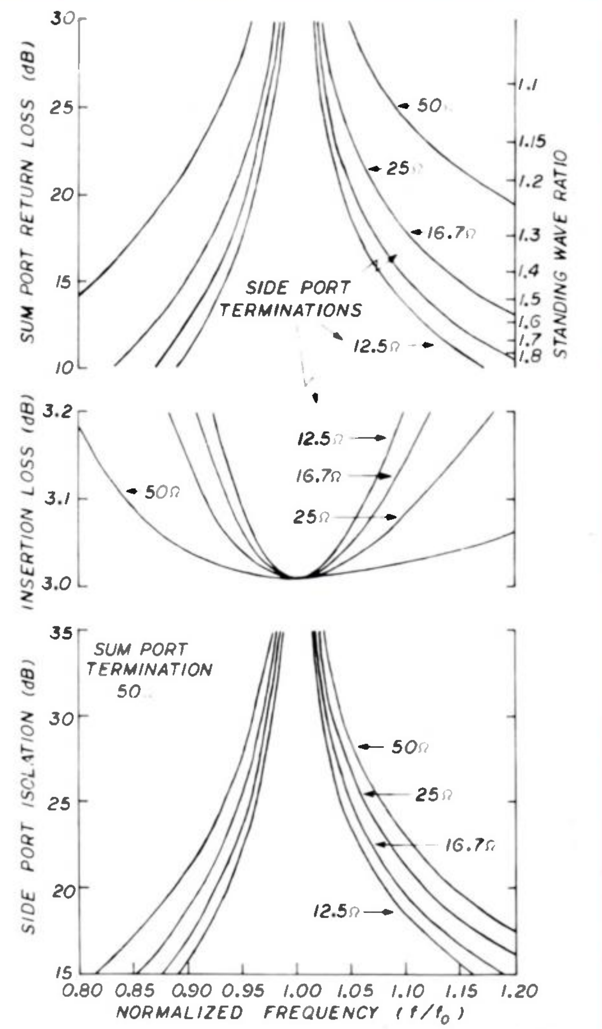

Figure 10. Theoretical response of the High-Pass "L" Network Hybrid for side port impedances of 50, 25, 16.7 and 12.5 ohms.

Normally, the low-pass version of the lumped-constant combiner is preferred because it does offer additional harmonic attenuation. The high-pass version does offer the advantage of having built-in coupling capacitors. It also offers space advantages if the balance inductor L6 is wound around R5.

The theoretical results for the high-pass version are shown in Figure 10. They differ from the low-pass curves in that they are skewed toward the higher frequencies.

References

1. E.J. Wilkinson, "An N-Way Hybrid Power Divider," *IRE Trans. on Microwave Theory and Techniques,* Vol. MTT-8, January 1960, pp. 116-118.
2. Ken Dufour, "60 Watt VHF Amplifier Using Splitting/Combining Techniques," *r.f design,* September/October 1981, pp. 18-23. Motorola Engineering Bulletin EB-93.
3. F.E. Terman, *Electronic and Radio Engineering,* McGraw-Hill, 1955, pp. 112-115.
4. Ernie Franke, WA2EWT, "Appreciating the L Matching Network," *Ham Radio,* September, 1980, pp. 26-30.

Hi-heat Pumping Thermoelectric Modules

A new series of Cambion *"Cool and Heat" TE devices* for low cost cooling and close temperature control uses are realized because the new TE modules are supplied *without metalized pads* with no sacrifice in performance. Thermal performance is at least as high as conventional TE devices having metal plates on the module surfaces.

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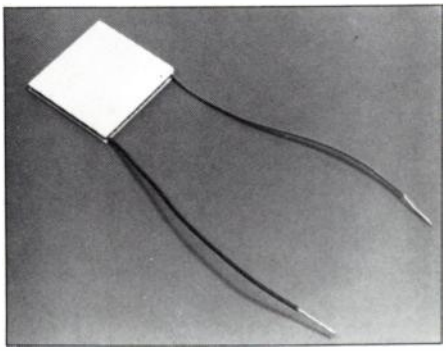
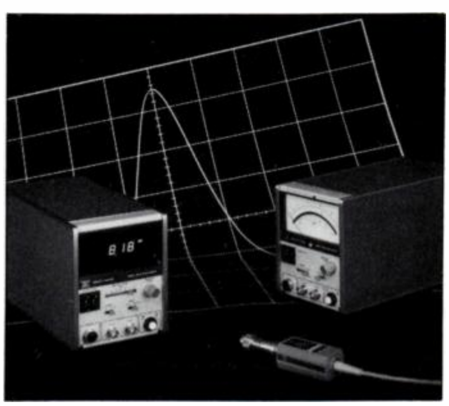


modules provide a range of 20 watts through 45 watts heat-pumping capacity at current ratings from 7 through 14 amperes.

Contact Cambion Customer Service at 445 Concord Avenue, Cambridge, MA 02238. (617) 491-5400. Please circle INFO/CARD #140.

Microwave Power Meters Measure Peak Power Directly

Two new microwave power meters — one digital, the other analog — have been announced by Hewlett-Packard Company. The meters will measure



peak power output of pulsed microwave systems from 1 to 100 mW, and from 100 MHz to 18 GHz.

