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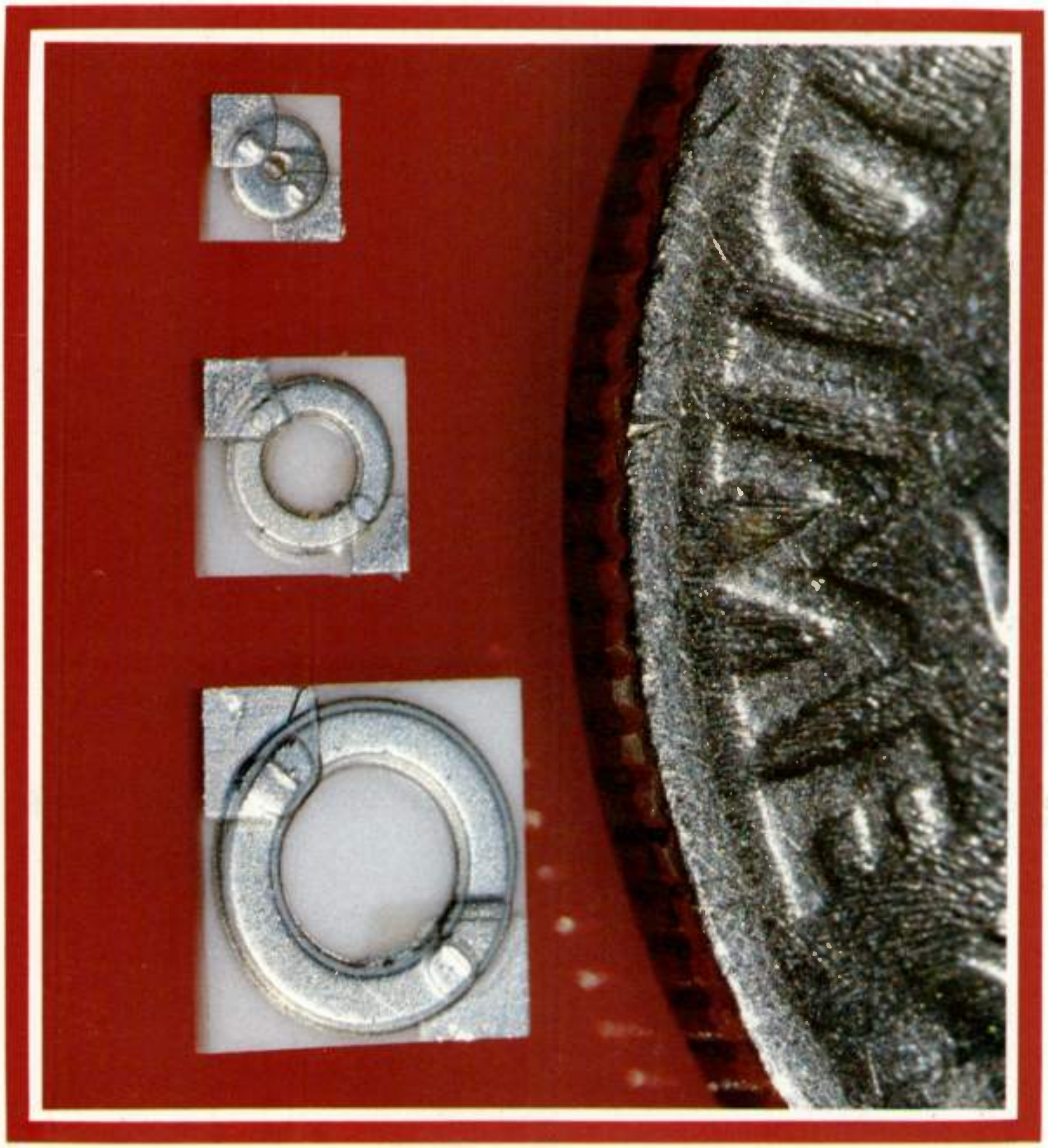

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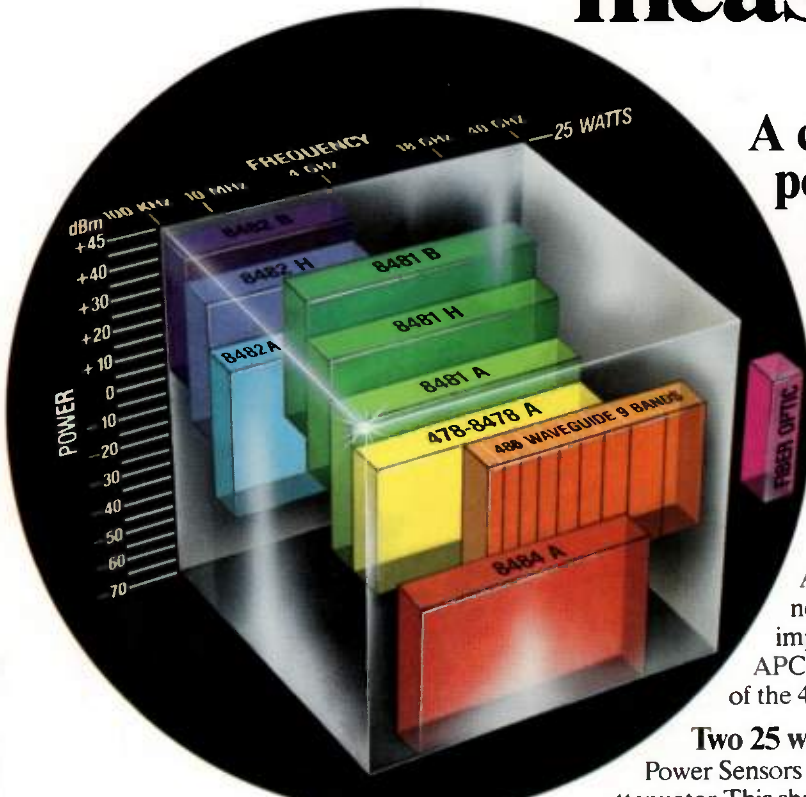

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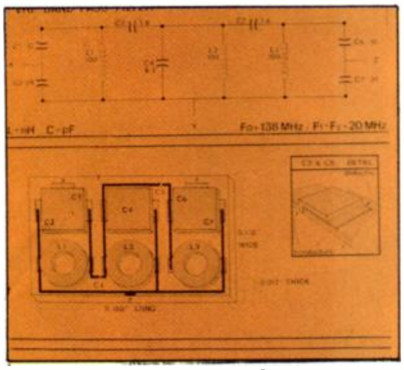
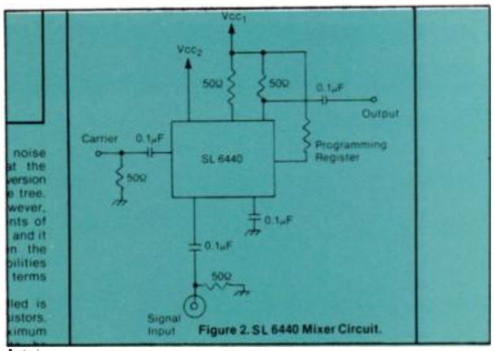


Pico Chip Passive RF Components



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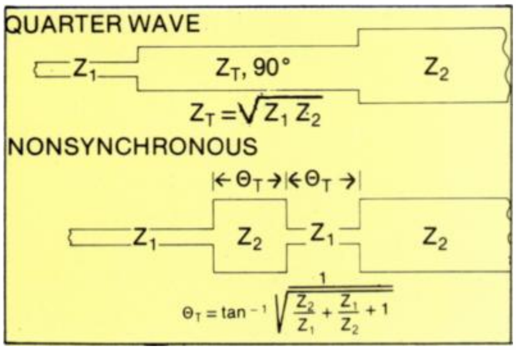

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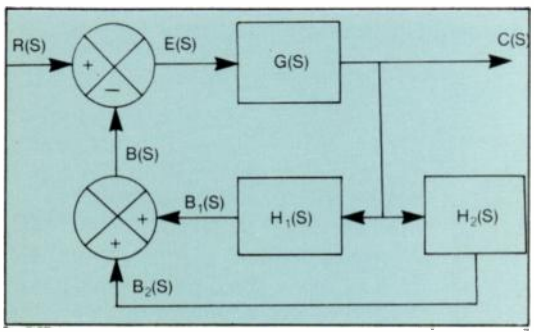

AGC Loop Design Using Control System Theory

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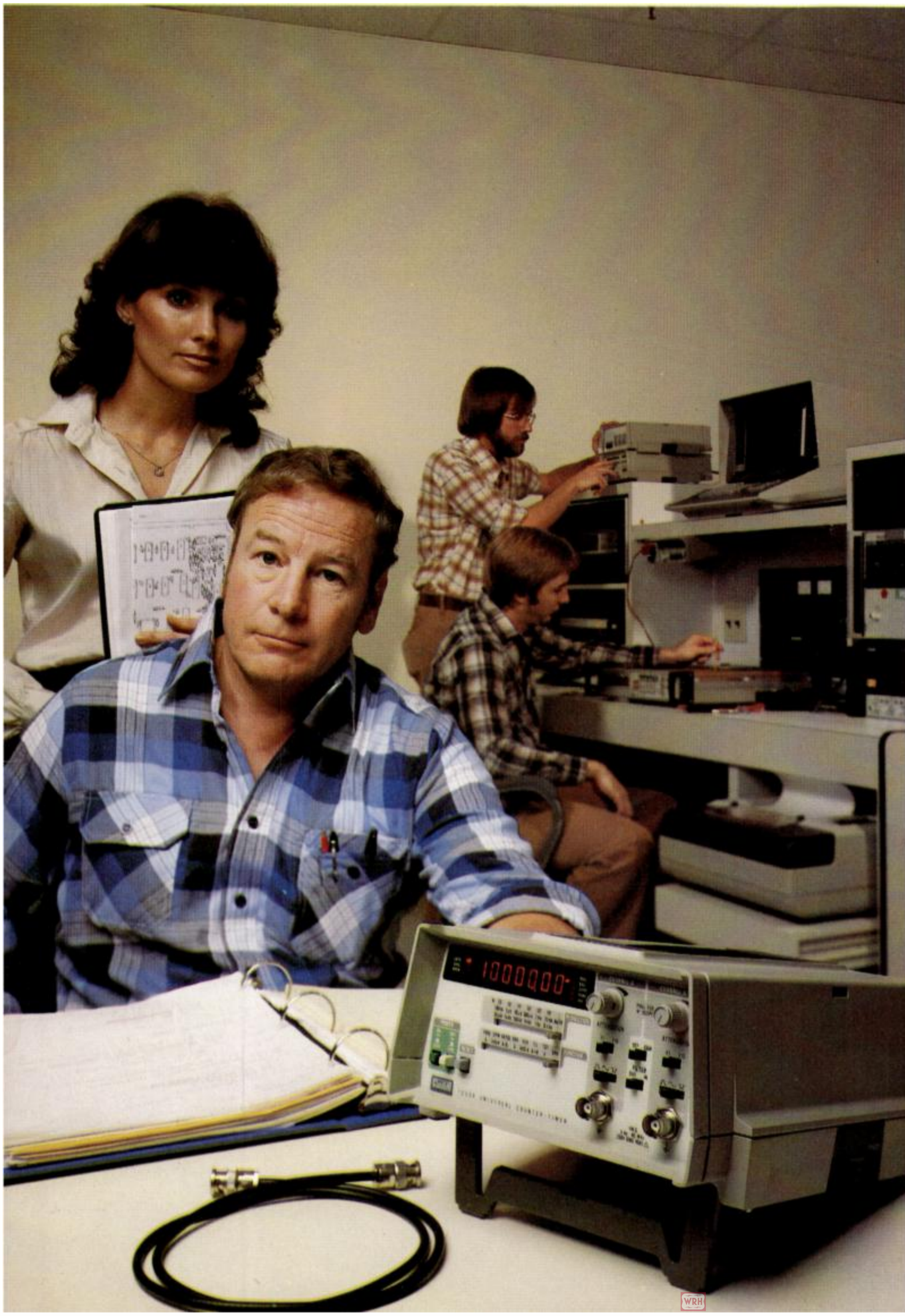

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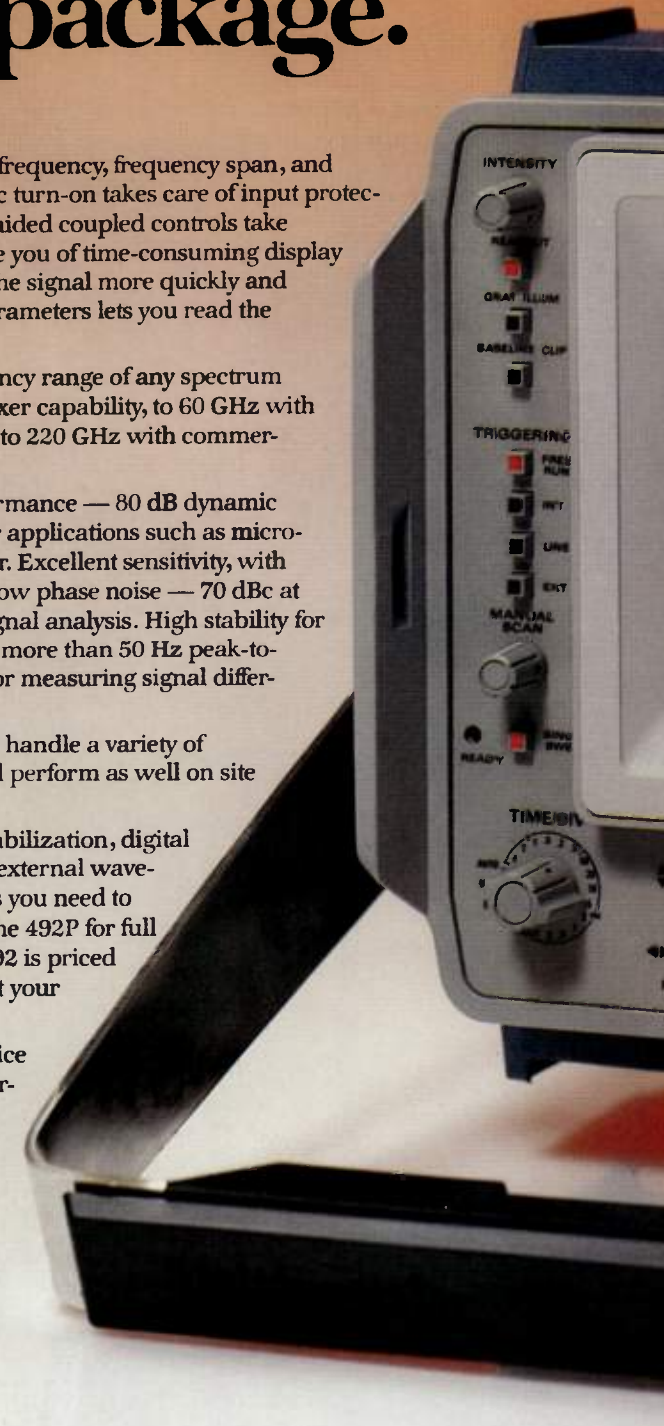

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The Last Obstacle

It does my heart good to see a job well done; an engineering task from start-to-finish attacked in the good ole scientific method. Several engineering types from a small company decided a particular problem (and thus need) existed, studied current approaches, decided new techniques were indicated and proceeded to develop them and *succeeded.*

Outstanding advances in microminiaturization of active circuits and packaging techniques have occurred in the recent past affecting the design engineer of every discipline. The last significant hurdle to overcome was (in this Editor's opinion) that of the lumped L and C component size.

Picominiaturization is the subject of this month's lead-off article. It is an exciting progress report on *not* only smaller inductors and capacitors with a technique that lends itself to consistent repeatability of results but most importantly on the (realizable) promise of integration of these components into multi-element filters and larger functional groupings.

An example of this is a 10 element VHF Band Pass Filter. A 2 inch by 2 inch substrate yields 210 of them. The future for the RF engineer is exciting. Developments like this are few and far between and will probably have ramifications not yet envisioned in our times. I wish success to the small company and their components and techniques.

Standing room only was the order of the day at several of the technical sessions of Electro/80 held this year in Boston. Naturally, they were programs centered on techniques for the *RF* engineer (sessions 5 and 8 hosted by Doug DeMaw). In my estimation the entire show was a success with probably more than 30,000 in attendance over the three day period. It's hard to think recession while elbowing one's way through literally thousands of excited engineers. It was a pleasure meeting many of you there and I look forward to a repetition at the upcoming Wescon and Southcon Shows in the months ahead.

To the many who have responded recently by means of the feedback cards — thank you. For an editor it is extremely gratifying to know that you enjoy the magazine, articles, format, etc. *Every* card is read and your constructive criticism, ideas, suggestions are given much thought *and* action.

To the lone gentleman from Aurora, Illinois — *NO*! — the contest was not rigged and all ballots are here in the office for your perusal.

Ed Oxner — welcome to our Editorial Review Board.

Without further ado, turn the page and let's get into "Pico Chip Passive RF Components."

Rich Rosen

Rich Rosen

Pico Chip Passive RF Components Inductors — Capacitors — Resistors

Making them smaller and smaller via a new multi-layer thin film manufacturing technique.

Hal DePalma
Thinco
Hatboro, PA

In this age of ever increasing microminiaturization, the RF design engineer has had problems in achieving significant reductions in the size and weight of his analog circuits. The problem has been one of component size constraints particularly in the area of passive components. (Figure 1 graphically illustrates the difference in size between a picominiaturized inductor and capacitor, two center components, and their conventional counterparts, shown either side of them.) In sum, the lumped constant inductor, capacitor and resistor have not kept pace with the semiconductor in terms of subminiaturization for very high packaging densities. This is clearly indicated in Figure 2 where a lumped capacitor far exceeds, in size, the adjacent active integrated circuit.

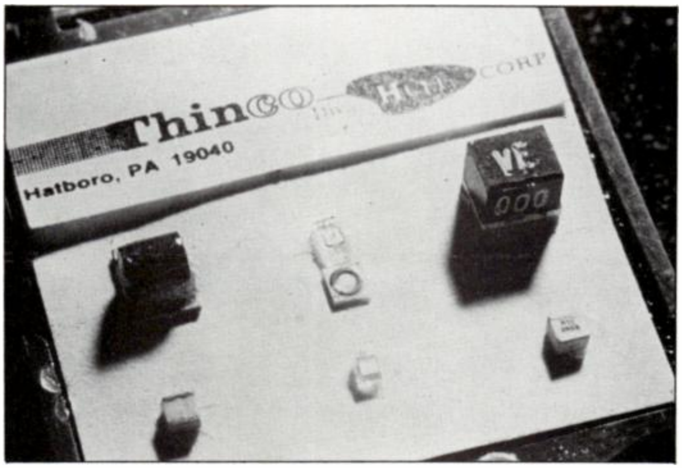

Figure 1.

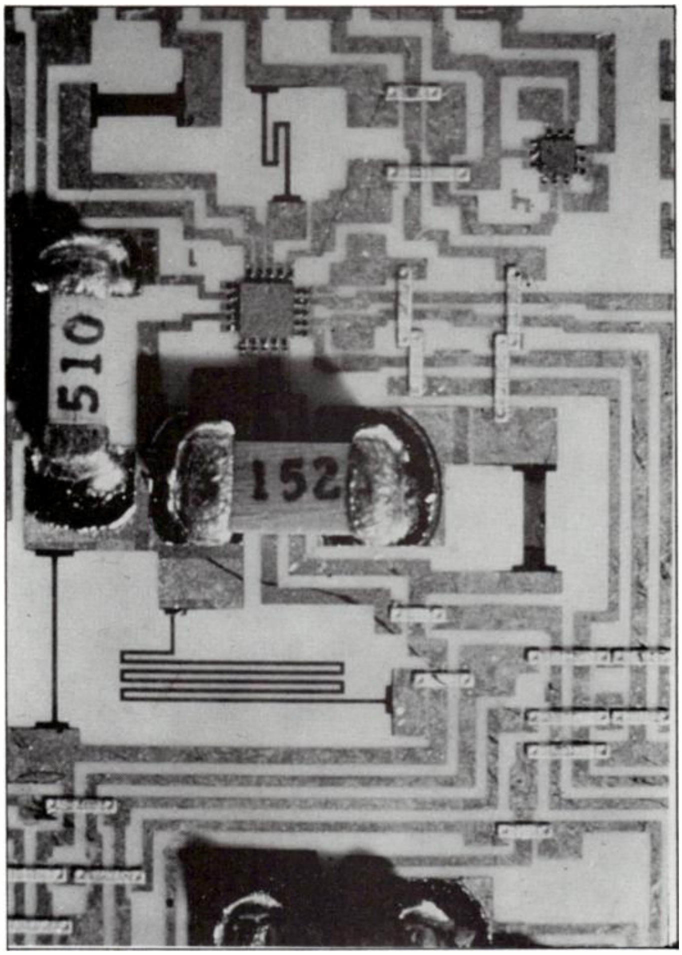

Figure 2.

Another Look

The Thinco Division of Hull Corporation was formed several years ago with the singular mission of devising techniques that would effectively deal with the passive component size problem. To establish a sense of direction, an in-depth study was conducted of the available miniaturized components with particular emphasis upon contemporary fabrication techniques and material systems. After considerable deliberation, Thinco engineers reached the conclusion that a new fabrication technology was required to obtain a significant advance in the state-of-the-art.

New Directions

Since a "ground zero" approach had been chosen, it was necessary to establish several criteria and then search out a manufacturing process approach that would satisfy these requirements which were as follows:

- Obtain a size reduction of at least 10:1 (X-Y-Z axis volume) over presently available components of like function.
- Devise a method that would enable the inductor, capacitor and resistor to be made in exactly the same way and simultaneously.
- Develop a manufacturing process exhibiting extreme repeatability, rendering precision tolerance control and lending itself to complete mechanization for cost effectiveness.

Underlying these criteria was the desire to develop a technology that would enable the same transition from the discrete chip component to the monolithic integrated circuit as had been accomplished by the semiconductor industry in the 1960's.

Making It Happen

With all of the above in mind, the final choice of a manufacturing process approach was multi-layer vacuum deposited thin films.

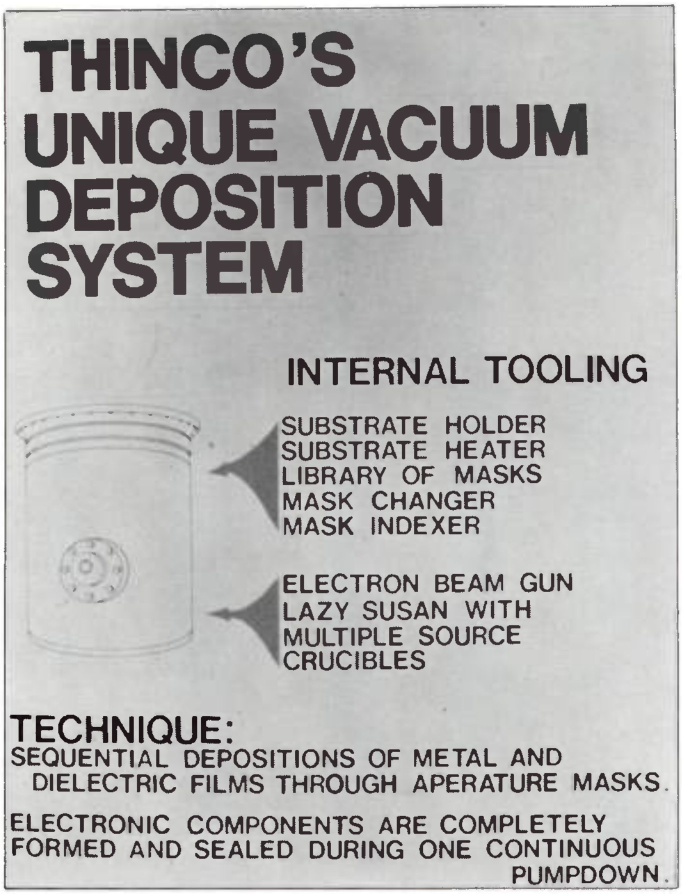

Figure 3.

The manufacturing process that evolved has been the result of considerable development of proprietary tooling, process techniques and special materials. Each of these three areas presented their own problems that had to be resolved to obtain the product performance levels being sought. At this time, product performance is at about the same stage as were transistors — around 1960 and constantly improving.

As shown in Figure 3 the important elements of the Thinco thin film processing equipment are contained within the vacuum system bell jar. The internal tooling is quite sophisticated with many unique features. (See Figure 4.)

It is the ability to sequentially deposit metal and ceramic thin films with precision thickness control through a variety of masks with precision aperatures and precision mask indexing that enables extreme control over the component geometry and the resultant tight parameter tolerances.

The discrete passive chip components are made in somewhat the same way that discrete transistor and diode chips are fabricated. Typically, hundreds or thousands of identical inductors are deposited simultaneously in a step-and-repeat pattern (Figures 5) on a 2" x 2" alumina substrate. Thinco's process is totally additive and the component structures are completely formed and hermetically sealed under one continuous vacuum thereby eliminating entrapment of foreign materials. Ultimately the individual components are diced with a digitally controlled diamond blade saw.

Figure 6 is a back-lighted photograph of one of six masks used in the fabrication of a typical inductor. The semicircular apertures are designed for half turn

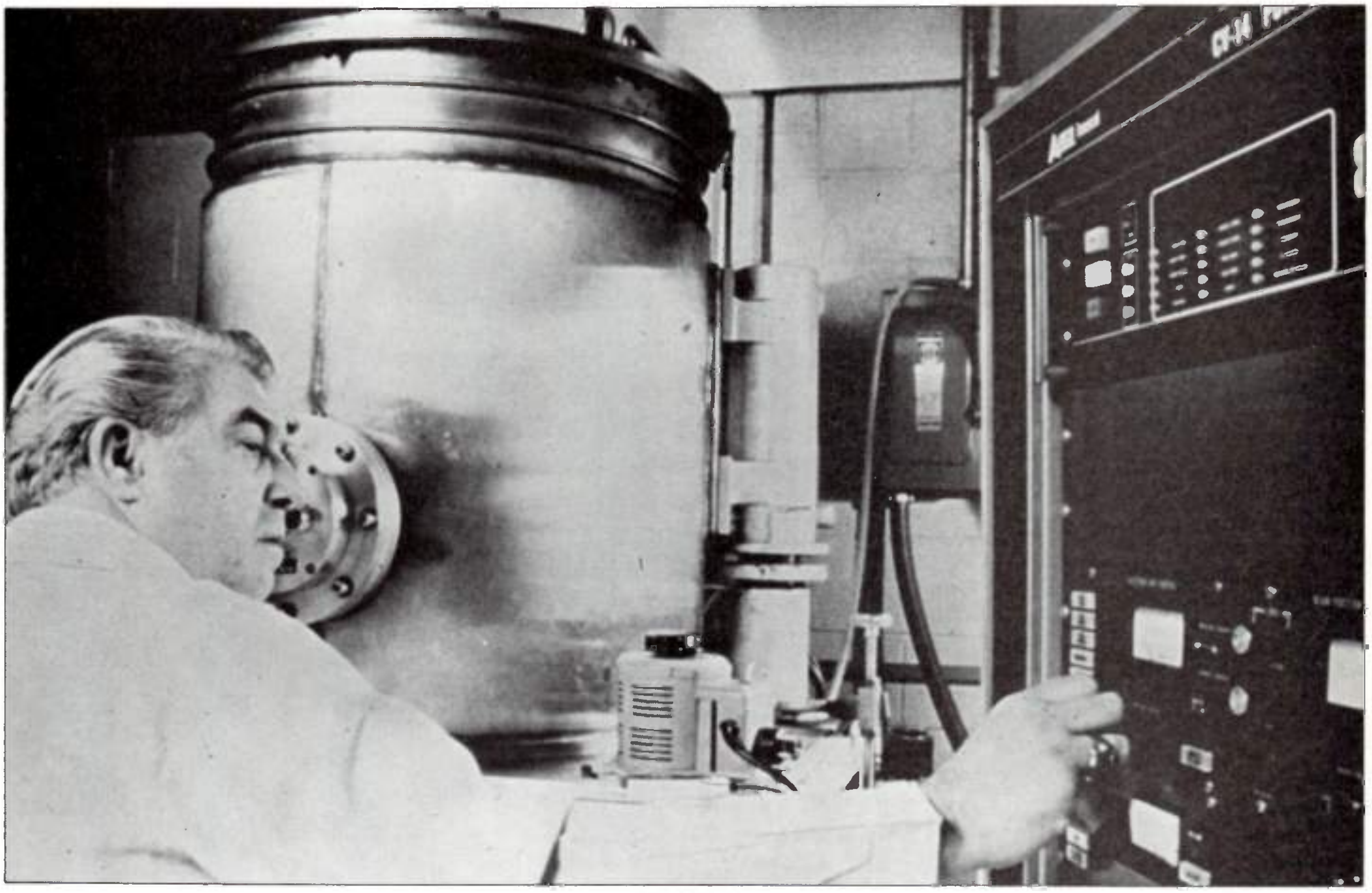

Figure 4.

Figure 5.

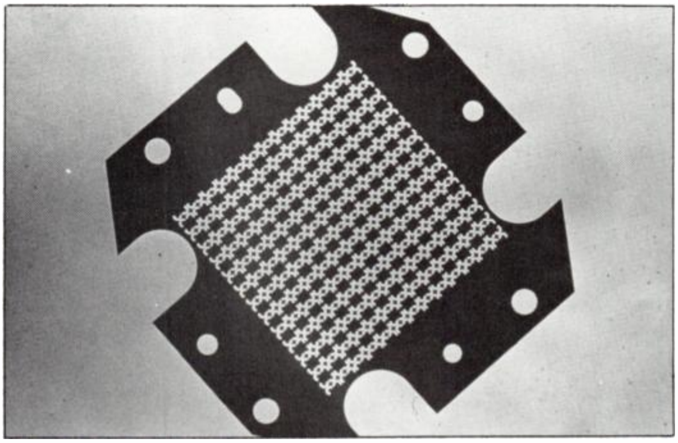

Figure 6.

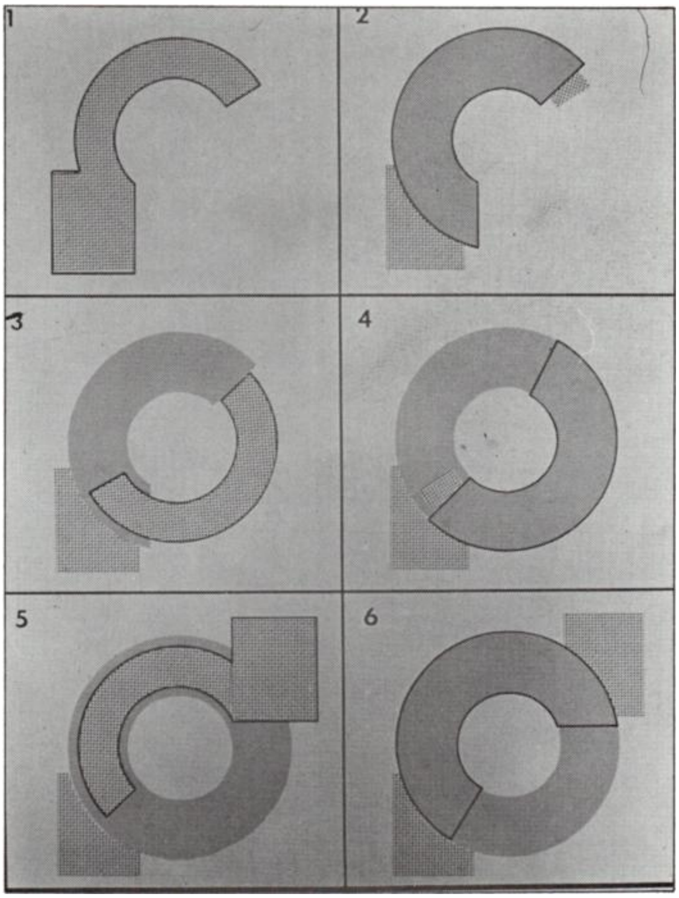

Figure 7.

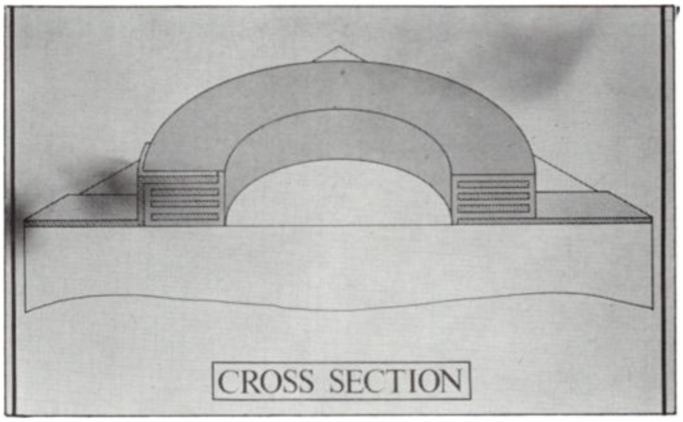

Figure 8.

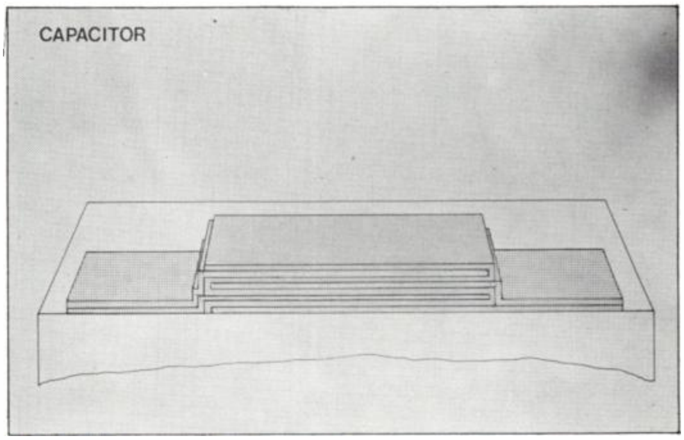

Figure 9.

depositions which are accomplished as described below.

Looking at just one component location on the substrate, Figure 7 illustrates the sequence of depositions used in the formation of an inductor:

Step 1: Position mask #1 upon the substrate and deposit aluminum in a pattern to form the first termination pad and the first half turn of the "winding".

Step 2: Position mask #2 upon the substrate and deposit a ceramic dielectric in a pattern leaving the termination pad and the tip of the first half turn exposed.

Step 3: Position mask #3 upon the substrate and deposit aluminum to form the second half turn.

Step 4: Position mask #4 upon the substrate and deposit ceramic over the second half turn leaving the tip exposed.

(The half turn depositions are repeated until the required number of turns for a specific inductance value are obtained.)

Steps 5 and 6: The last half turn with the second termination pad and the final ceramic encapsulation seal are deposited.

The cross sectional views in Figure 8, 9 and 10 illustrate how true three dimensional geometries are obtained with the above multilayering technique. Currently, up to 25 layers are being deposited with all of the completed component structures having a height less than 0.003".

Close examination of the cross-sections also reveal how the same process is used to form all three components which can be done simultaneously to fabricate networks. Additionally, complex multi-layer printed wiring interconnects can also be incorporated.

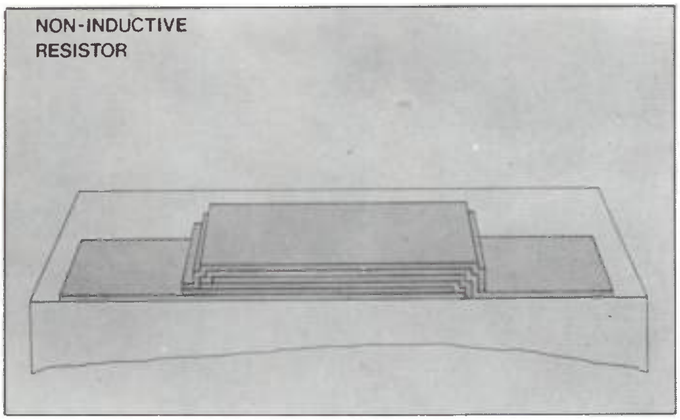

Figure 10.

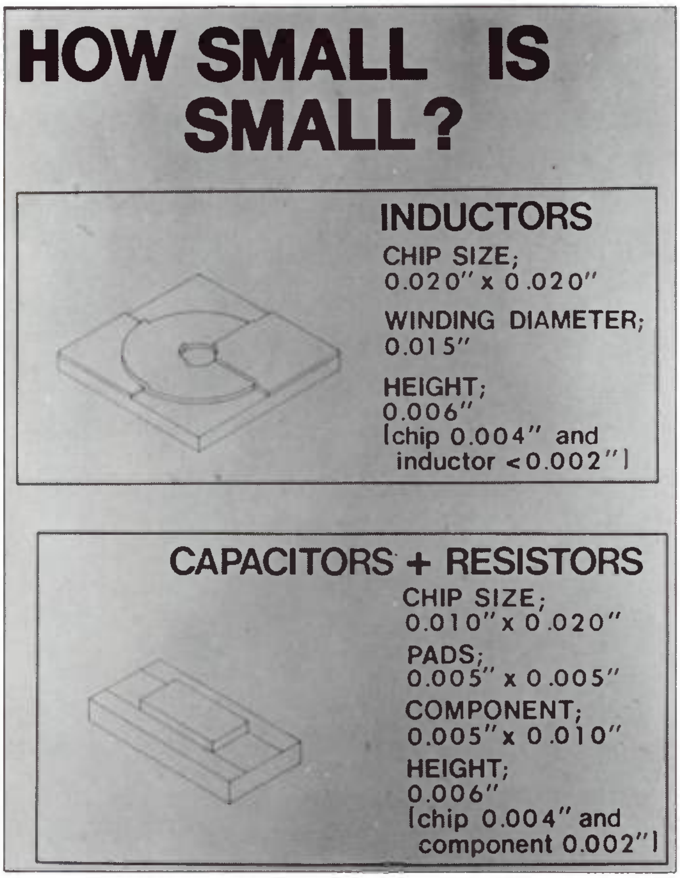

Figure 11.