The HP 8900C has an analog meter for easy peaking of pulsed power while the HP 8900D uses a digital display with range annunciators for unambiguous readings.

Both measure peak power directly instead of the old method of using an average power meter and a duty cycle computation. This new technique reduces the chance for both instrument and computation error.

In conjunction with the HP 84811A peak power sensor, the new method covers most RF and microwave applications of pulsed power systems.

Each meter has two operating modes:

- *Direct mode* senses and displays peak power with no adjustments.
- *Compare mode* drives an external oscilloscope to provide a video reproduction of the pulse envelope along with an adjustable reference line. The operator then positions the line to correspond to various points of a pulse — overshoot, for example — and reads out the power level of the reference line.

The HP 84811A peak power sensor responds accurately to peak power from 1 mW to 100 mW, pulse widths from 1 microsecond to CW and repetition rates from 100 Hz to 100 kHz. Each sensor also is furnished with a correction factor table vs. frequency that can be set into the meter to correct the reading for sensor frequency response.

SWR is less than 1.5 from 10 MHz to 12 GHz and less than 2.0 from 12 GHz to 18 GHz. Maximum power overload is 250 mW.

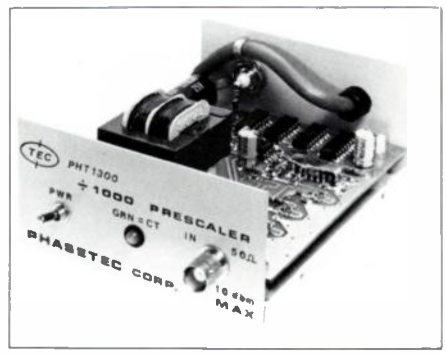
As an additional convenience for system measurements that sample power with a directional coupler, the HP 8900D digital meter features 0-60 dB power offset in 10 dB steps. By setting in the coupling value, the meter thus displays up to 100 kilowatts directly.

Overall measuring uncertainty depends on frequency but is approximately ± 1.5 dB at 12 GHz, exclusive of mismatch effects. Recorder output allows leveling of peak power sources, and the video output permits digitizing of the waveform with external equipment. INFO/CARD #139.

1400 MHz Prescaler

Phasetec Corporation of West Peabody, Mass., a growing electronics manufacturer, has introduced its new model PHT 1300 Divide by 1000 Prescaler, which can extend the range of your existing counter to 1400 MHz, thus providing a high performance, low cost alternative for microwave frequency measurement.

An ideal addition to any lab bench or test station, this easy-to-use AC operated Prescaler divides input frequency by 1000. The PHT 1300 Prescaler features low input VSWR, high sensitivity, and wide dynamic range over the 10-1400 MHz spectrum. Internal threshold circuitry provides output



gating plus a clear visual indication of minimum input operating level.

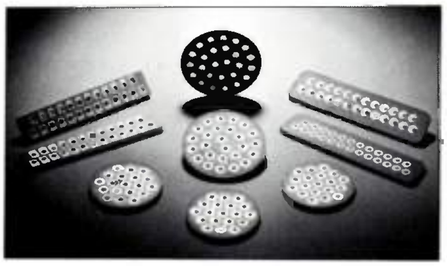
The PHT 1300 Prescaler is now available from stock for $239. For additional information contact Jeff Schiffer, president, Phasetec Corporation, P.O. Box 2086, West Peabody, MA 01960. (617) 535-4833. INFO/CARD #138.

Planar Capacitor Array

Viclan, Inc. introduces a new Planar Capacitor Array (PCA) that provides a combination of multi-line EMI filtering with mass termination at the connector interface. One Viclan PCA may be used to replace as many as 128 single-line filtering devices — allowing significant reductions in weight, space, and cost.

The Viclan PCA consists of a monolithic ceramic package with a customer-specified number and pattern of capacitors to fit connector pin arrangements.

Viclan PCA's are available in circular, subminiature D, dual-in-line, and



other custom shapes. Planars may be used in the design of "C," "L," "T," and "Pi" style filters. Capacitance values up to 10,000 pF. Voltage ratings up to 600 VDC or 230 VAC.

Engineering samples of your custom requirements available in four weeks. Pricing per line as low as 25 cents, depending on quantity and complexity of design.

For more information contact your local Viclan representative, or call Maurice Westbury, Viclan, Inc., (714) 292-1411. Write 7373A Engineer Road, San Diego, CA 92111. INFO/CARD #137.

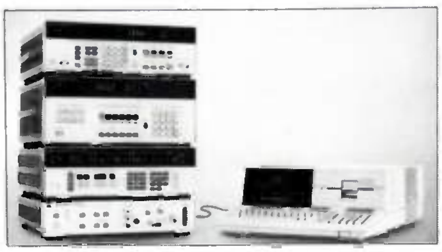
Automatic/Manual Transceiver Test Set

A new transceiver test set that provides fully automatic or manual operation by combining general-purpose instruments into one test system has been announced by Hewlett-Packard Company.

Performing both automatic and manual in-channel tests of AM, FM and ϕM communications receivers and transmitters, the new test set meets transceiver testing needs at the lowest prices ever set for an HP product with this capability.

Modular in structure, the HP 8953A combines the measurement power of the HP 8901A modulation analyzer, HP 8903A audio analyzer and HP 8656A synthesized signal generator with the HP 8954A transceiver interface, all necessary cables and accessories, and a choice of instrument controllers. As a combination of standard HP instruments, it can be assembled from on-hand components and purchased items or obtained as a complete test set.

Both the HP-85F instrument controller and the recently introduced HP 9826A desktop computer are available as controller options. Together these instruments and controllers per-



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form simple tests such as frequency and distortion as well as complex measurements such as receiver-usable sensitivity and audio flatness in just a few seconds.

For those requiring additional measurement capabilities such as out-of-channel or SSB testing, the test set's HP 8954A transceiver interface provides additional instrument connections.

A second RF signal generator such as an HP 8662A synthesized signal generator may be added for out-of-channel receiver testing and a second RF monitor such as the HP 8568A spectrum analyzer or the HP 3586C selective-level meter is available for out-of-channel transmitter and SSB testing.

The HP 8953A transceiver test set's three key instruments — the HP 8901A, HP 8903A and HP 8656A — are general-purpose, fully programmable instruments with HB-IB (IEEE-488) interface for remote control. Both manually and under remote control, these instruments perform complex measurements quickly and accurately.

Front-panel annunciators continuously indicate the filter, detector, measurement keys, modulation and display formats in use. The technician can monitor the state of the instruments at all times, even when they are being controlled remotely.

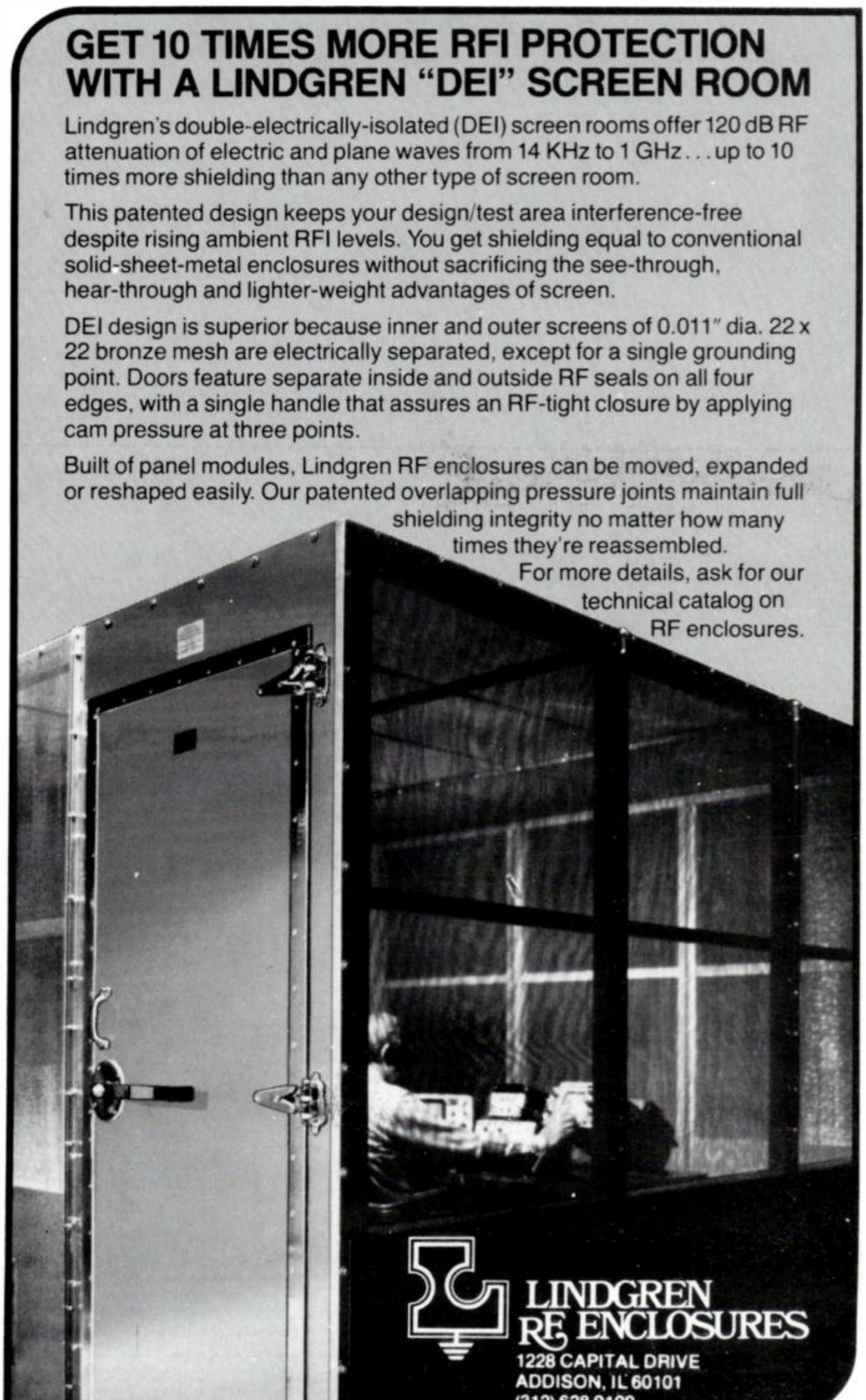
The HP 8656A signal generator has calibrated AM, FM and simultaneous internal and external modulation, which can generate simultaineous squelch and test tones.