Q?

How Small Is Small?

The three inductors pictured in Figure 12 are the L-30C, L-55C and L-100C with chip dimensions of 0.047", 0.068" and 0.117" on a side. Figure 13 shows the companion MWCT-1 capacitor on a 0.04" and 0.05" chip. The height of all component chips is determined by the combined thickness of starting substrate and the deposited structures which could be taken down to 0.006" total if required. Figure 14 shows a profile view of two PC boards. The upper one has conventional lumped components and the lower picominiature components.

The size of these first inductors and capacitors were chosen strictly for ease in handling during the development and characterization phase. They do not represent the smallest components that could be made with current techniques.

How small is small is a function of mask making

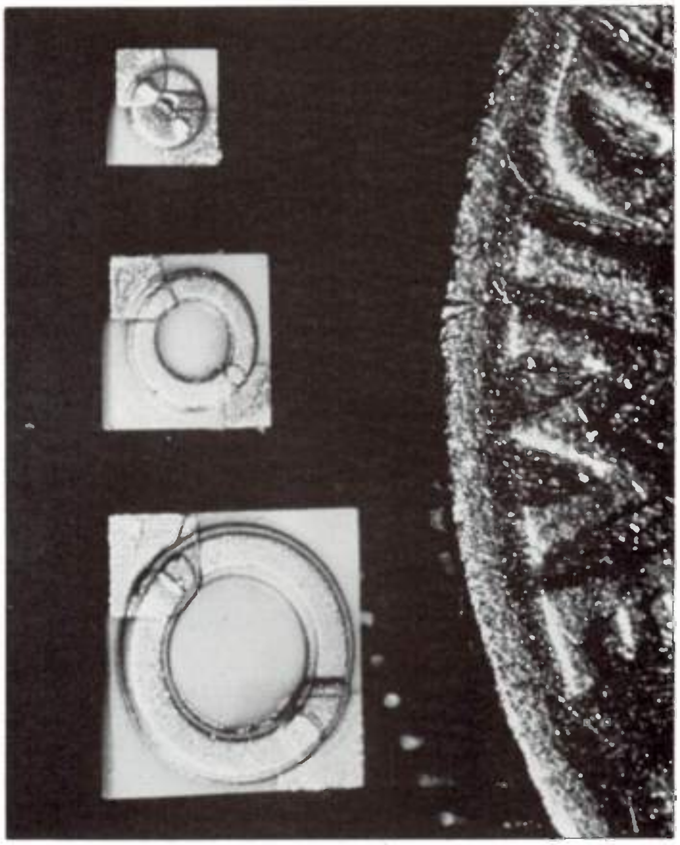

Figure 12.

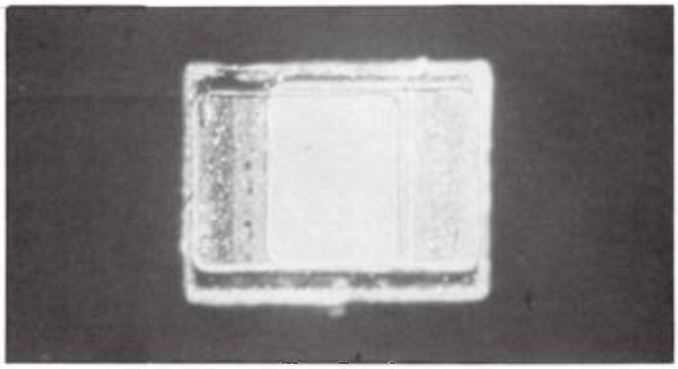

Figure 13.

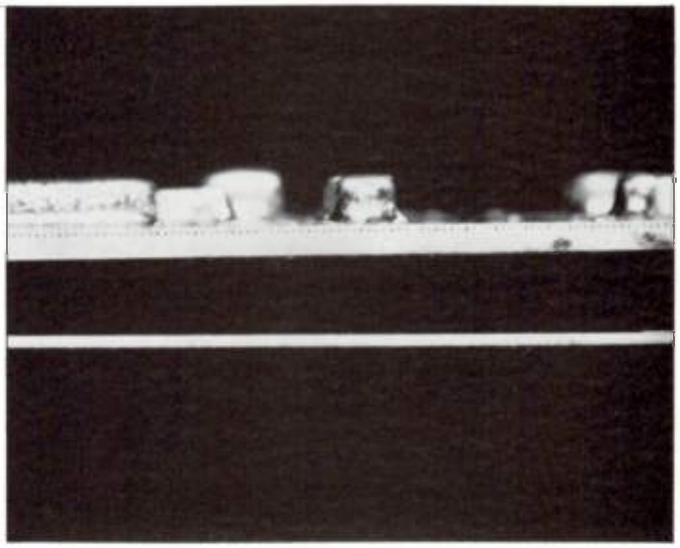

Figure 14.

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technology. It is a question of how small the apertures can be made with precision tolerances for a given mask thickness. There are several technical trade-offs involved in this matter but a safe figure at this time would be 0.005".

Size reduction in the inductor does impose certain restrictions upon inductance value. However, with the aid of a ferrite substrate as shown in Figure 15, impressive values of inductance into the microhenry range have been obtained.

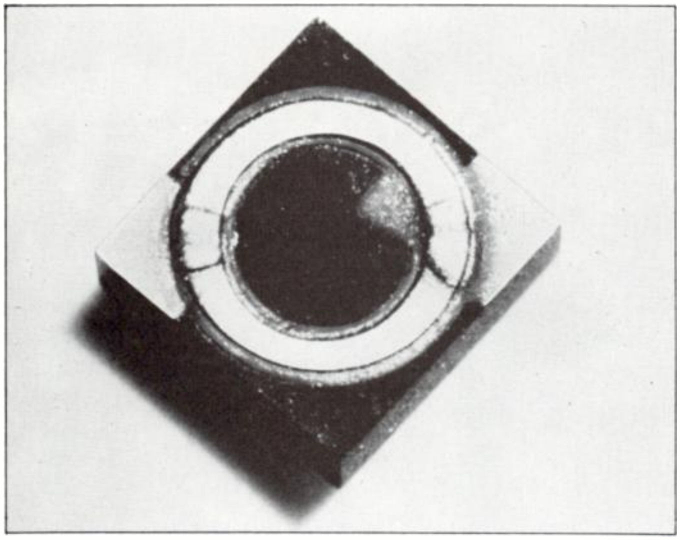


Figure 15.

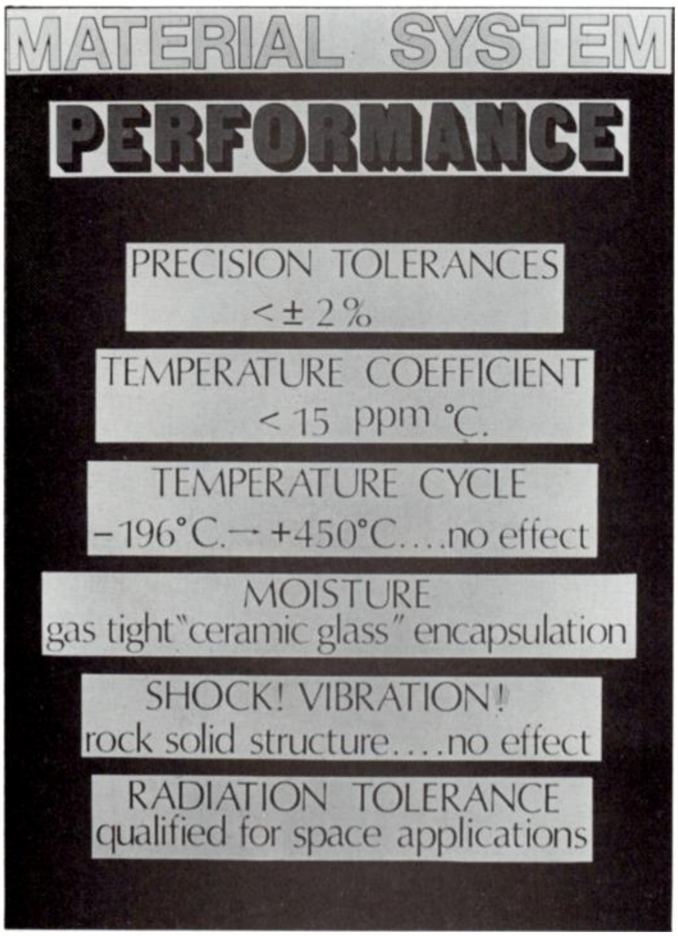


Figure 16.

Material System Properties

Size alone is not enough particularly in stringent military/aerospace applications. Fortunately, as the product evolved and a certain material system was developed, other highly desirable properties appeared as shown in Figure 16.

Geometry Properties

Additional desirable properties attributable to the geometry made their appearance during characterization tests that made the new components even more versatile.

1. Power Handling: As shown in Figure 17 the components are planar with a very short thermal path. With heat sinking, exceptionally high current levels can be carried.

2. Self-Resonant Frequencies: The very small geometries inherently have very low parasitic capacitance and inductance. The inductor can be scaled to operate out to about *5 GHz* and the capacitor out to *20 GHz* or beyond.

3. Mutual Coupling: Two inductors mounted side by side exhibit exceptionally low mutual coupling thereby eliminating the need for shielding in most cases.

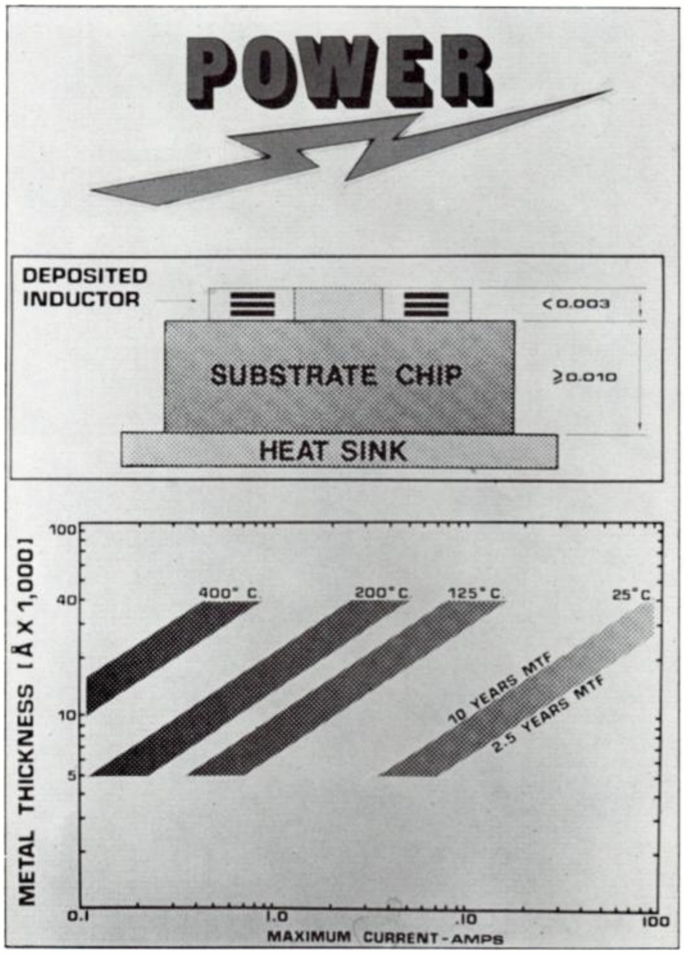


Figure 17.

A Look Into The Future

With the basic feasibility of the multi-layer thin film approach having been established, a logical product variation progression suggests itself. Figure 18 displays some

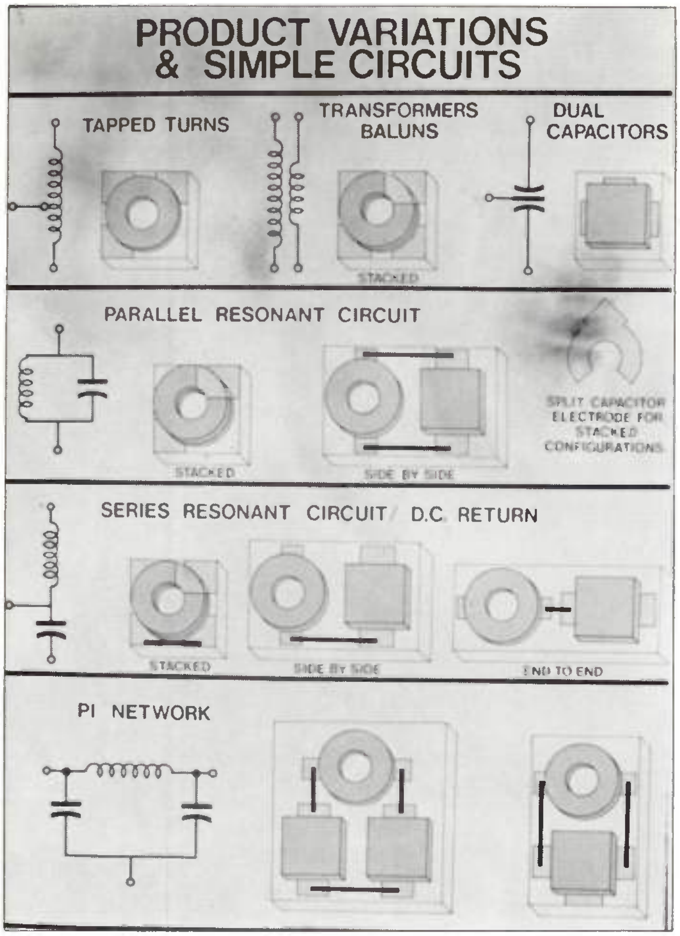


Figure 18.

of the anticipated variations including simple two and three component monolithic integrated circuits. None of these products are tooled as yet but will be as market demand dictates.

The ultimate technical and cost saving benefits of this new technology will be obtained at the monolithic integrated circuit level. Every aspect of the technology lends itself to computer aided design and computer aided manufacturing. Additionally, tooling for a new product is simply a matter of masks which are relatively inexpensive and easy to obtain.

The first step toward monolithic integrated circuit capability is now underway via a U.S. Navy development contract #N66269-80-C-0024 from the Naval Air Development Center, Warminster, Pa. Figure 19 illustrates the utilization of picominiature discrete form inductors in a 2 channel VHF/UHF converter. The objective of the contract is to develop and fabricate monolithic chip bypass filters consisting of one each capacitor and inductor in a three

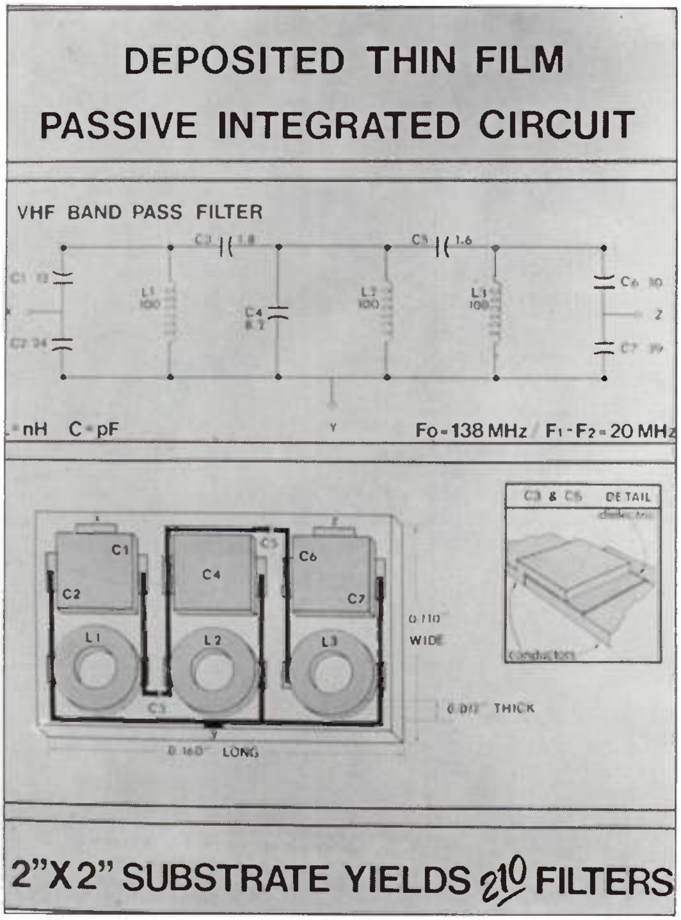


Figure 20.

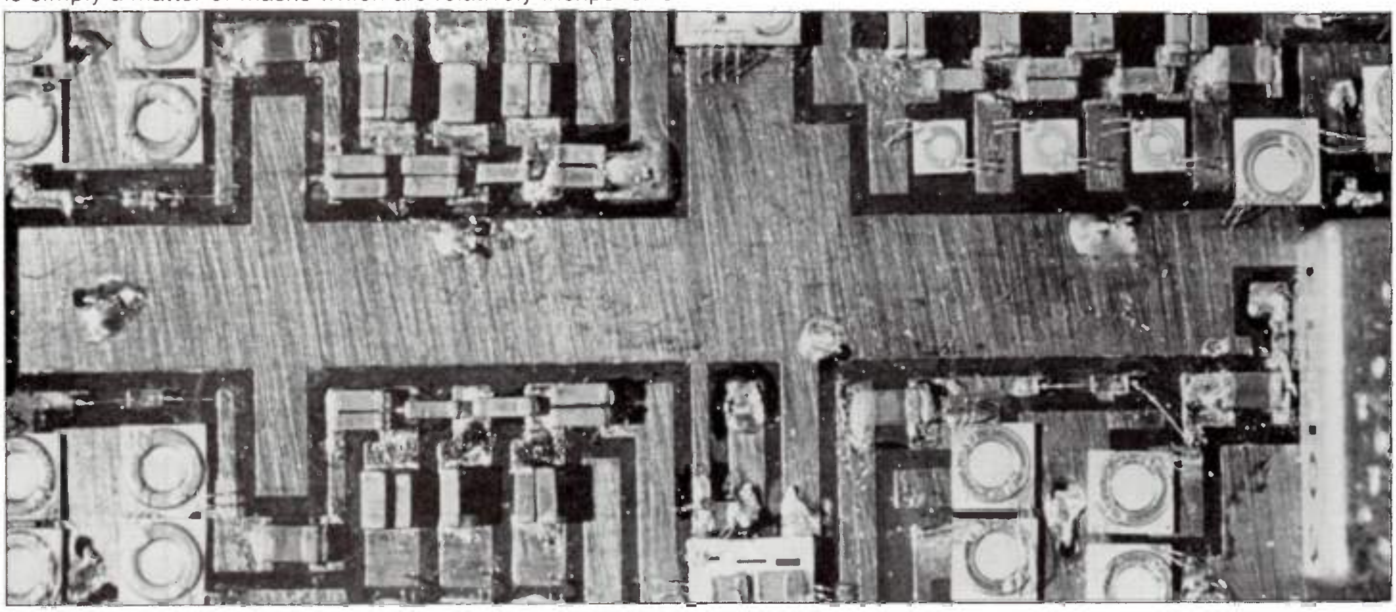


Figure 19.