For tests such as transmitter distortion, the HP 8903A audio analyzer's selectable high-pass and low-pass input filters remove unwanted signals such as noise or squelch tones. Circle INFO/CARD #136.

Flange Attenuators

New 3mm/SMA flange attenuators from EMC Technology, Inc., Cherry Hill, N.J. function as coax connectors for MIC packages while providing precise attenuation values. Series 7400 flange attenuators eliminate connector pairs to permit smaller, lower cost and more reliable packages.

These and other advantages of the new attenuators make them ideal for applications including channel amplitude balancing in receiver circuits for RF or IF packages. They are available in 1 dB increments up to 20 dB for highest flexibility, and are only 3/4" long to accommodate a package without creating design problems.



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A re-designed version of the company's model 4400 attenuator, series 7400 models incorporate EMC's unique interchangeable transition pins which provide precise matches between the circuit and the transmission line. The pins are available in four different configurations to meet specific requirements. Configurations include slotted, tab, standard SMA pins and machinable blank pins.

The attenuators operate from DC to 18 GHz. To assure highest electrical and mechanical performance, attenuators are constructed with passivated stainless steel bodies and gold-plated beryllium copper pins. Operating temperature ranges from −55°C to 125°C. Power rating is 1 watt at 25°C. Full calibration is also offered as an option.

Cost for the new series 7400 four-hole flange termination is $24.64 each in 100 lot quantities; delivery is from stock to six weeks depending upon attenuation value.

Contact EMC Technology, Inc., 1971 Old Cuthbert Road, Cherry Hill, NJ 08034. (609) 429-7800. INFO/CARD #135.

Solid State Modulator

EDCO model PS103048 Modulator is designed to control 40 Watt mini-TWTs in parallel providing a 400 Volt pulse with operating frequency of CW to 2 MHz. Rise and fall delays are less than 40 nanoseconds, with a RF rise time of 5 ns. The modulator input accepts a 10 kHz square wave, 40 V peak-to-peak, and operates on 15 Watts

of standby power. Trigger pulse is 10 Volts into a 50 ohm impedance. The unit is ruggedized and designed to operate under military airborne environments. Volume is less than 34 cubic inches.

Contact Edco Engineering Corporation, 3255 Scott Blvd., Bldg. 6A, Santa Clara, CA 95050. (408) 241-7226. Telex: 171-598. INFO/CARD #133.

High Power RF Gated Amplifier Generates Pulses Up to 3 KW

A high power rf gated amplifier for use with the manufacturer's main frame gating modulator and an external cw source in studies of ultrasonics and nuclear resonance is being introduced by Matec, Inc. of Warwick, R.I.

The Matec model 515A-HP R.F. Gated Amplifier features an rms pulse power output rated at 3 KW from 0.5 to 8 MHz and 1.75 KW from 8 to 20 MHz. Plugged into the firm's model 5100 Gating Modulator (Main Frame) and an external cw source, it produces sequences of coherent, high power rf pulses with pulse width and separation variable continuously or in steps of one or 10 cycles of the cw frequency.

The Matec model 515A-HP R.F. Gated Amplifier accepts cw input levels from less that 10 mV to more than 100 mV with no change in output. The unit also accepts externally generated rf pulses: i.e. a multiple rf pulse program for NMR studies, or a single rf pulse from a low level pulsed oscillator in ultrasonic applications.

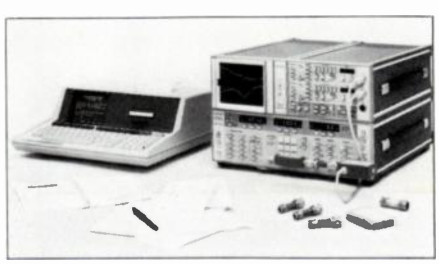
The Matec model 515A-HP R.F. Gated Amplifier is priced at $4595 (domestic); model 5100 Gating Modulator (Main Frame) at $4450. Literature is available on request. For more information contact Matec, Inc., Bruce B. Chick, 60 Montebello Road, Warwick, RI 02886, (401) 739-9030. INFO/CARD #118.