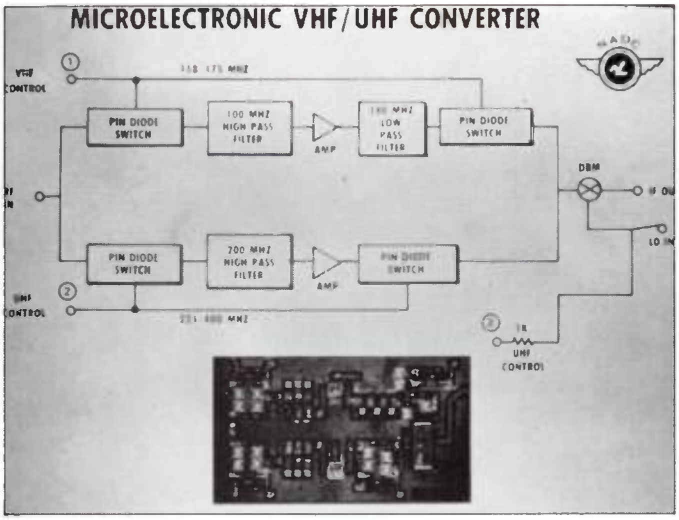


Figure 21.

terminal series configuration. Both components will be deposited simultaneously and the L-C ratio will be varied over a considerable range with the same masks by means of special deposition techniques. This work is being performed for an avionics program under the guidance of Mr. Elliott Ressler who was among the first to recognize the potential of the multi-layer thin film technique.

The simple integrated circuit capability will be established this year which will set the stage for more complex medium scale integration such as a bandpass filter (Figure 20) containing 10 components. Certainly, large-scale integration would be the next step. (See Figure 21.) Cordless telephones and a host of other RF related products in wrist-watch cases are a future prospect. (The wrist-watch TV so prevalent in sci-fi cartoons of the past may indeed be nearer to realization using pico-miniaturization techniques as humorously illustrated in Figure 22.)

The direction that potential future military/aerospace development programs might take are blocked out in Figure 23.

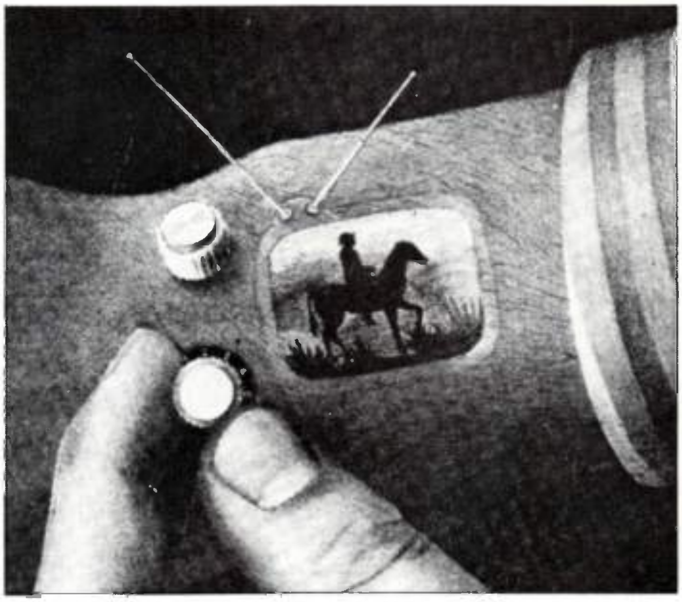


Figure 22.

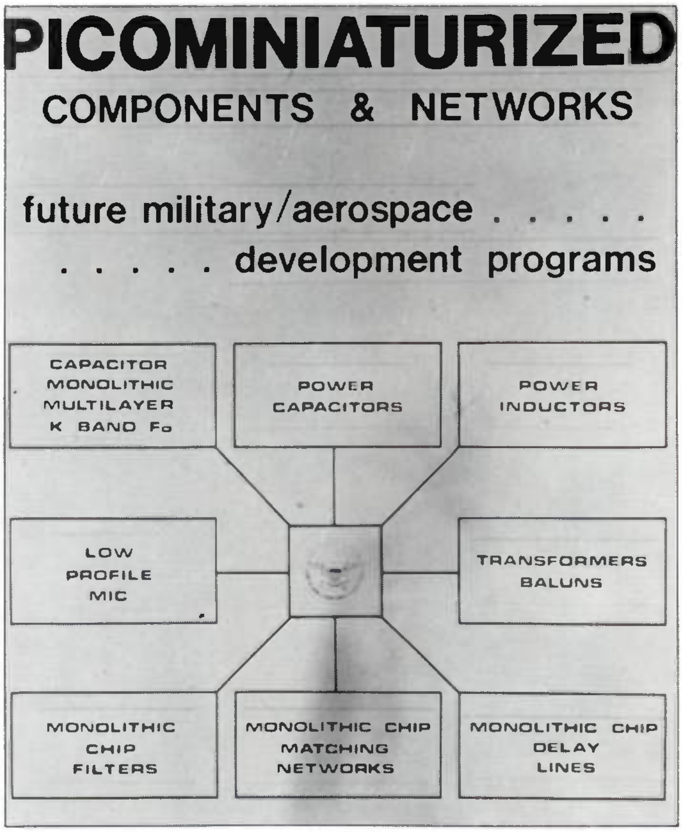


Figure 23.

Much of the credit for making this technology a production reality belongs to Dr. Stephen T. Chen and Leonard Altman who are key members of the Thinco staff.

Correction to "Chebyshev Filters Using Standard Value Capacitors"

Dear Editor:

While reviewing the calculations in my article, "Chebyshev Filters Using Standard-Value Capacitors," *r.f. design*, February 1980, I found a simpler procedure for transforming the pre-calculated 50-ohm designs to any other impedance level. This simpler procedure does not require the use of the normalized element values of Table A-1 — only Tables 1 and 2 are required for the lowpass and highpass transformations.

The first five steps of the simpler procedure are identical with the originally published procedure. The two concluding steps to calculate the new cutoff frequency and the new inductance values are as follows.

(6) Calculate the exact F_3 frequency of the selected design by dividing the F_3^{50} of the tabulated design by the termination resistance ratio, $R_r = R/50$.

(7) Calculate the new inductor values by multiplying the tabulated L values of the selected design by the square of the resistance ratio. [Both capacitance values, C1,7, and C3,5, are taken directly from the table.]

This concludes the transformation procedure.

Repeating the example in the text using the simplified procedure where a 75-ohm lowpass filter was desired with a 3-dB cutoff frequency of 3.00 MHz: $R_r = 1.5$ and $F_3^{50} = 4.5$ MHz. Choose lowpass design #40 which is the design closest to 4.5 MHz. $F_3\#40 = 4.645$ MHz with $C1,7 = 820$ pF and $C3,5 = 1500$ pF. (These will be the capacitance values of the new 75-ohm filter.)

From new step (6):

the new $F_3^{75} = (4.645/1.5)\text{MHz} = 3.0967$ MHz.

$L2,6 = (1.5^2)\cdot 2.64\ \mu H = 5.94\ \mu H,$

$L4 = (1.5^2)\cdot 2.94\ \mu H = 6.615\ \mu H.$

If the results obtained with this simpler procedure are compared with the originally published results, they will be seen to be essentially identical. The same procedure is used for transforming the 50-ohm highpass filters to different resistance levels. Note that the highpass transformation procedure is considerably shortened with the simpler procedure.

I would appreciate it if you would bring this simplified filter transformation procedure to the attention of your readers. I apologize for any inconvenience that the more involved (although correct) procedure may have caused them. Those who would like a full-page copy of Tables 1 and 2 may obtain copies by sending me a stamped self-addressed envelope.

Ed Wetherhold

S-Parameter Amplifier Design With Your HP-67, May '80 *r.f. design*

The following corrections to the program listing for card #2 are indicated below:

Step 109	STO 9	33 09
Step 112	f GSB 3	31 22 03

Both steps were inadvertently omitted.

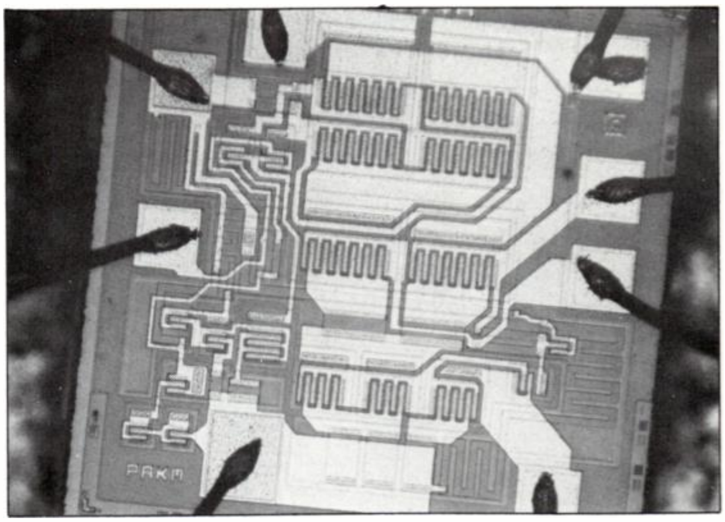
The program should read as follows:

Step	Key Entry	Key Code
108	RCL 5	34 05
109	STO 9	33 09
110	f GSB 5	31 22 05
111	RCL 9	34 09
112	f GSB 3	31 22 03
113	f GSB 5	31 22 05

Thanks to Ed Oxner of Siliconix who first brought it to my attention ... Editor

"RF Coaxial Cables. . . " May '80, Page 14, Formula 6 should read: $V_f = \dfrac{100}{\sqrt{E}}$

High Performance Integrated Circuit Mixers



Design considerations, typical performance and the merits of the double-balanced I.C. mixer are discussed. Mixer terminology are detailed in an appendix.

P.E. Chadwick
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Swindon Wilts, England

The doule balanced mixer has, in the last 10 years, become as ubiquitous an RF component as the attenuator or coaxial connector. The application of active devices to this area of RF circuitry has been slow, basically because of the failure of active devices to meet the performance requirements. Currently, packaged double balanced mixers are available from a multiplicity of companies, and offer bandwidths of 5-500, 10-1000 MHz or even 2 or 3 GHz at low cost. These mixer packages contain transformers (sometimes as transmission line hybrids) and Schottky diodes. Prices range from $2 to $200, depending upon frequency range, compression point, intermodulation intercept point, local oscillator power requirements, connector types, packaging format, temperature range, etc.

The majority of these devices have two parameters in common however, — maximum operating temperatures of 85°C and high local oscillator driver power. The diode ring mixer (Figure 1) has been exhaustively analyzed in various literature over the years, and is now fairly well understood — which may account partially for its popularity. Nevertheless, it suffers from some major defects, and these are particularly:

- The intercept point is usually critically dependant upon the load impedance.
- High intercept and compression points require high drive power.
- Local Oscillator (L.O.) leakage of energy to input and output ports is usually appreciable (around 40dB).
- There is a conversion loss of around 6dB.

There have been several attempts to get around these disadvantages. By suitably terminating the output of the mixer, it is possible to reduce the loss to around 3 dB, but this is achieved at the expense of the intermodulation intercept point. Replacing the diodes as switches by MOSFET's has been tried to achieve a reduction in L.O. power requirements. This has introduced the difficulty of intermodulation distortion production because of modulation of the MOSFET 'ON' resistance by the signal. Finally, J FET mixers have been produced which give reasonable performance, but do not seem to be able to compete in terms of cost, especially when the external circuitry that is required is considered.

The integrated circuit mixer represents another approach. It offers the potential advantages of low cost, low L.O. drive requirements, gain and the stability of signal suppression becauce of the close thermal tracking of transistors on the monolithic chip. However, very few of these circuits have offered very good intermodulation response figures, and those that have, produce high noise figures.

Against this technical background, then, Plessey Semiconductors looked at improving the integrated circuit double-balanced mixer. The market size available is obviously large, and the sectors into which the IC mixer can offer competition are those where the frequency range is low (under 200 MHz), the availability of gain is useful, a noise figure higher than the ring mixer is immaterial, and costs are important. During the design phase it was realized that a device meeting the full MIL temperature range could be produced if required, and, combined with a process giving adequate reliability factors with junction temperatures up to 175° C, the possibility of a significant advance became apparent.

The equipment markets for this device are large: HF/VHF radio receivers for marine, civil, military, and aircraft use, including CB and amateur HF SSB transmitters, frequency synthesizers, test equipment and many other areas. However, the wide divergence of these markets means that the requirements of one market are unlikely to be met by those of another. Either a family of devices are required, or some way of broadening the spectrum of application for one device has to be achieved. This is done by programming the device performance externally by means of a current and allowing wide choice of supply current.

Design Considerations

The basic requirements for a mixer for general application suggests that the use of a single balanced mixer is not really acceptable, and that the double balanced form is required. This suggests that the "tree" arrangement should be used (Figure 1.) Consequently

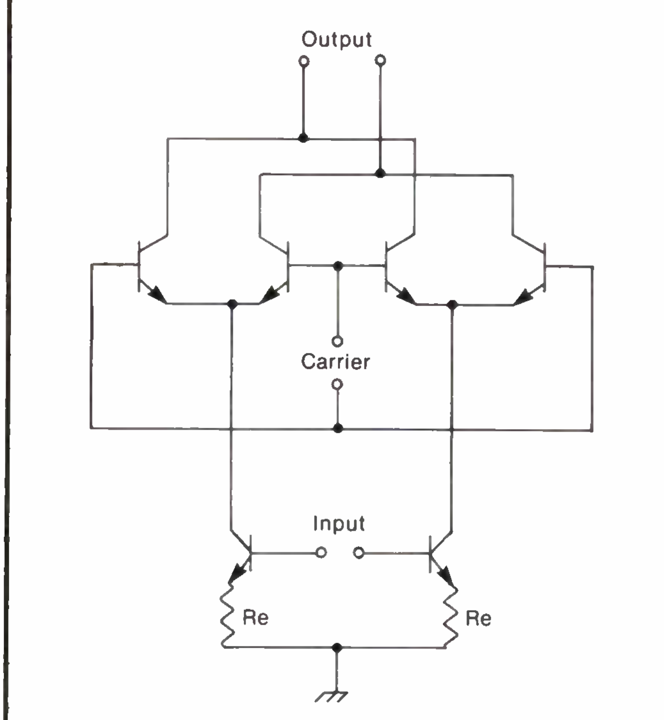

Figure 1. Diode Ring Mixer.

some investigation is required into the sources of noise and non-linearity. This investigation showed that the primary non-linearity was in the voltage-to-current conversion in the bases of the transistors at the bottom of the tree. the voltage-to-current conversion is linearized. However, at the expense of the noise figure. The requirements of linearity and noise figure are mutually contradictory, and it is necessary to achieve a compromise between the two. This compromise is complicated by the capabilities of the semiconductor manufacturing process in terms of resistance values.

The maximum signal level which can be handled is dependent on the current flowing through these transistors. Higher signal levels require higher currents. For maximum flexibility in all applications it is necessary to be able to vary the current through the tree, and in the SL6440, this current is designed to be externally programmable, allowing operation over a wide range of conditions. However, for the ultimate compression point and intercept point capabilities to be achieved the dissipation can become quite high, and would require the use of a heat sink.

The current is programmed by the use of a current mirror, such that the current programmed through each transistor at the bottom of the tree is equal to the programming input. The total current drawn through the open collector outputs is twice the programming current.

One of the major problems of the diode ring mixer is that in most cases intermodulation distortion production is critically dependent on the load at the output port. Certain more complex mixers avoid this, at the cost of extra hybrids. However, the transistor tree is non-critical in this respect, and can be said to be independent of load, as far as IMD is concerned. However, the choice of load impedance allows the available gain to be chosen, although higher voltages are required if the output stage is not to compress because of signal swing exceeding the supply.

Typical Performance

Very wide range of operating modes are possible due to the flexibility inherent in using a programmable design with differential inputs and outputs.

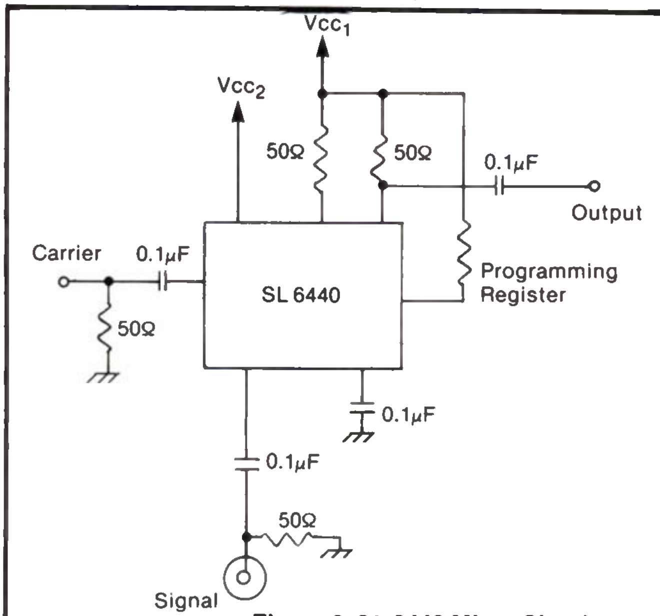

Figure 2. SL 6440 Mixer Circuit.

Figure 2 represents the simplest circuit, which is capable of performing as indicated in Table 1.

However, by using a somewhat more complex circuit as in Figure 3, it is possible to produce a suitable "front end" for a high performance HF receiver. The results are tabulated in Table 2.

3rd order Intercept point	+ 30 dBm
1 dB compression point	+ 15 dBm
SSB Noise Figure	12 dB
Conversion Gain	– 1 dB
Supply Volts Vcc 1	10V
Supply volts Vcc 2	8V
Programmable current	25 mA
Carrier Input Voltage	220 mV

Table 1. Typical Mixer Performance.

Sensitivity	15 dB SINAD	– 113 dBm input, 3 kHz bandwidth f_{IN}. 1-30 MHz
3rd Order IMD	– 70 dB	2 signals, each – 4 dBm 10 kHz separation.
2nd Order IMD	> – 80 dB	2 signals each – 10 dBm
Carrier Radiation from input port	< – 65 dBm	Measured in 50 ohms at input port.
IF rejection	30 dB	Measured at input port
Input matching	22 dB	Return loss, 50 ohms 1-30 MHz
Gain	10 dB	

Table 2. Front-End Mixer Performance.

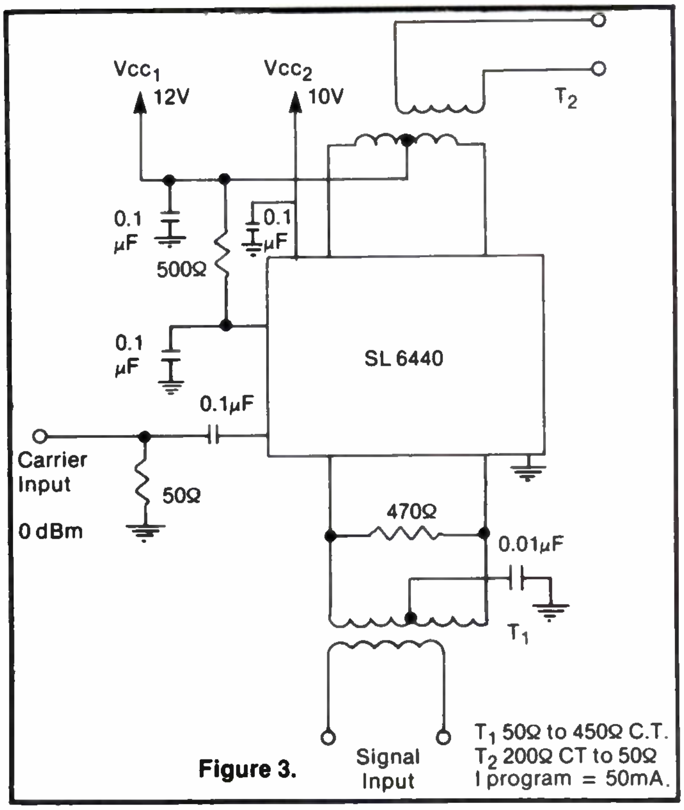

Figure 3.

The achieved dynamic range is quite high: 106 dB. The eventual system performance may well be limited by the performance of any crystal filters following the device. By suitable design of the input transformers, the frequency range may be extended to 100 MHz without significant performance degradation. However, the sensitivity may well be insufficient for use above 30 MHz without the use of an RF amplifier.

The ability of a double balanced mixer such as the SL6440 to produce signals with low IMD ratios may also be used to advantage in SSB transmitters. Because of the high signal levels which can be applied to the IC with negligible distortion, the amount of amplification required at the final frequency can be reduced, while the gain and low noise performance assist in the maintenance of low noise floors.

As a result of the programmable capability, it is feasible to use the SL6440 or similar mixer for other functions as well. This has the effect at simplifying manufacturing and maintenance logistics.

For the very highest performance in the military temperature range, the use of a heat sink is obligatory but for many purposes, this is not required.

Performance Measurement

Difficulties are encountered in measuring the performance, especially IMD of double-balanced mixers such as the SL6440. Merely combining two signal generators via resistive combining networks generally leads to the generators intermodulation products being larger than those introduced by the device under test. Therefore, the combination of the signals is critical, and usually requires a directional coupler, 180° hybrid or similar device. A typical test set-up is shown in Figure 4. The attenuator following the coupler provides a good match to the coupler, maintaining the isolation between generators. Where the signals are close together in frequency, as in receiver measurements, it may well happen that the noise floor of the signal generators may be high enough to cause problems. it is therefore necessary to ensure that the signal generators used are adequate in this respect.

Further, a high output level is required from the generators to allow for losses in couplers and attenuators. When measurements are made with a spectrum analyzer, it is often necessary to ensure that there is sufficient attenuation prior to the spectrum analyzer mixer. This ensures that the analyzer is operating linearly. In this respect, a good test is to change the RF attenuation prior to the analyzer by 10 dB. If the intermodulation products on the analyzer change by other than 10 dB, then the analyzer is being overloaded. If these precautions are followed, then accurate and repeatable IMD measurements can be made.

The SL 6440 offers a new level of performance at a price much lower than that hitherto attainable. In addition, the capability of covering the military temperature range, the low levels of oscillator radiation and the large dynamic range attainable offer the RF designer a new component in an area where design difficulties have previously limited performance.

Appendix

Definition Of Mixer Terms

Mixers consist of two input ports and one output port. One input port is for the RF signal (f_R) and one is for the local oscillator (f_L) signal. The mixed signal ($f_R + f_L$) appears at the output port.

Mixers may be unbalanced, where both RF signal, local oscillator and the IF signal appear at the output; single balanced, where either local oscillator or RF signal are

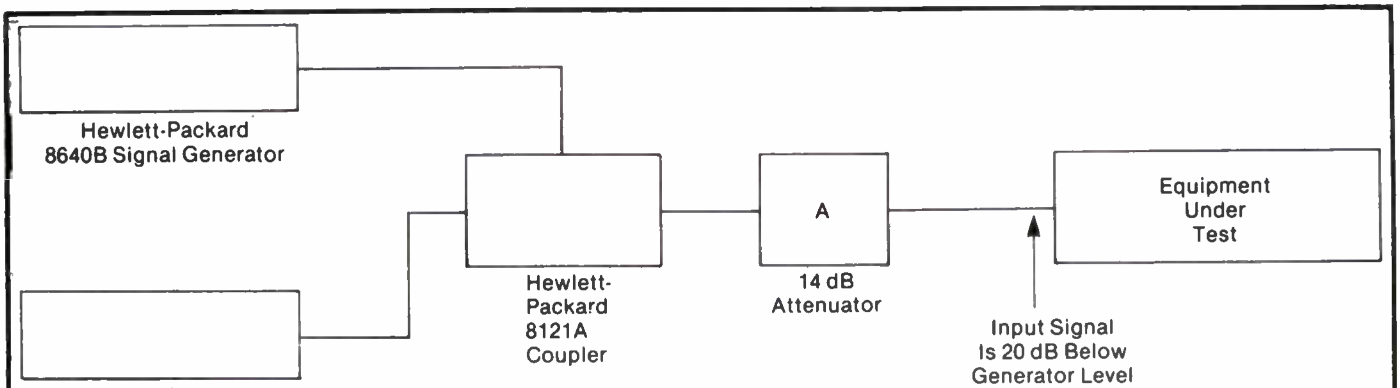

Figure 4. IMD Measurement System.

balanced out and do not appear at the output; double balanced, where both the RF signal and local oscillator are balanced out and do not appear at the output.

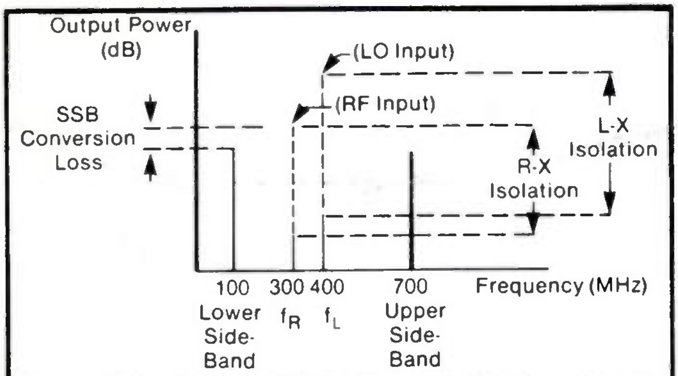

Figure 5. Typical Output Spectrum of Double-Balanced Mixer.

Conversion Gain

Conversion loss is a measure of the gain between the RF input signal and the output IF signal. For a given frequency translation two outputs are produced, a lower sideband and an upper sideband signal. Since only one sideband is generally of value, the specification is usually for a single sideband output. Conversion gain of a mixer is equal to the ratio of the IF single sideband output to the RF input level. All measurements are usually based on a 50 ohm system but this is not always the case.

Conversion Compression

Conversion compression is a measure of the maximum RF signal for which the mixer provides linear operation. Normally the IF output signal is a constant ratio of the RF input level i.e. the conversion gain. However, as the RF level is increased there will be a point where the conversion gain will fall. The IF output level does not exactly follow the increase in RF input level. The criteria used to describe the deviation from linearity between the RF input level and the IF output level is a fixed amount of compression. The 1 dB compression point is the level normally specified.

Isolation

Isolation is a measure of circuit balance within the mixer. When the isolation is high the amount of 'leakage' or 'feedthrough' between the mixer ports will be small. The LO to RF isolation is the amount the LO drive level is attenuated when it is measured at the RF port. The LO to IF isolation is the amount the LO drive level attenuates when it is measured at the IF port. Often only the LO isolation is specified since the RF signal is usually lower than the LO drive level and therefore, is not usually a problem.

Spurious Response (Harmonic Intermodulation)

A mixer due to its non-linearity will generate the entire spectrum of harmonics, which includes harmonics of the LO and RF frequencies, and high order intermodulation products of the form $n\,f_L \pm mf_R$ where n and m are integers. The double balanced mixer, by virtue of its symmetry, suppresses those spurious outputs which are the result of even-order components of the LO or RF signals.

Harmonics of F_{rf}		F_{LO}	$2F_{LO}$	$3F_{LO}$	$4F_{LO}$	$5F_{LO}$	$6F_{LO}$	$7F_{LO}$	$8F_{LO}$
$7F_{rf}$	99	95	100	90	95	99	100	85	95
$5F_{rf}$	97	90	95	90	97	95	100	90	100
$5F_{rf}$	85	80	85	70	90	67	92	65	85
$4F_{rf}$	85	80	95	85	95	85	95	85	90
$3F_{rf}$	70	60	65	70	70	55	75	55	75
$2F_{rf}$	60	70	70	65	80	60	73	70	98
F_{rf}	30	0	33	13	35	35	42	35	45
$F_{rf} \pm F_{LO}$		35	40	46	54	54	55	56	55

1. Conditions: RF = − 10 dBm at 100 MHz (Broadband resistive terminations at all ports.)
LO = + 7 dBm
2. The mixing products are referenced to the desired output $f_{LO} + f_{RF}$
3. RF harmonics are referenced to the RF input
4. LO harmonics are referenced to the LO input.

Figure 6. Typical Spurious Response Chart.

Two-Tone Intermodulation

Two tone intermodulation (IM) distortion results from the arrival of a second input signal with the desired signal at the RF port of a mixer. A measure of this distortion provides a representative means of evaluating a mixer's linearity and spurious performance. When two RF signals, f_{R1} and f_{R2} are simultaneously applied to a mixer, they will combine with the local oscillator frequency, LO, producing the normal terms $f_L \pm f_{R1}$ and $f_L \pm f_{R2}$, as well as the undesired intermodulation products $f_L \pm nf_{R1} \pm mf_{R2}$. The first significant products are $f_L \pm 2f_{R1} \pm f_{R2}$ and $f_L \pm 2f_{R2} \pm f_{R1}$ usually referred to as third order IM products. "Third" refers to the three RF factors involved, e.g. $2\,f_{R1} + 1\,f_{R2}$.

The two-tone IM ratio is the ratio of a third order IM product to an IF output level, at a specified power level for the two RF inputs.

When the output products versus RF input power are plotted on a log-log graph the third order IM products have a 3 : 1 slope and the IF outputs have a 1 : 1 slope. The extrapolations of these two slopes meet at the third order intercept point. The intercept point (specified in terms of its RF input power level) is a common means of expressing a mixer's spurious performance. The higher the intercept point, the better the performance.

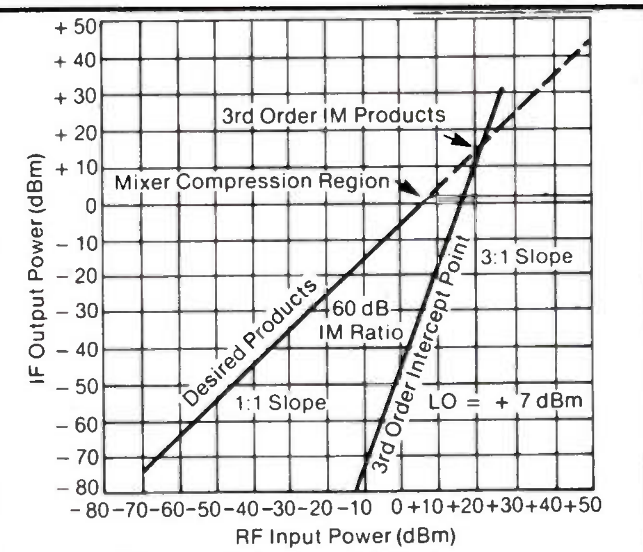

Figure 7. Third-Order IM Performance.

Footnote

1. Such as the Plessey Semiconductors SL6440 I.C. (The only IC mixer with programmable performance).

Line Matching Transformers

Transmission line techniques are developed for matching two real impedances by means of quarterwave *and* non-synchronous transformers. An HP 19/29 program and example are provided.

A. Przedpelski
A.R.F. Products, Inc.
Boulder, CO

The usual means of matching two real impedances, Z_1 and Z_2, is by a section of 90° line (quarter-wave length) of Z_T characteristic impedance (where $Z_T = \sqrt{Z_1 Z_2}$). This applies to coaxial transmission lines as well as printed circuit lines. The main difficulty with this approach is the non-availability of coaxial lines at all desired values of Z_T. This problem does not exist, of course, with printed circuit lines.

Matthaei[1] describes a less known matching configuration, the nonsynchronous transformer, where sections of Z_1 and Z_2 alternate to provide the desired impedance

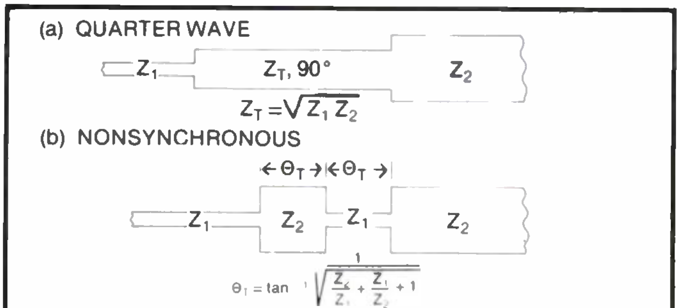

Figure 1. Matching Transformer Design Data.

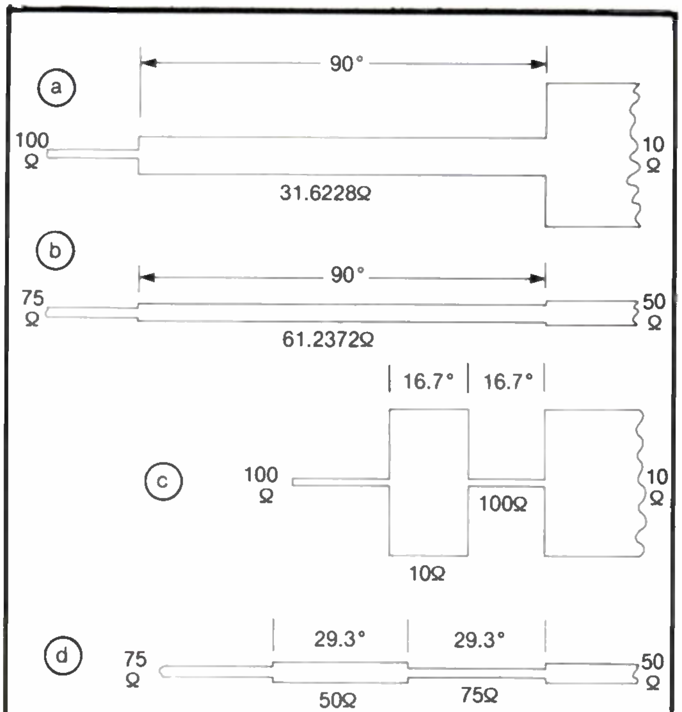

Figure 2. Examples of Matching Transformers.
(a) AND (b) QUARTERWAVE
(c) AND (d) NONSYNCHRONOUS

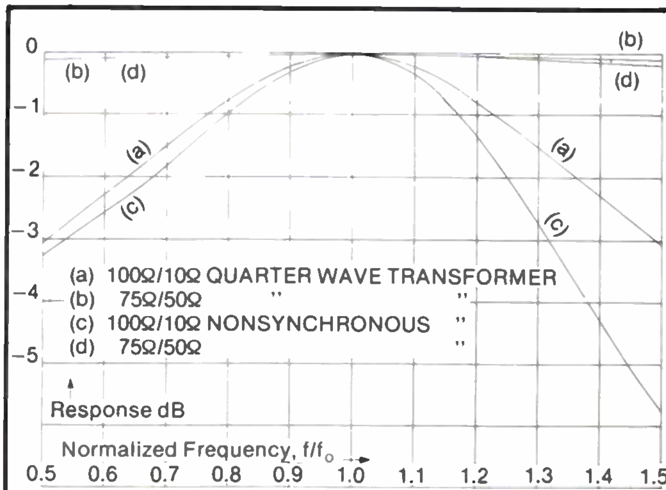

Figure 3. Frequency Response.

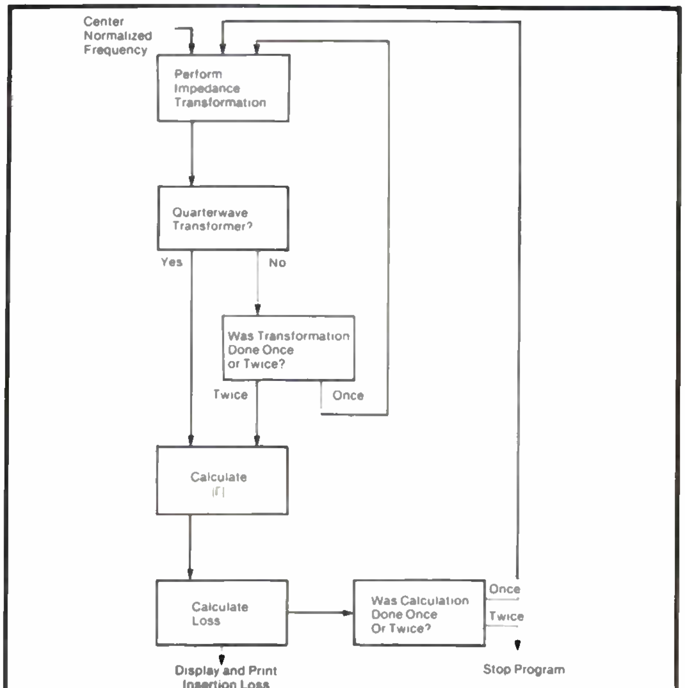

Figure 4. Simplified Calculation Flow Diagram.

References

1. Matthaei, Young and Jones; Microwave Filters, Impedance-Matching Networks, and Coupling Structures; McGraw-Hill Book Company, N.Y., 1964.
2. Handbook of Coaxial Microwave Measurements; General Radio, 1968.

match. The two types of matching transformers are shown in Figure 1. For illustration purposes a match between 100 and 10 ohms and 75 and 50 ohms, using microstrip techniques, is shown in Figure 2. The main advantage of the nonsynchronous transformer, is in coaxial line circuits, since additional impedance lines are not needed. As Matthaei points out, the nonsynchronous transformer is shorter (the maximum transformer length is 60° versus 90° for the quarterwave), especially for large Z_1/Z_2 ratios.

He also states that the bandwidth of the nonsynchronous transformer is smaller, but does not mention that the shrinkage occurs mainly at the higher frequencies. The actual frequency response of either type of transformer can be easily calculated using the program in Table 1. This program is for an HP-19C RPN-type calculator. It can also be used on an HP-29C by changing PRx to R/S and using appropriate key codes.

Step	Instructions	Input Data/Units	Keys		Output Data/Units
1	Enter Program				
2	Store	90	STO	1	
		Z_T	STO	2	
		Θ_T	STO	3	
		Z_1	STO	4	
		Z_2	STO	5	
3	Input Normalized Frequency	f/f_o	GSB	5	f/f_o
				DB_{go}	DB_{NS}
					DB_{90}
4	Repeat Step 3 For Other Frequencies	f_n/f_o	GSB	5	f/f_o
					DB_{NS}
					DB_{90}
	Note: Do Not Use 1.000 for f/f_o In Steps 3 or 4: Use 9.999 or 1.001				

Table 1.