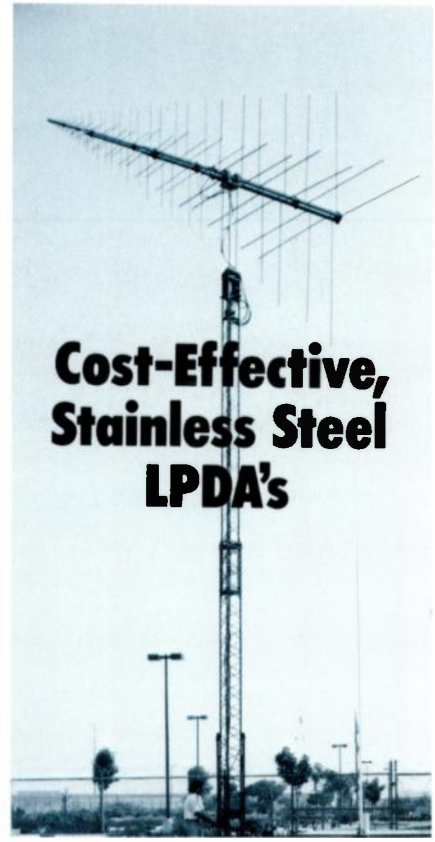
New Testing Services

Retlif, Inc. announces the addition of two new testing services. The laboratory is now equipped to perform 200 V/M susceptibility testing over the frequency range of 10 kHz to 40 GHz. Retlif, Inc., a leader in the field of FCC compliance testing, has recently completed an expansion program in regards to its FCC test division. This program included the addition of another shielded enclosure along with automated CISPR test instrumentation in order to provide rapid testing in line with FCC Subpart J type testing. For additional information please contact our sales department at (516) 751-4600 or write Retlif, Inc., Flowerfield Bldg. 25, St. James, NY 11780. Please circle INFO/CARD #131.

Automated Scaler Network Analyzer

The Wiltron 5609/75 Automated Scaler Network Analyzer measures return





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Measurements begin by inserting the preprogrammed cartridge supplied in the model 85 Controller. The operator is then guided step-by-step through the straightforward test procedure. Return loss and transmission characteristics are displayed, permitting the operator to adjust the test device before a hard-copy graph or tabulation of test data is made.

Model 5609/75 is priced at $27,490 with 90 day delivery. Contact Walt Baxter, Wiltron Company, 805 East Middlefield Road, P.O. Box 7290, Mountain View, CA 94042-7290. (415) 969-6500. TWX: 910-379-6578. Please circle INFO/CARD #132.

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CATV Hybrid Return Amplifiers

CATV hybrid return amplifiers in the 5 to 200 MHz frequency range offer high dynamic range to eliminate critical fine tuning problems. The CA 4400 series, from TRW Semiconductors, are designed for mid-split and high-split systems. They achieve a substantial improvement in dynamic range (typically 5 dB), compared with earlier models, due to the incorporation of a new-generation transistor.

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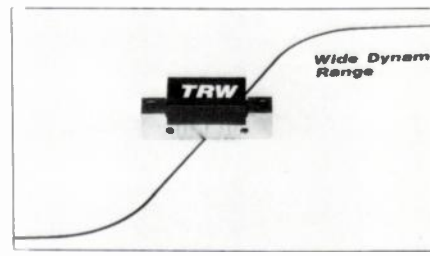

and 22 dB of gain respectively. CA 4400 series amplifiers are available immediately from TRW Semiconductors' worldwide sales and distribution network.

For further information, contact TRW Semiconductors, 14520 Aviation Blvd., Lawndale, CA 90260. (213) 679-4561. TWX: 910-325-6206. Telex: 67-7148. INFO/CARD #130.

Log-periodic Antenna

Watkins-Johnson Company has introduced the WJ-8344 Log-Periodic Antenna featuring extremely broadband operation. This antenna covers the full 1.0 to 18.0 GHz range and is designed for use in applications where size, weight and cost are factors.

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Applications for the WJ-8344 include surveillance, RFI, EMC, TEMPEST, point-to-point communication and detection.

Complete details, including price and delivery, are available from Watkins-Johnson Company, Antenna Applications Engineering, 2525 North First Street, San Jose, CA 95131. (408) 262-1411, ext. 247. Please circle INFO/CARD #127.

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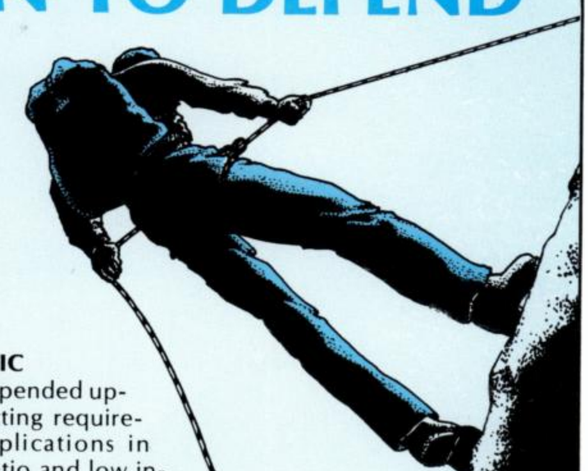

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Hold Button disconnects mouthpiece mike so caller can't hear you. Gives you privacy without covering mouthpiece with your hand.

Dial Button. Press to retrieve and automatically dial stored numbers.

Electronic Tone Ringer. Pleasant sounding.

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INFO/CARD 21

CODE-A-PHONE®

features a recessed-gate structure and is available in both chip and package form. Typical performance of the ALF 3000 chip is 0.9 dB noise figure, 13.5 dB associated gain at 4.0 GHz, and 1.5 dB noise figure, 11.0 dB associated gain at 8.0 GHz. This low noise figure and high gain combination, coupled with a device geometry that facilitates bonding, makes the ALF 3000 series an obvious choice for all oscillator and LNA applications.
The price is $62.50 each in quantities of 100 and availability is from stock. For further information contact Alpha Industries, Inc., 20 Sylvan Road, Woburn, MA 01801, (617) 935-5150, TWX: 710-393-1236, Telex: 949436. INFO/CARD #128.