Step	Key Entry	Key Code
001	(g) LBL 5	25 14 05
	PR_x	65
	STO.0	45 .0
	4	04
	STO 0	45 00
	(g) LBL 0	25 14 00
	RCL.0	55 .0
	(g) DSZ	25 45
	RCL i	55 12
010	X	51
	(f) tan	16 44
	STO 8	45 08
	RCL 5	55 05
	X	51
	(g) ISZ	25 55
	RCL i	55 12
	(g)→P	25 34
	STO.5	45 .5
	R↓	12
020	STO.4	45 .4
	RCL 8	55 08
	RCL i	55 12
	X	51
	RCL 5	55 05
	(g)→P	25 34
	RCL.5	55 .5
	÷	61
	RCL i	55 12
	X	51
030	X ≷ y	11
	RCL.4	55 .4
	−	31
	X ≷ y	11
	(f)→R	16 34
	STO 6	45 06
	X ≷ y	11
	STO 7	45 07
	(g) DSZ	25 45
	GTO j	14 12
040	(g) LBL 3	25 14 03
	RCL 8	55 08
	RCL 6	55 06
	X	51
	RCL 5	55 05
	RCL 8	55 08
	RCL 7	55 07
	X	51
	−	31
	(g)→P	25 34
050	STO.5	45 .5
	R↓	12
	STO.4	45 .4
	RCL 8	55 08
	RCL 5	55 08
	RCL 5	55 05
	X	51
	RCL 7	55 07
	+	41
	RCL 6	55 06
	(g)→P	25 34
060	RCL.5	55 .5
	÷	61
	RCL 5	55 05
	X	51
	X ≷ y	11
	RCL.4	55 .4
	−	31
	X ≷ y	11
	(f)→R	16 34
	STO 6	45 06
070	X ≷ y	11
	STO 7	45 07
	(g) LBL 1	25 14 01
	X ≷ y	11
	RCL 4	55 04
	4	04
	X	51
	X	51
	RCL 6	55 06
	RCL 4	55 04
080	+	41
	(g) X^2	25 53
	RCL 7	55 07
	(g) X^2	25 53
	+	41
	÷	61
	(f) log	16 33
	1	01
	0	00
	X	51
090	PRx	65
	(g) DSZ	25 45
	GTO 0	14 00
	(g) SPC	25 65
	(g) RTN	25 13

1	2	3	4	5
90	Z_T	Θ_T	Z_1	Z_2

Program Steps

The program is based on the following steps:

Step 1

Using

$$\frac{Z_{IN}}{Z_T} = \frac{\dfrac{Z_{OUT}}{Z_T} + j \tan \Theta_T}{1 + j \dfrac{Z_{OUT}}{Z_T} \tan \Theta_T} \qquad (1)$$

convert Z_2 to Z_{IN} for the quarterwave transformer or to an intermediate Z_3 for the nonsynchronous type.

Step 2

Skip for quarterwave transformer. For synchronous type repeat Step 1 and convert Z_3 to Z_{IN} using equation (1).

Step 3

Calculate magnitude of reflection coefficient, $|\Gamma|$, using

$$|\Gamma| = \left| \frac{Z_{IN} - Z_1}{Z_{IN} + Z_1} \right| \qquad (2)$$

Step 4

Calculate ratio of transmitted power to incident power using

$$\frac{P_t}{P_i} = 1 - |\Gamma|^2 \qquad (3)$$

Step 5

Calculate response in dB, using

$$10 \log \frac{P_t}{P_i} \qquad (4)$$

Steps 1 through 5 are repeated at the desired normalized frequencies, f/f_o, by changing Θ_T in equation (1) to

$$\Theta_T' = \Theta_T \frac{f}{f_o} \qquad (5)$$

The above procedure is rather tedious and lengthy. The program in Table 1 uses the indirect control feature of the HP-19C/29C (indirect address, increment R_0, decrement R_0) and reduces the total program length by repeating the transformation calculation for the nonsynchronous transformer and repeating the entire program for the quarterwave transformer. Thus, for each normalized frequency input, the attenuation of both transformers is given. A simplified flow diagram is shown in Figure 4.

To demonstrate the use of this program, the responses of the four cases shown in Figure 2 were calculated. The results were plotted in Figure 3. The expected results were obtained: the larger the difference between the impedances, the poorer the frequency response; and, the nonsynchronous transformer has a poorer frequency response, especially at the higher frequencies, as compared to the quarterwave. ☐

AGC Loop Design Using Control System Theory

Normal and delayed AGC loop design theory is presented along with a 75 dB AGC range receiver application.

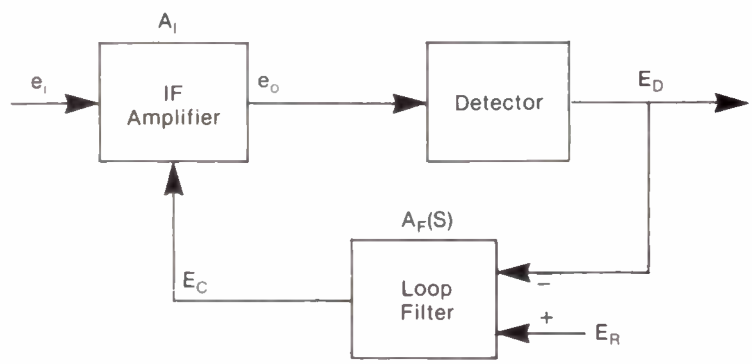

Figure 1. AGC Loop.

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AGC loops, such as the one in Figure 1, are simple automatic control systems, but most explanations of their operation make them appear complex.[1 2 3 4] The apparent complexity is caused by the fact that the change in output voltage resulting from an input voltage change isn't proportional to it, but to the log of the relative input voltage change. That is, the number of millivolts the output changes is proportional to the number of dB the input changes, not to the number of millivolts. This can be shown quite easily.

In Figure 1, the IF amplifier output voltage is related to the input voltage by the amplifier voltage gain.

$$e_o = A_I e_i \tag{1}$$

In a closed loop system all of the variables in (1) are functionally related to the control voltage E_C. Differentiating (1) with respect to E_C,

$$\frac{de_o}{dE_c} = A_I \frac{de_i}{dE_c} + e_i \frac{dA_I}{dE_c} \tag{2}$$

Dividing (2) by (1),

$$\frac{\left(\frac{de_o}{e_o}\right)}{dE_c} = \frac{\left(\frac{de_i}{e_i}\right)}{dE_c} + \frac{\left(\frac{dA_I}{A_I}\right)}{dE_c}$$

or, using the relationship $d\,(ln\,U) = \frac{du}{u}$

$$\frac{d\,(ln\,e_o)}{dE_c} = \frac{d\,(ln\,e_i)}{dE_c} \quad \frac{d\,(ln\,A_I)}{dE_c} \tag{3}$$

Next, define the detector gain constant

$$K_D = \frac{dE_D}{d\,(ln\,e_o)} \tag{4}$$

and the IF amplifier control gain constant

$$K_n = \frac{d\,(ln\,A_I)}{dE_c} \tag{5}$$

Since lnA_I is by definition the IF amplifier gain in nepers, K_n has the dimensions nepers/volt. Similarly, K_D is in volts/neper.

Substituing (4) and (5) in (3),

$$\frac{dE_D}{K_D\,dE_c} = \frac{d\,(ln\,e_i)}{dE_c} + K_n \tag{6}$$

Assuming that the IF amplifier bandwidth, the detector bandwidth and the IF amplifier control circuit bandwidth are all much greater than the AGC loop bandwidth (which is usually the case), K_D and K_n are not functions of frequency. The loop filter gain $A_F(S)$ and the closed loop gain M(S) are:

$$A_F(S) = \frac{E_c}{E_R - E_D} = -\frac{dE_c}{dE_D} \quad (7)$$

$$M(S) = \frac{dE_D}{d(\ln e_i)} \quad (8)$$

E_R is a reference voltage which is adjusted to set the detector DC output voltage E_D.

Substituting (7) and (8) in (6),

$$-\frac{1}{K_D A_F(S)} = -\frac{1}{M(S)\,A_F(S)} + K_n\text{, or}$$

$$M(S) = \frac{K_D}{1 + K_D K_n A_F(S)} \quad (9)$$

If the value of K_n changes with control voltage the AGC loop bandwidth will change with input signal level. In most amplifiers K_n is approximately constant and the variations in loop bandwidth remain within tolerable limits. If the variations are too great, linearizing circuits similar to those found in some VCO's can be used.

K_n is usually evaluated by plotting attenuation in dB versus control voltage for the type of IF amplifier to be used and then finding the slope of the line at various points, or by tabulating the data and determining the value by numerical differentiation. In these calculations it's convenient to express K_n as

$$K_n = K_c K_I \quad (10)$$

where K_I is the control voltage sensitivity in dB/volt and K_c is a conversion factor.

$$K_c = \frac{\ln 10}{20} = .11513\ \text{nepers/dB} \quad (11)$$

In order to keep the loop bandwidth constant with varying values of E_D, a log amp between the detector and the loop filter would be required. However, this is rarely necessary, since the purpose of the AGC loop is to keep E_D constant. E_D is set to a predetermined level by adjusting E_R and K_D is calculated for this value of E_D.

$$K_D = \frac{dE_D}{d(\ln e_o)} = e_o\frac{dE_D}{de_o} \quad (12)$$

For a linear diode detector the DC output voltage is $E_D = ke_o - E_d$, where E_d is the diode voltage drop. Then

$$\frac{dE_D}{de_o} = k \text{ and}$$

$$K_D = ke_o = E_D + E_d \quad (13a)$$

For a square law detector,

$$E_D = ke_o^2,\ \frac{dE_D}{de_o} = 2\,ke_o \text{ and} \quad (13b)$$

$$K_D = 2k\,e_o^2 = 2E_D$$

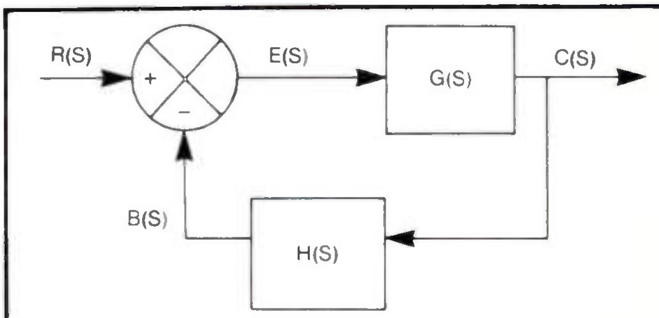


Figure 2. Standard Terminology Closed Loop.

Equation (9) can be written using the standard closed loop transfer function notation[5] shown in Figure 2,

$$M(S) = \frac{C(S)}{R(S)} = \frac{G(S)}{1 + G(S)\,H(S)}\text{, where}$$

$$G(S) = K_D,\ H(S) = K_c K_I A_F(S)$$

and the loop transfer function is

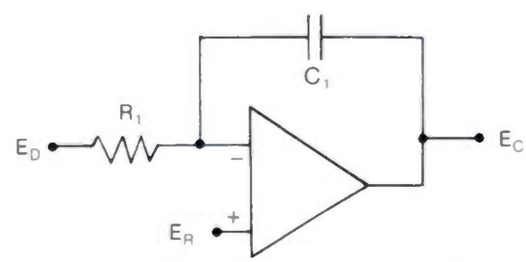
$$\frac{B(S)}{E(S)} = G(S)\,H(S) = K_D K_C K_I A_F(S) \quad (14)$$

The loop error transfer function

$$\frac{E(S)}{R(S)} = \frac{1}{1 + K_D K_C K_I A_F(S)} \quad (15)$$

with a unit step function input is used in evaluating the transient response. The attack time of the AGC loop is the time required for the resulting error e(t) to fall to a specified value, usually .37 or .10.

The most common type of loop filter is the simple integrator shown in Figure 3.



$$A_F(S) = -\frac{E_C}{E_D} = \frac{A_1}{S} \qquad A_1 = \frac{1}{R_1 C_1}$$

Figure 3. Simple Integrator.

Substituting

$$A_F(S) = \frac{A_I}{S} \text{ in (15),}$$

$$\frac{E(S)}{R(S)} = \frac{1}{1 + \dfrac{K_D K_c K_I A_1}{S}} = \frac{S}{S + K_V} \quad (16)$$

where the velocity error constant is

$$K_V = K_D K_c K_I A_1$$

To calculate the transient response we assume a unit (1 neper) step function input r(t) = u(t), then R(S) = 1/S and from (16)

$$E(S) = \frac{1}{S + K_V} \tag{17}$$

The loop error as a function of time is found by evaluating the inverse Laplace Transform of (17),

$$e(t) = \varepsilon^{-K_V t} \tag{18}$$

Ohlson[4] shows that for an amplitude-modulated input signal with modulation index m, the detector output is

$$E_D = E_R \left[\frac{1 + m \sin \omega t}{1 + \beta m \sin(\omega t + \Theta)} \right] \tag{19}$$

where

$$\beta = \frac{1}{\sqrt{1 + \left(\frac{\omega}{K_V}\right)^2}} \tag{20a}$$

$$\text{and } \Theta = -\tan^{-1}\left(\frac{\omega}{K_V}\right) \tag{20b}$$

When the loop filter is a simple integrator.

The quantitites β and Θ are easily identified as the magnitude and phase angle of the unity feedback closed loop gain,

$$\beta = \left| \frac{B(S)}{R(S)} \right|_{S = j\omega} \tag{21a}$$

and

$$\Theta = \arg \left[\frac{B(S)}{R(S)} \right]_{S = j\omega} \tag{21b}$$

From equation (19) it can be seen that at low frequency ($\beta = 1$) the detector output is a DC level $E_D = E_R$ and at high frequency ($\beta = 0$) the output is the undistorted modulation envelope $E_D = E_R(1 + m \sin \omega t)$. The distortion which the denominator of (19) describes is dependent on the modulation index. We arbitrarily assume that in the worst case (m = 1) the demodulated output signal distortion is tolerable for $\beta < .2$ and thus define a low-frequency cutoff by solving

$$\beta^2 = \left| \frac{B(j\omega_c)}{R(j\omega_c)} \right|^2 = \frac{1}{26} \tag{22}$$

Then

$$\frac{K_V^2}{K_V^2 + \omega^2} = \frac{1}{26} \qquad \text{or}$$

$$\omega_c = 5K_V \tag{23}$$

Since the loop gain will have a similar effect on the modulation with any type of loop filter, equation (2) will also be used to define the cutoff frequency when the filter is more complex than a simple integrator.

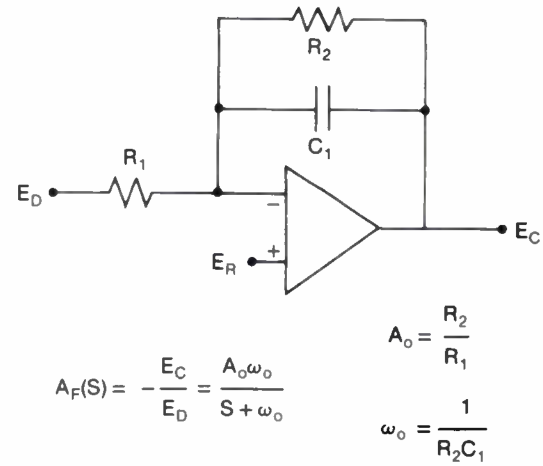


Figure 4. Finite DC Gain Filter.

For a filter with finite DC gain such as the one shown in Figure 4,

$$G(S)\,H(S) = K_D K_C K_I A_o \left(\frac{\omega_o}{S + \omega_o} \right) \tag{24}$$

Substituting the DC loop gain or positional error constant

$$K_p = K_D K_c K_I A_o \qquad \text{in (24)}$$

$$G(S)\,H(S) = \frac{K_p \omega_o}{S + \omega_o} = \frac{N(S)}{D(S)}$$

Then

$$\frac{E(S)}{R(S)} = \frac{D(S)}{N(S) + D(S)} = \frac{S + \omega_o}{S + (1 + K_p)\,\omega_o}$$

For a unit step function input,

$$E(S) = \frac{S + \omega_o}{S[S + (1 + K_p)\,\omega_o]} \tag{25}$$

The inverse transform of (25) can be found in Reference 6, transform pair 116:

$$e(t) = \frac{1}{1 + K_p} + \frac{K_p \varepsilon^{-(1 + K_p)\,\omega_o t}}{1 + K_p} \tag{26}$$

The unity gain closed loop transfer function is

$$\frac{B(S)}{R(S)} = \frac{N(S)}{N(S) + D(S)} = \frac{K_p \omega_o}{S + (1 + K_p)\,\omega_o} \tag{27}$$

Substituting (27) in (22) and solving for ω_c,

$$\omega_c = \omega_o \sqrt{25K_p^2 - 2K_p - 1} \tag{28}$$

For $K_p >> 1$ this becomes $\omega_c \approx 5\,K_p\,\omega_o$,

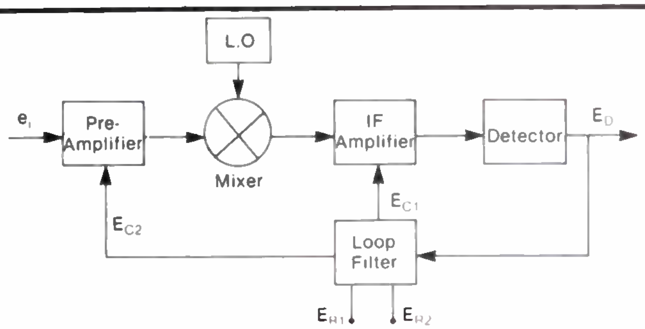
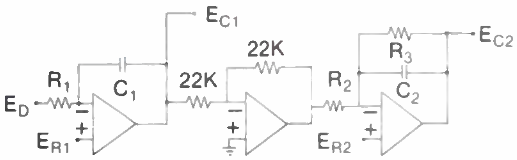


Figure 5. Receiver Block Diagram.



$$A_1 = \frac{1}{R_1C_1} \qquad A_2 = \frac{R_3}{R_2} \qquad \omega_1 = \frac{1}{R_3C_2}$$

$$A_{F1}(S) = -\frac{E_{C1}}{E_D} = \frac{A_1}{S}$$

$$A_{F2}(S) = -\frac{E_{C2}}{E_D} = \frac{A_1}{S}\left[\frac{A_2\omega_1}{S+\omega_1}\right]$$

Figure 6. Loop Filter.

The design becomes somewhat more complicated if delayed AGC is used in addition to normal AGC. For operation over a wide dynamic range, gain control should be applied to the input stages to prevent strong input signals from overloading the following stages. However, applying AGC to the input stages almost always degrades receiver noise figure. This makes it desirable to control only stages near the output at threshold and to "delay" control of the input stages until the input signal is so strong that output signal-to-noise ratio is satisfactory even with an increased receiver noise figure. Figure 5 is a block diagram of such a receiver. The loop filter is shown in Figure 6. E_{R1} sets the output level and the input signal level at which the DAGC cuts in is set by E_{R2}.

A loop filter of this type produces the transfer functions and control system block diagram shown in Figure 7. The closed loop bandwidth is set by the AGC loop parameters and, when ω_1 is properly chosen, the bandwidth changes very little when the DAGC threshold is reached.

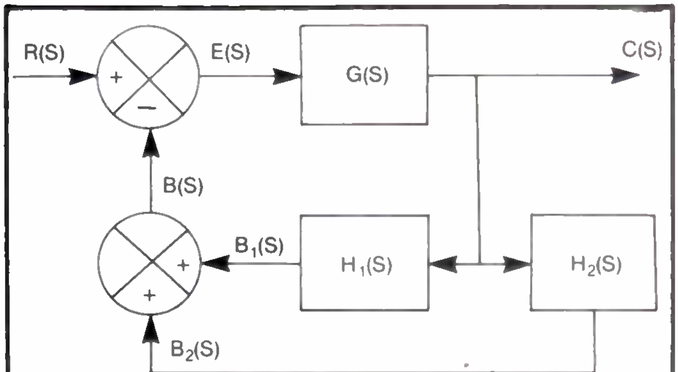


Figure 7. AGC and DAGC Closed Loop.

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2. W.K. Victor and M.H. Brockman, "The Application of Linear Servo Theory to the Design of AGC Loops", Proc. IRE, vol. 48, pp. 234-238, Feb. 1960.
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4. J. Ohlson, "Exact Dynamics of Automatic Gain Control", IEEE Trans. on Communications, vol. COM-22, pp. 72-75, January 1974.
5. B.C. Kuo, "Automatic Control Systems", 2nd. ed. Prentice-Hall 1967.
6. P.A. McCollum and B.F. Brown, "Laplace Transform Tables and Theorems", Holt, Rinehart and Winston 1965.

In Figure 7, $G(S) = K_D$, $H_1(S) = K_c K_{I1} A_{F1}(S)$ and $H_2(S) = K_c K_{I2} A_{F2}(S)$, where K_{I1} and K_{I2} are the control voltage sensitivities of the amplifiers controlled by the AGC and DAGC voltages respectively, and $A_{F1}(S)$ and $A_{F2}(S)$ are shown in Figure 6.

The loop transfer function is

$$\frac{B(S)}{E(S)} = G(S)\,[H_1(S) + H_2(S)],$$

or

$$\frac{B(S)}{E(S)} = \frac{K_D K_c K_I A_1}{S} + \frac{K_D K_c K_{I2} A_1 A_2 \omega_1}{S(S+\omega_1)} = \frac{K_D K_c K_{I1} A_1}{S} \times \left[1 + \frac{\left(\frac{K_{I2}}{K_{I1}}\right) A_2\omega_1}{S+\omega_1}\right]$$

Define $K_1 = K_D K_c K_{I1} A_1$ (29a)

$$\text{and } K_2 = A_2\left(\frac{K_{I2}}{K_{I1}}\right) \qquad (29b)$$

$$\text{Then } \frac{B(S)}{E(S)} = \frac{K_1}{S}\left(1 + \frac{K_2\omega_1}{S+\omega_1}\right) =$$

$$\frac{K_1 S + K_1(1+K_2)\omega_1}{S^2 + \omega_1 S} = \frac{N(S)}{D(S)} \qquad (30)$$

The loop error transfer function is

$$\frac{E(S)}{R(S)} = \frac{D(S)}{N(S)+D(S)} = \frac{S(S+\omega_1)}{S^2 + (K_1+\omega_1)S + K_1(1+K_2)\omega_1} \qquad (31)$$

and the unity-feedback closed loop transfer function is

$$\frac{B(S)}{R(S)} = \frac{N(S)}{N(S)+D(S)} = \frac{K_1 S + K_1(1+K_2)\omega_1}{S^2 + (K_1+\omega_1)S + K_1(1+K_2)\omega_1} \qquad (32)$$

The transient response of the loop is found by substituting $R(S) = 1/S$ in (31). Then

```
100 DIM T (3)
110 T (1) = 1
120 T (2) = 2
130 T (3) = 5
140 P2 = 6.283185
150 N1 = 20
160 PRINT "AGC LOOP TRANSIENT RESPONSE"
170 PRINT
180 PRINT "K1, K2, W1";
190 INPUT K1, K2, W1
200 IF K1 = 0 THEN 730
210 IF W1 < = 0 THEN 230
220 IF K2>0 THEN 270
230 F2 = 5*K1/P2
240 T2 = K1
250 IO = 4
260 GO TO 390
270 A1 = (K1 + W1)/2
280 W6 = K1*(1 + K2)*W1
290 Z2 = A1*A1/W6
300 Z1 = SQR (Z2)
310 W5 = SQR (W6)
320 T2 = A1
330 Z4 = 1 - Z2
340 I0 = SGN(Z4) + 2
350 B2 = 13*K1*K1 + (1 - 2*Z2)*W6
360 W2 = B2 + SQR (B2*B2 + 25*W6*W6)
370 F2 = SQR (W2)/P2
380 PRINT "ZETA = "; Z1, "WN = "; W5
390 PRINT "FC = "; F2; "HZ"
400 PRINT
410 GO TO 420, 500, 480, 500 ON IO
420 WO = W5*SQR( - Z4)
430 X1 = 1/(2*W0)
440 B1 = A1 - W0
450 C1 = A1 + W0
460 T2 = B1
470 GO TO 500
480 W0 = W5*SQR(Z4)
490 A2 = (W1 - A1)/W0
500 T1 = 5E3/(N1*T2)
510 FOR I1 = - 3 TO 3
520 T2 = 10^11
530 FOR I2 = 1 TO 3
540 TO = T2*T(12)
550 IF TO >T1 THEN 580
560 NEXT I2
570 NEXT I1
580 PRINT "T (MILLISECS)", "ERROR"
590 FOR I1 = 0 TO N1
600 T1 = I1*TO
610 T2 = 1E - 3*T1
620 GO TO 630, 650, 670, 690 ON I0
630 E1 = X1*(W1 - B1)*EXP ( - B1*T2) - (W1-C1)*EXP ( - C1*T2))
640 GO TO 700
650 E1 = (1 + (W1 - A1)*T2)*EXP( - A1*T2)
660 GO TO 700
670 E1 = (COS (WO*T2) + A2*SIN (W0*T2))*EXP( - A1*T2)
680 GO TO 700
690 E1 = EXP ( - K1*T2)
700 PRINT T1, E1
710 NEXT I1
720 GO TO 170
730 END
```

Figure 8. AGC Loop Transient Response Program.

$$E(S) = \frac{S + \omega_1}{S^2 + (K_1 + \omega_1) S + K_1 (1 + K_2)\omega_1} = \qquad (33)$$

$$\frac{S + \omega_1}{S^2 + 2\zeta\omega_n S + \omega_n^2}$$

The undamped natural frequency ω_n and damping factor ζ are

$$\omega_n = \sqrt{K_1 (1 + K_2)\, \omega_1} \qquad (34a)$$

and

and

$$\zeta = \frac{K_1 + \omega_1}{2\omega_n} \qquad (34b)$$

To find the transient response the following auxiliary parameters are defined:

$$a = \zeta\omega_n \qquad (35a)$$

$$b = \omega_n (\zeta - \sqrt{\zeta^2 - 1}) \qquad (35b)$$

$$c = \omega_n (\zeta + \sqrt{\zeta^2 - 1}) \qquad (35c)$$

$$\omega_o = \omega_n \sqrt{1 - \zeta^2} \qquad (35d)$$

The transient response is then given by:

$$e(t) = \left[\cos \omega_o t + \left(\frac{\omega_1 - a}{\omega_o} \right) \sin \omega_o t \right] \varepsilon^{-at} \quad \zeta < 1 \ (36a)$$

$$e(t) = \left[1 + (\omega_1 - a)t \right] \varepsilon^{-at} \quad \zeta = 1 \ (36b)$$

$$e(t) = \frac{1}{c - b} \left[(\omega_1 - b)\, e^{-bt} - (\omega_1 - c)\varepsilon^{-ct} \right] \quad \zeta > 1 \ (36c)$$

For $K_2 = 0$ (or $\omega_1 = 0$) e(t) can be found from equation (18) with $K_v = K_1$.

The computer program in Figure 8 calculates e(t) for a range of values of t.

Substituting (34) and (32) in (22), then solving for ω_c,

$$\omega_c = \sqrt{d + \sqrt{d^2 + 25\omega_n^4}} \qquad (37a)$$

where

$$d = 13\, K_1^2 + (1 - 2\zeta^2)\omega_n^2 \qquad (37b)$$

Design of the AGC loop filter consists of selecting values of K_1 K_2 and ω_1 and then determining the filter components which will result in these values. Below the DAGC threshold the loop is described by equations 16 through 23, thus K_1 is determined by the cut-off frequency or the response time. In most cases it is determined simply as $K_1 = \omega_{c/5}$.

K_2 is determined by the relative gain reductions of the IF amplifier and preamplifier which are required above DAGC threshold. K_2 is the ratio of preamplifier gain reduction in dB to IF amplifier gain reduction in dB produced by a certain change in the control voltage E_{c1}. Since

$$K_I = \frac{dA_I(dB)}{dE_c} \quad \text{and at DC}$$

$$A_2 = \frac{dE_{c2}}{dE_{c1}} \quad \text{from (29b)}$$

$$K_2 = \left(\frac{dE_{c2}}{dE_{c1}}\right) \bullet \left(\frac{dA_{I2}}{dE_{c2}}\right) \bullet \left(\frac{dE_{c1}}{dA_{I1}}\right) \quad \text{or}$$

$$K_2 = \frac{dA_{I2}(dB)}{dA_{I1}(dB)} \tag{38}$$

For given values of K_1 and K_2, ω_1 adjusts the damping factor. In most cases approximately critical damping ($\zeta = 1$) is desirable. For $\zeta = 1$,

$$\omega_1 = K_1 \left[(1 + 2K_2) - \sqrt{(1 + 2K_2)^2 - 1}\right] \tag{39}$$

Example

Assume that a receiver with 75 dB AGC range is required. The IF amplifier gain is to be reduced 25 dB before DAGC threshold is reached. Above AGC threshold the IF amplifier gain must be reduced an additional 20 dB and the preamplifier gain 30 dB. The required cutoff frequency is 250 Hz.

Then $K_1 = 2\pi 250/5 = 314.6$ from (23)
$K_2 = 30/20 = 1.5$ from (38)
$\omega_1 = 314.6\,(4 - \sqrt{16 - 1}) = 39.903$ from (39)

A Bode plot of the loop gain described by these parameters is shown in Figure 9.

Typical values for K_D, K_{I1}, and K_{I2} are:

$K_D = 2.0$ V/neper $\quad K_{I1} = 10$ dB/V $\quad K_{I2} = 5$ dB/V

Using these values, the component values for the filter in Figure 6 can be calculated as follows:

From (29a),

$$A_1 = \frac{K_1}{K_D K_c K_{I1}} = \frac{314.16}{2.0 \times .1153 \times 10} = 136.43$$

Select $C_1 = C_2 = .10\mu F$

$$\text{then } R_1 = \frac{1}{A_1 C_1} = 7.329 \times 10^4 \approx 75K$$

$$A_2 = \left(\frac{K_{I1}}{K_{I2}}\right) K_2 = 3.0 \text{ From (29b)}$$

$$R_3 = \frac{1}{\omega_1 C_2} = 2.506 \times 10^5 \approx 240K$$

$$R_2 = \frac{R_3}{A_2} = 8.353 \times 10^4 \approx 82K$$

Figure 10 is a tabulation of the transient response of this loop, both for above DAGC threshold and below it.

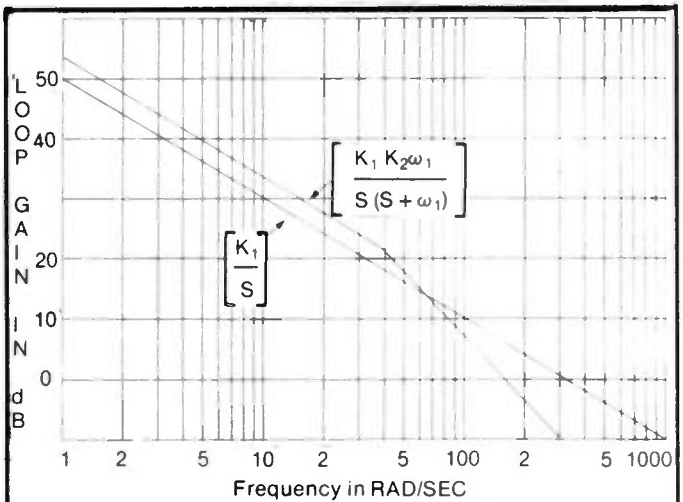


Figure 9. Loop Gain Bode Plot.

K1, K2, W1? 314.16, 1.5, 39.903
ZETA = 1. WN = 177.031
FC = 252.309 HZ

T (MILLISECS)	ERROR
0	1.
2	.509358
4	.222393
6	6.12363E—02
8	− 2.35428E—02
10	− 6.32179E—02
12	− 7.71315E—02
14	− 7.71567E—02
16	− 7.02863E—02
18	− 6.06597E—02
20	− 5.05290E—02
22	− 4.10419E—02
24	− 3.27210E—02
26	− 2.57146E—02
28	− 1.99768E—02
30	− 1.53744E—02
32	− 1.17404E—02
34	− 8.90703E—03
36	− 6.71925E—03
38	− 5.04433E—03
40	− 3.77085E—03

K1, K2, W1? 314.16, 0, 39.903
FC = 250.001 HZ

T (MILLISECS)	ERROR
0	1
1	.730402
2	.533487
3	.38966
4	.284609
5	.207879
6	.151835
7	.110901
8	8.10023E—02
9	5.91643E—02
10	4.32137E—02
11	3.15634E—02
12	2.30540E—02
13	1.68387E—02
14	1.22990E—02
15	8.98321E—03
16	6.56136E—03
17	4.79243E—03
18	3,50040E—03
19	2.55670E—03
20	1.86742E—03

K1, K2, W1? 0, 0, 0

Figure 10. AGC Loop Transient Response.

CRT Application Notes

A series of application notes on the CRT are available from Syntronic Instruments, Inc. of Addison, IL. These series of notes provide basic and specific information for the beginning through seasoned design engineer. It enables the designer to understand the operation, limitations and features of a basic component, the CRT, so integral to many pieces of instrumentation utilized by himself.

The application note titles and contents are:

CRT Resolution Basics (A.P. #3)
- Resolution Defined
- Focus Defined
- Types of Focus
- The Yoke
- Spot Growth

Pincushion Distortion (A.P. #1)
- Geometric Distortion
- Basic Considerations
- Performance Example
- Displacement Error Calculation

Residual Magnetism (A.P. #2)
- General Definition
- Typical Residual Performance
- Operational Effects of Residual Magnetism
- External Factors Affecting Residual Magnetism

Cross Talk and Ringing In CRT Displays (A.P. #4)
- Effect on Parasitic Oscillations
- What Can Be Done?
- Conclusions

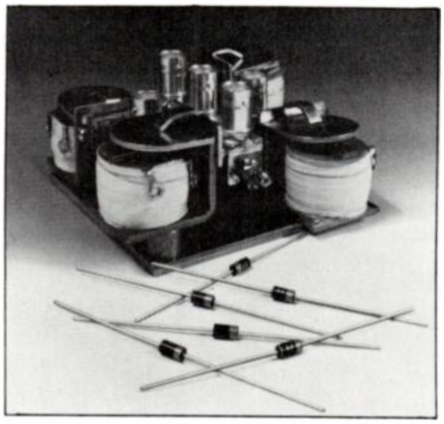
Contact Syntronic Instruments, Inc., 100 Industrial Road, Addison, IL 60101, (312) 543-6444. INFO/CARD #126.

HF Power Pin Diodes

KSW Electronics Corp. has announced a new family of HF Power Pin diodes, KS1001-KS1003, capable of switching 1000 watts of RF power at 2-30 MHz. Diodes have carrier lifetimes of 6 microseconds and distortion more than 80 dB below the fundamental.

Designed for switching HF transmitters, antenna filters and matching networks, the 1200 volt devices can replace relays in most applications. Series resistance is only 0.14 ohms at 750 mA of forward bias current, and diode capacitance is 3 pF at 100 volts of reverse bias. Up to 1 KW is handled by the KS1001 over 2-30 MHz at 1:1 VSWR while the KS1002 handles 300 watts. KS1003 will switch 250 watts over the 10-30 MHz range.

Contact KSW Electronics Corp., South Bedford Street, Burlington, Massachusetts 01803, (617) 273-1730. INFO/CARD #76.



More Oscillators For Less Money

Narda Microwave announces a price reduction on Model 911 VGO Series Oscillators. A Narda design team recently discovered that the 911-Type Varactor Gunn Oscillator was perfect for use in their latest piece of test equipment, the Model 7000A Microwave Multimeter. The 7000A is an extraordinary piece of test gear and Narda anticipates record-breaking sales of this equipment. Because of the increased demand for the oscillator, production of 911's has been intensified and the cost to manufacture the unit has dropped substantially. Besides the dollar savings, 911 VGO Series Oscillators are lightweight, small in size and offer exceptionally low noise performance specifications. For these reasons, they have been used successfully in a multitude of military and commercial applications, most recently as replacements for Klystron local oscillators. A typical frequency specification for the solid-state oscillators offers a tuning range of 1.1 GHz from 8.5 to 9.6 GHz; other 911-Type oscillators are available at operating frequencies from 6 to 18 GHz. Typically, units have a tuning voltage range of +0.5 to −40 volts and nominal tuning sensitivity of 25 MHz/volt. The operating temperature range for these oscillators is from −35°C to +75°C with stability of 0.5 MHz/°C and all units meet or exceed MIL-E-5400 requirements.

Contact: The Narda Microwave Corporation, 75 Commercial Street, Plainview, New York 11803. (516) 349-9600. INFO/CARD #138.

Wide-Frequency Microwave Synthesizer

Watkins-Johnson Company introduces the WJ-1292 Microwave Frequency Synthesizer which covers 10 MHz to 18 GHz in a single 5-1/4-inch-high by 22-inch-deep rack mounting unit. The WJ-1292 provides stable, clean, keyboard-selected signals with 100 kHz resolution. It can be operator-programmed for digital sweeps over any portion of its range. Performance options such as leveling, high-power operation and many others are available by customizing the WJ-1292 with standard building blocks. Coverage to 40 GHz can be achieved through the use of additional RF source heads available from Watkins-Johnson. Also, units providing attenuation, modulation, signal switching and related functions can be combined with the WJ-1292 to produce a custom synthesizer system. Standard computer or IEEE-488 interfaces can be specified. The WJ-1292 Microwave Frequency Synthesizer is specially suited for applications requiring stability, spectral purity and computer programmability. It is ideal for use in automatic microwave test systems, receiver test stations, general microwave laboratory work, metrology labs and for antenna testing. It may also be used as a stable local oscillator for special-purpose receivers and for testing space-communications systems. Its step-sweep capability is useful in making swept measurements of narrow- or broad-band components, subassemblies and systems.

Contact Watkins-Johnson Company, 3333 Hillview Avenue, Palo Alto, CA 94304. INFO/CARD #137.