Synthesized Signal Generator

Eaton Corporation's Electronic Instrumentation Division has introduced the AILTECH model 384, an advanced direct synthesized signal generator capable of providing signals to 4 GHz. The instrument features fast switching — 25 μs across its entire range of 1 MHz to 4 GHz.

The 384 is the latest in the AILTECH 360/380 family of signal generators. A microprocessor controlled frequency synthesized instrument, it features keyboard entry and IEEE Standard 488 interface BUS.

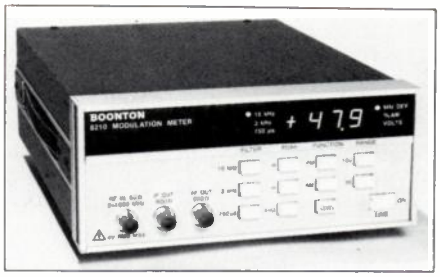
The unit has a wide range of laboratory and field applications for frequency agile systems including production ATE, EW target simulators, secure communications systems and avionics ground support consoles.

Contact Eaton Corporation, Electronic Instrumentation Division, 2070 Fifth Avenue, Ronkonkoma, NY 11779. (516) 588-3600. INFO/CARD #129.

Modulation Meter

Boonton's new 8210 Modulation Meter is the first low-cost instrument that automatically calibrates both AM and FM channels each time power is applied. As a result of this self-calibration, AM accuracy is one percent of reading, from 10 percent to 90 percent AM, for carrier frequencies from 2 MHz to 520 MHz and modulation frequencies of 50 Hz to 5 kHz. FM accuracy is also one percent of reading for deviations up to 150 kHz, carrier frequencies of 2 MHz to 1.5 GHz, and modulation rates of 50 Hz to 5 kHz. Modulation frequencies from 30 Hz to 15 kHz are accepted on both AM and FM channels with degraded accuracy.

Tuning and leveling are fully automatic with the 8210. The instrument acquires the largest signal present at

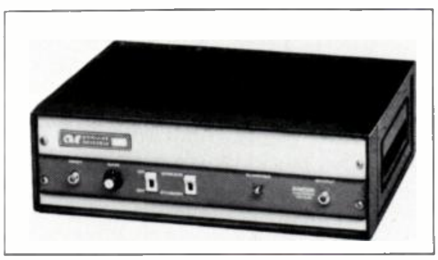


the input connector and adjusts its local oscillator and measurement channel gain to indicate the fully calibrated AM or FM on a 3-1/2 digit display. Selectable 3 kHz and 15 kHz loss-pass filters are standard, as is a 750 μs de-emphasis switch. Residual FM of less than 150 Hz at 1.5 GHz, decreasing linearly to a floor of less than 5 Hz with 3 kHz bandwidth, and residual AM of less than 0.15 percent AM allow the measurement of unmodulated carriers. FM rejection is less than one percent AM at 100 kHz peak deviation; AM rejection is less than 100 Hz deviation at 50 percent AM.

The meter is priced at $1,795 with 90 day delivery. Contact Scott Elkins, VP Marketing, Boonton Electronics Corp., P.O. Box 122, Parsippany, NJ 07054. (201) 887-5110. INFO/CARD #126.

Compact 150 Watt Amplifier

Considerably smaller and lighter than other RF amplifiers in its class, the new model 150LA from Amplifier Research is rated at 150 Watts minimum power, 1-110 MHz (60 watts minimum to 150 MHz). It will continue

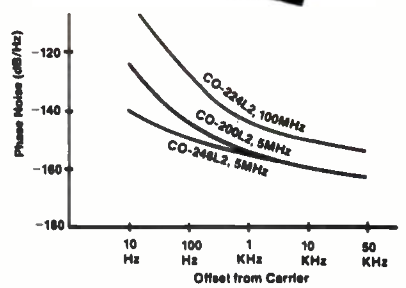


to operate regardless of the magnitude or phase of source and load impedances, without oscillation, shutdown, or damage.

Model 150LA combines solid-state low power stages with a vacuum tube final amplifier, providing instantaneous bandwidth (full power at any frequency in its operating spectrum, without tuning or bandswitching) and exceptional pulse and blanking characteristics. A typical RF envelope exhibits pulse droop of less than one percent.

At only 23 kg (50 lb), the new unit measures 50 cm wide, 16 cm high, and 45 cm deep (19.8" x 6.3" x 18.0") and is priced at $4,900. Contact Amplifier Research, 160 School House Road, Souderton, PA 18964. (215) 723-8181. TWX: 510-661-6094. INFO/CARD #125.

Low Noise Crystal Oscillators



HF MODELS		
Series:	CO-200L2	CO-246L2
Frequency:	4-25 MHz	5 MHz (opt. to 100 MHz)
Aging:	$1x10^{-9}$/day $3x10^{-7}$/year	$1x10^{-10}$/day $3x10^{-8}$/year
Phase noise:	at 5 MHz	
@100 Hz:	−145 dB/Hz	−150 dB/Hz
@ 1 KHz:	−155 dB/Hz	−155 dB/Hz
@ 50 KHz:	−163 dB/Hz	−163 dB/Hz

VHF MODELS		
Series:	CO-224L2	CO-220L2
Frequency:	25-300 MHz	25-125 MHz
Aging:	$1x10^{-8}$/day $2x10^{-6}$/year	$1x10^{-9}$/day $3x10^{-7}$/year
Phase noise:	at 100 MHz	
@100 Hz:	−130 dB/Hz	−120 dB/Hz
@ 1 KHz:	−145 dB/Hz	−130 dB/Hz
@ 50 KHz:	−155 dB/Hz	−140 dB/Hz

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INFO/CARD 20

Advertiser Index

SECTION MANAGER R.F. PRODUCTS

Lead a creative group of R.F. Design Engineers through the development of a new line of innovative Broadcast Products. This position requires a strong background in the design of AM and FM transmitting equipment along with proven skills in project management. B.S.E.E. or M.S.E.E. with at least 3-5 years of R.F. Design experience and some supervisory skills are desired. This position is an excellent opportunity for a successful project engineer to move into engineering management.

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Broadcast Electronics is known for high quality broadcast audio, automated programming, and FM transmitting equipment.

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For additional information call or send resume in complete confidence to: Vice President, Engineering.

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Programmable Broadband Solid State Noise Generators

Replacing conventional sweep generators, Micronectics' model PNG 5100 through 5110 series of Noise Generators feature full noise source capability from 10 Hz to 1 GHz in a compact instrument or rack-mounted panel, only seven inches high. Output is greater than + 10 dBm (10 mW) across the given frequency range.

IEEE-488 Interface Bus Remote Programming includes on and off standby, external pulse and attenuation control, as well as local and remote capability.

Standard frequency bands are: 10 Hz-20 kHz, 10 Hz-100 kHz, 10 Hz-500 kHz, 100 Hz-3 MHz, 100 Hz-10 MHz, 100 Hz-25 MHz, 100 Hz-100 MHz, 1 MHz-200 MHz, 1 MHz-500 MHz, 10 MHz-1 GHz.

Optional frequency bands are available up to 40 GHz.

For complete details on these new Programmable Solid State Noise Sources, contact Micronetics, Inc., 36 Oak Street, Norwood, NJ 07648. (201) 767-1320. TWX: 710-991-9603. INFO/CARD #123.

Manual A-B Data Switch For Coax, Twinax Introduced

A manual A-B Coax or Twinax switch with several housing and connector options, for switching between two data, video or digital sources, has been announced by MarLee Switch Company.

The single pole, double throw switch, model 981, covers the DC to 100 MHz frequency range. Isolation is better than − 100 dB, insertion loss less than 1/2 dB. Characteristic impedances are 50 or 75 ohms, with other impedances and terminated versions available. The unit can allow, for instance, manual switching of two CRT terminals to one controller, or, switching of one display between two controllers.

The model 981-0001 desk model Coax A-B manual switch is $140.00 in single piece quantity. Delivery is typically 60 days after receipt of order. Normal quantity discounts are offered.

Marlee Switch Company, 933-D North Central Ave., Upland, CA 91786, designs and manufactures switches, matrices, controls and displays for RF, video, data and audio switching and interconnecting. INFO/CARD #124.

New Literature

Inductive Components Catalog

CAMBION, a leading manufacturer and supplier of components to the electronics industry, has just published its *Inductive Components Catalog 111.* This new and completely updated 34-page catalog features comprehensive listings of Cambion's Variable Coils (shielded and unshielded), RF Chokes (including military types), Coil Forms (both PC board and panel mount), Micro-inductors (see section below), Testing Fixtures, and selected Capacitors.

The new Catalog 111 contains full documentation for all Inductive Components with accompanying dimensioned line drawings and tabulated specifications. Also included in this compact, information-filled 34-page engineering design literature is a "tear-out" MIL Standard/Cambion RF Choke Cross Reference for convenient wall-mounting or other fast reference. And, the new catalog contains an Index by part sequence to simplify ordering, plus a Table of Contents for easy reader reference.

Response to the new Catalog 111 offering is expected to be very high so Cambion anticipates 4-6 weeks delivery of this useful data after receipt of request. Letterhead inquiries will receive more prompt attention, and all requests should be directed to Cambion Customer Service at 445 Concord Avenue, Cambridge, MA 02238. (617) 491-5400. Telex: 92-1480. TWX: (710) 320-6399. INFO/CARD #118.

Ceramic Disc Capacitor Brochure

The Passive Components Division of Thomson-CSF Components is offering free a new eight-page, two-color brochure on its ceramic disc capacitor line.

The brochure describes the three types of ceramic disc capacitors presently available from Thomson-CSF: temperature compensating, general purpose, and barrier layer capacitors. For each of these three types, the brochure provides full technical data including capacitance ranges, technical characteristics, and performance graphs. In addition, details of capacitor markings and coatings, and full ordering information is included.

Contact Thomson-CSF Components Corp., 6660 Variel Avenue, Canoga Park, CA 91303. (213) 887-1010. Please circle INFO/CARD #113. □

Our SAW Filters give you less for your money.

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Today, Andersen SAW filters are priced competitively with other filter technologies. We have hundreds of standard designs suitable for thousands of applications. So you save on design charges. And if you need something tailor-made, our fast, computerized design capabilities and large production volume can reduce your costs far below what you'd expect.

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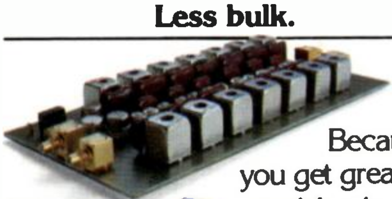
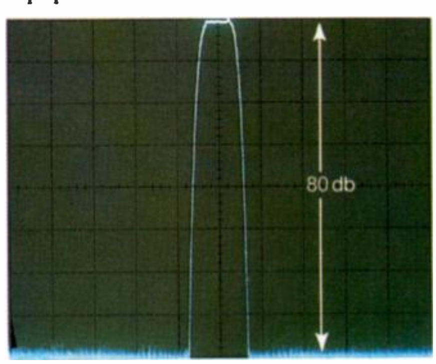


Andersen SAW filter vs. equivalent LC network.

Because you get greater precision in a smaller package. See for yourself how Andersen SAW filters stack up in size against LC filters.

Less tuning.

Because there's nothing to tune. Our filters are passive analog solid-state devices. They have permanently fixed performance characteristics. Once that design is checked by our computers, it never has to be checked again. Unlike LC filters, our SAW filters need no tuning, assembly, checking, retuning, and rechecking. So you save time, labor and capital equipment costs.



Typical frequency response of Andersen BPP70-1300-3-133A of 1.5MHz bandwidth.

Less maintenance.

Because there's nothing that can ever get out of tune. Our SAW filters not only cost less than you'd think to begin with. They won't nickel and dime you to death after you buy them.

Less rejects.

Because SAW filters of the same design are virtually identical. That helps you produce a more consistent end product. And our proven designs assure top performance. You get a superior shape factor. Improved close-in rejection. Inherent linear phase. And less distortion than with conventional filter designs.

Less confusion.

Because if there's one thing we give you more of, it's information. Volume I of our Handbook of Acoustical Signal Processing will tell you everything you need to know about the right SAW filter for your application. Plus, for a limited time, the Handbook is free. So send for your copy now.

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Andersen SAW products are available in the United Kingdom and Europe through our sister company, Signal Technology Ltd., Swindon, Wiltshire, UK.

INFO/CARD 1

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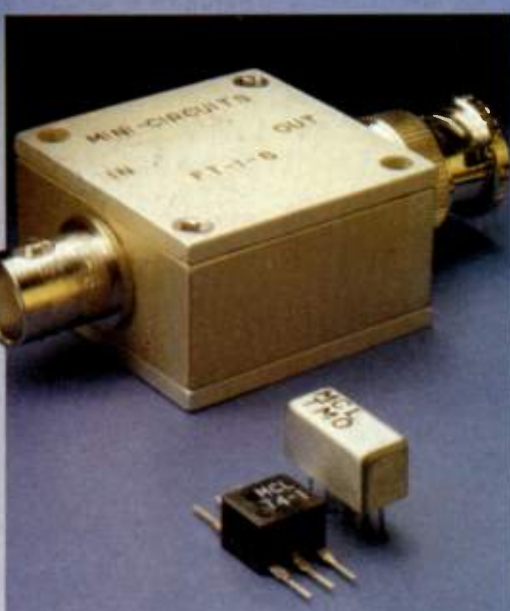

Select from the economical, microminiature T-series (plastic case) or TMO series (hermetically-sealed metal case) covering 10 KHz to 800 MHz. These models operate from 12.5 to 800 ohms with insertion loss typically less than 0.5 dB.

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

Need a connector version? Select from the FT or FTB series, available with unbalanced or balanced outputs. Connector choices are female (BNC, Isolated BNC, and Type N) and male (BNC and Type N). These units operate from 10 KHz to 500 MHz with impedances of 50 and 75 ohms.

Of course, Mini-Circuits' one-year guarantee is included.

DC ISOLATED PRIMARY & SECONDARY

Model No.	T1-1 TMO1-1	T1-1H	T1.5-1 TMO1.5-1	T2.5-6 TMO2.5-6	T4-6 TMO4-6	T9-1 TMO9-1	T9-1H	T16-1 TMO16-1	T16-1H
Imped. Ratio	1	1	1.5	2.5	4	9	9	16	16
Freq. (MHz)	.15-400	8-300	1-300	.01-100	.02-200	.15-200	2-90	.3-120	7-85
T Model (10-49)	$2.95	$4.95	$3.95	$3.95	$3.95	$3.45	$5.45	$3.95	$5.95
TMO model (10-49)	$4.95		$6.75	$6.45	$6.45	$6.45		$6.45	

CENTER-TAPPED DC ISOLATED PRIMARY & SECONDARY

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

UNBALANCED PRIMARY & SECONDARY

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

FT FTB

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

Mini-Circuits

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