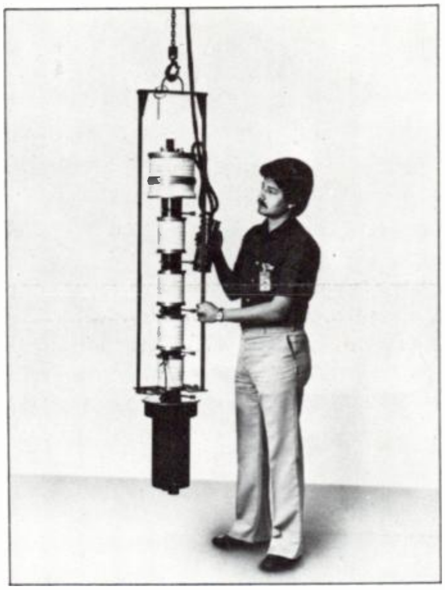
New Amplifier Tubes Cut UHF-TV Transmitter Energy Use 10 Percent

Four new amplifier tubes for UHF-TV transmitters have been announced by the Palo Alto Microwave Tube Division of Varian Associates, Inc. The newest external cavity Klystrons use 10 percent less power, which could save UHF-TV stations at least $4,000/year and up to $100,000 and more, depending on hours of operation, power levels, and number of amplifier tubes.

Savings for a mid-size TV station operating two 55-kw Klystrons 18 hours/day would be $19,000 per year with the new tubes, assuming a 4¢/kwh power cost. In certain areas, where power costs 10¢ to 12¢/kwh, the savings would be much greater. Additional savings would result from reduced heat-exchanger loading and increased tube life due to the lower power levels.

The new Klystrons are directly interchangeable with existing tube models, which are standard in UHF-TV transmitters used throughout the United States and Canada, including those manufactured by General Electric, Ampex, Townsend, and CCA. No modifications to existing hardware are required to install the new tubes.

To save power, the new tubes have been designed so that a minimum gain of 35 dB produces 35 to 58 kw peak-of-sync power outputs with less than 10 watts of RF drive power. The tubes now have a peak-of-sync efficiency of 40 to 42 percent — 8 to



10 percent better than existing models.

Of the four new Klystrons, the models 4KM100LA-H and 4KM100LF-H are intended for operation in the 470 to 596 MHz UHF band, while the models 4KM150LA-H and 4KM150LF-H cover the 596 to 710 MHz UHF band. All models permit multiplexing of video and audio signals with improved linearity over existing models.

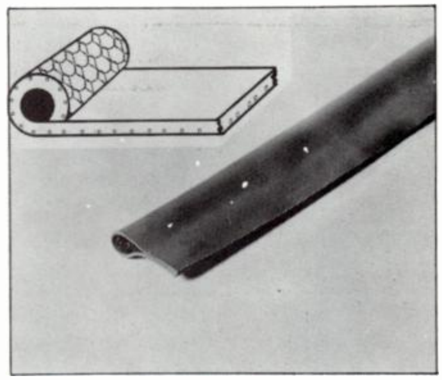
Contact Varian VHF-TV Marketing, 611 Hansen Way, Palo Alto, CA 94303. (415) 493-4000. INFO/CARD #127.

Silicone P-Strips With Knitted Metal Wire Embedded Surface

A new calendared silicone rubber sheeting, Metex PMP I, one of whose surfaces is embedded with knitted metal wires to provide electrical conductivity, is available in a range of special P-strip configurations from Metex Corporation, Edison, N.J. Available wall thicknesses of the material range from .015" to .092" in sheet widths up to 20" wide. Durometers can be varied from 40 to 70, thus providing more effective control. The material works efficiently and effectively under very low compression forces. The new PMP I material was specifically developed for use on Kevlar or carbon fiber filled plastics that are widely employed in the housing of military electronics used on military aircraft, ground support systems, and shelters.

Finished gaskets produced from PMP I maintain the full integrity of the silicone characteristics while simultaneously acting as both an environmental seal and a means of electrical protection. Since the knitted wires are an integral part of the surface, the problems associated with loose wires are completely eliminated. Similarly, there are no powders, no conductive particles. As a result, the life span of the material has increased 10 times over any other equivalent usage gasket.

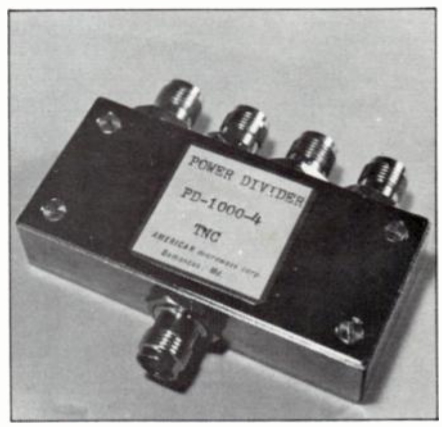
Contact Metex Corporation, Electronic Shielding Group, 970 New Durham Road, Edison, NJ 08817. Circle INFO/CARD #133.



Power Divider Covers 10-1000 MHz

Model PD-1000-4 power divider covers the frequency range 10-1000 MHz. It features low loss, typically 0.6 dB from 10-400 MHz, and 1.25 dB from 400-1000 MHz over split loss. Isolation between output ports is 25 dB, minimum up to 20 MHz, and 30 dB, minimum from 20 to 1000 MHz. Return loss is 10 dB from 10-40 MHz, and 20 dB from 40-1000 MHz on all ports. Impedance is 50 ohms. Amplitude balance between outputs is 0.3 dB, maximum. Available with SMA, TNC or BNC connectors.

Contact American Microwave Corp., P.O. Box 41, Damascus, MD 20750. (301) 253-6782. INFO/CARD #139.



Tantalum Hermetically Sealed Tubular Capacitor

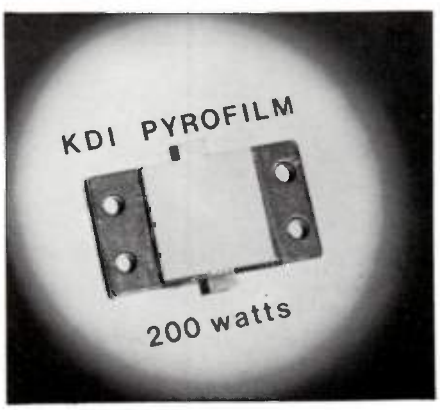
Plessey Capacitors, Westlake Village, California, is now in production of the W15 series "all tantalum hermetically sealed" tubular capacitor. This capacitor is presently available in the popular T3 and T4 case sizes. The standard range for the T3 case size is 86uf 100 VDC to 750 uf 10 VDC. The W15 is manufactured to meet the requirements of Mil-C-39006/22. Qualification to this spec is currently being pursued. The all tantalum units provide large values of capacity in minimum space requirements. It is specifically designed to provide input and output filter needs of static inverters, pulse modulated and switching regulators, and other power supply applications.

Contact Plessey Capacitors, Inc., Tantalum Division, 5334 Sterling Center Drive, Westlake Village, CA 91361. Telephone (213) 889-4120. INFO/CARD #135.

Pyrofilm Attenuator Dissipates 200 Watts

KDI Pyrofilm has added a new power attenuator to its rapidly growing line of conduction cooled power devices. The PPA-200 power attenuator will dissipate 200 Watts when the mounting flange is maintained at 80°C. It is available in attenuation values from 1 to 20 dB. Operating in the frequency range from DC to 500 MHz, the attenuator has a VSWR of 1.25 from DC to 200 MHz and 1.50 from 200 to 500 MHz. This compact attenuator measures 1 inch by 1-7/8 inches.

Contact Al Arfin, KDI Pyrofilm, 60 South Jefferson Road, Whippany, NJ 07981. (201) 887-8100. INFO/CARD #128.



Low-Cost Systems Counter

Extensive Hewlett-Packard Interface Bus (IEEE-488) capability and the reciprocal-taking frequency measurement technique are among the many features provided at a low price in this new counter from Hewlett-Packard. Called the Model 5316A Universal Counter, this instrument measures frequency, frequency burst, frequency ratio, time interval, time interval average and period. It will also totalize. Its two input channels operate over the counter's 100 MHz frequency range. An optional third input channel is available to cover frequencies to 1.0 GHz for communications applications. Combining a wide variety of measurement functions and HP-IB capability, the HP 5316A is ideal for applications where cost is important in design, production and test of electronic and mechanical products, and data logging in scientific research, industrial processes and communications systems.

The Model 5316A uses reciprocal-taking measurements for frequencies below 10 MHz, thus providing eight full digits of resolution (seven digits per second of gate time) over its complete frequency range. This is particularly advantageous in measuring low frequencies rapidly. The counter includes an easy-reading eight-digit LED display (plus exponent), trigger level and sensitivity controls for both channels, and pushbutton measurement selection. The Model 5316A becomes Hewlett-Packard's lowest priced full HP-IB (IEEE-488) programmable systems counter. It can function as both talker and listener on the bus, and all major measurement functions are programmable. Trigger level programmability, usually provided only as an expensive option in programmable counters, is standard in the HP 5316A.

Contact Inquiries Manager, Hewlett-Packard Company, 1507 Page Mill Road, Palo Alto, CA 94304.

RF Absorption Milliwattmeter

This new broad-band Termaline® RF Milliwattmeter terminates and measures the output of low power signal sources directly without the use of charts. A front-panel range-switch selects one of three ranges, 0-200 mW, 800mW and 3 watts, without the need to transfer crystals. The wide fre-

quency range of the model 6257 accommodates communications measurements, all the way from 100 kHz Maritime Mobile/Maritime Radio Navigation to one gigahertz Aeronautical Radio Navigation, and all services in between. The unit is designed to measure output of broad-band oscillators, signal generators, hand-held transceivers or any low-powered device. Used in conjunction with compact TENULINE® Attenuators, the Milliwattmeter's maximum full scale range can be expanded to 25 or 100 watts. Each of the three ranges are field-calibratable (e.g. for tighter accuracy at a specific frequency). The wattmeter's diode detector also serves as a demodulator of AM transmission envelopes. The demodulated (audio) signal is available at a front-panel miniature phone jack to feed into high impedance display or analysis instrumentation. VSWR of model 6257 is below 1.1 to 512 MHz and less than 1.15 to 1000 MHz in 50 ohm coaxial systems.

Contact Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139. (216) 248-1200. INFO/CARD #129.

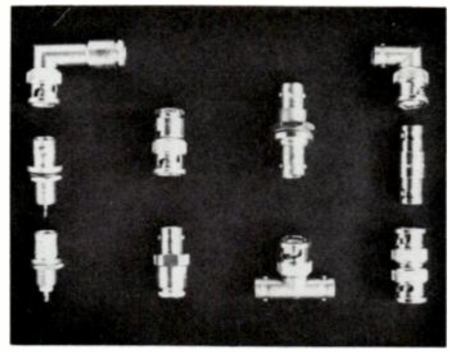
Mil-Spec BNC Connectors And Adapters

A broad line of standard and special Mil-Spec BNC connectors and adapters that now appear on the QPL is available from Delta Electronics Mfg. Corp. of Beverly, Massachusetts.

The Delta BNC Connector and Adapter Series includes straight and right angle plugs; cable, bulkhead, and panel jacks; bulkhead and panel receptacles; and straight adapters that meet Mil-Spec and now appear on the QPL.

The Delta BNC Connector and Adapter Series conforms to Mil-C-39012 and Mil-A-55339. They are part of an extensive connector line including N, TNC, SMA, C, SC, GHV, and HN series. Miniature to LC sizes can be produced, and special connectors can also be provided.

Contact: Delta Electronics Mfg. Corp., 93 Park Street, Beverly, MA 01915, (617) 927-1060. INFO/CARD #141.



Direct Reading Ground Tester

A new direct reading ground tester is now available from AEMC Corporation. The Earth Tester Model 2 offers direct ground resistance readings, easy portability and ruggedness. The direct reading Earth Tester Model 2 is the ideal instrument for all electrical utilities, REA's, municipals, electrical contractors and inspectors, and all maintenance crews checking that their ground resistances are in accordance with their regulations or with the NEC, OSHA and IEEE regulations. The direct reading Earth Tester is a

portable unit built into a metal case and lid, with a padded shoulder strap for carrying it in the field. It is supplied by 3 C cells, fuse protected against potential differences between

The HP 8620 Sweeps the Spectrum–

10 MHz to 22 GHz.

A versatile mainframe.

With the 8620C Mainframe, you get the sweep modes and markers you need for both wide and narrow band measurements: a full band sweep with 3 markers, a marker sweep at the touch of a button and a △F sweep that can be as wide as 100% of band. Plus precise frequency setability with the convenient CW vernier and △F "expand" controls: you can set a 1 MHz △F Sweep even at 18 GHz. 8620C Mainframe, $2850.

Wide RF coverage –and high power.

Choose from ultra wideband RF plug-ins—10 MHz to 2.4 GHz with the HP 86222, 2 to 18.6 GHz (optional to 22 GHz) with the HP 86290B—or from octave and double-octave plug-ins, several of which offer 40 mW or more output power. These RF units also provide the excellent frequency accuracy, linearity, and spectral purity that make them ideal sources for general purpose bench and field test applications. They're especially useful in swept frequency test systems such as the HP 8410 and 8755 Network Analysis Systems.

Popular plug-ins include:

Model No.	Freq. Range	Output Power	Price
86222A/B	10 MHz-2.4 GHz	20 mW	$4100/4800
86235A	1.7-4.3 GHz	40 mW	$3500
86240A	2.0-8.4 GHz	40 mW	$4150
86240C	3.6-8.6 GHz	40 mW	$4950
86245A	5.9-12.4 GHz	50 mW	$4600
86260A	12.4-18 GHz	10 mW	$3950
86290A	2-18 GHz	5 mW	$14,250
86290B	2-18.6 GHz	10 mW	$16,250
86290B opt. H08	2-22 GHz	2 mW	$19,250

HP-IB programmability option adds value.

DESIGNED FOR HP-IB SYSTEMS

HP-IB: Not just IEEE-488, but the hardware, documentation and support that delivers the shortest path to a measurement system.

When you add the HP-IB option ($950) to the 8620C Mainframe, you can program up to 10,000 frequency points per band on up to 4 bands with a variety of sweep modes. With the precision 86222 or 86290 RF units installed in the 8620C, excellent repeatability is achieved thanks to the exceptional frequency accuracy and linearity of these plug-ins.

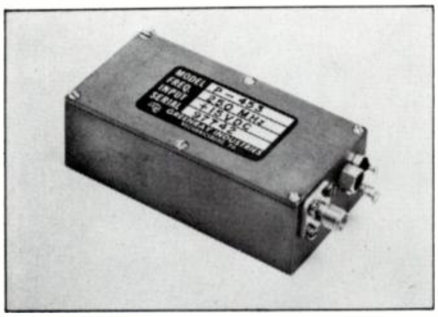
A real advantage of HP-IB is you get not only the bus architecture you need, but the documentation support that can help you get your system up and running in weeks, instead of months.

A prime example is the scores of HP-IB automatic microwave network analyzer systems now in use. Major elements of these systems are the 8620C/86290B Sweeper, HP 8410 Network Analyzer, and HP 9825A Desktop Computer. Systems like this are fully described in HP Application Notes 221 and 187 Series. You can get copies of these plus information on the 8620C from your local HP sales office, or write Hewlett-Packard, 1507 Page Mill Rd., Palo Alto, CA 94304.

Domestic U.S. prices only.

45007A

HEWLETT PACKARD



is 3.0"L x 1.5"W x 1.0" High.

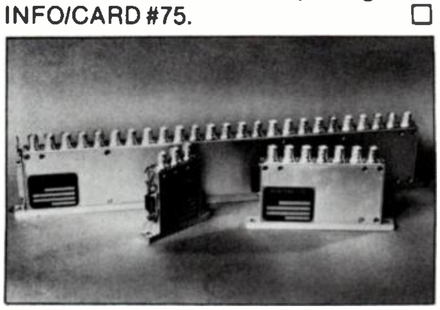
Contact Greenway Industries, Inc., 840 West Church Road, Mechanicsburg, PA 17055. (717) 766-0223. Circle INFO/CARD #140.

New Coaxial-Line Switching Arrays

R.F. Components believe that they have a world lead in the design and production of coaxial switches for handling data and RF transmission applications, by simultaneously switching both the inner conductor and outer sheathing. This U.K. firm, based at Hoddesdon, north of London, has had 15 years' experience in the design of stackable multipoint switchable arrays, used mainly in the past for data and speech communications. With the advent of high-speed data handling the importance of switching screens as well as central conductors became apparent, and many of the arrays in the R.F. Components' catalogue can now be ordered with this facility. Specials can also be considered, if the size of the order should justify such special development and production.

Dual reed switches, operated by a single coil, connect for example one output with one of several inputs, or — thanks to the stackability of the basic blocks — can give a crossbar switching effect from a 10 by 10 array of coaxial lines. Coil operating voltages can be ordered in the range 4 to 80 VDC. Various coaxial connectors can be supplied for the inputs and outputs, including B.N.C.

Contact: R.F. Components, Plumpton House, Plumpton Rd., Hoddesdon, Hertfordshire EN110EB, England. INFO/CARD #75. □



Advertiser Index

Classifieds

Food for thought.

Our new Universal Tone Encoder lends it's versatility to all tastes. The menu includes all CTCSS, as well as Burst Tones, Touch Tones, and Test Tones. No counter or test equipment required to set frequency-just dial it in. While traveling, use it on your Amateur transceiver to access tone operated systems, or in your service van to check out your customers repeaters; also, as a piece of test equipment to modulate your Service Monitor or signal generator. It can even operate off an internal nine volt battery, and is available for one day delivery, backed by our one year warranty.

- All tones in Group A and Group B are included.
- Output level flat to within 1.5db over entire range selected.
- Separate level adjust pots and output connections for each tone Group.
- Immune to RF
- Powered by 6-30vdc, unregulated at 8 ma.
- Low impedance, low distortion, adjustable sinewave output, 5v peak-to-peak
- Instant start-up.
- Off position for no tone output.
- Reverse polarity protection built-in.

Group A

67.0 XZ	91.5 ZZ	118.8 2B	156.7 5A
71.9 XA	94.8 ZA	123.0 3Z	162.2 5B
74.4 WA	97.4 ZB	127.3 3A	167.9 6Z
77.0 XB	100.0 1Z	131.8 3B	173.8 6A
79.7 SP	103.5 1A	136.5 4Z	179.9 6B
82.5 YZ	107.2 1B	141.3 4A	186.2 7Z
85.4 YA	110.9 2Z	146.2 4B	192.8 7A
88.5 YB	114.8 2A	151.4 5Z	203.5 M1

- Frequency accuracy, ± .1 Hz maximum - 40°C to + 85°C
- Frequencies to 250 Hz available on special order
- Continuous tone

Group B

TEST-TONES:	TOUCH-TONES:		BURST TONES:			
600	697	1209	1600	1850	2150	2400
1000	770	1336	1650	1900	2200	2450
1500	852	1477	1700	1950	2250	2500
2175	941	1633	1750	2000	2300	2550
2805			1800	2100	2350	

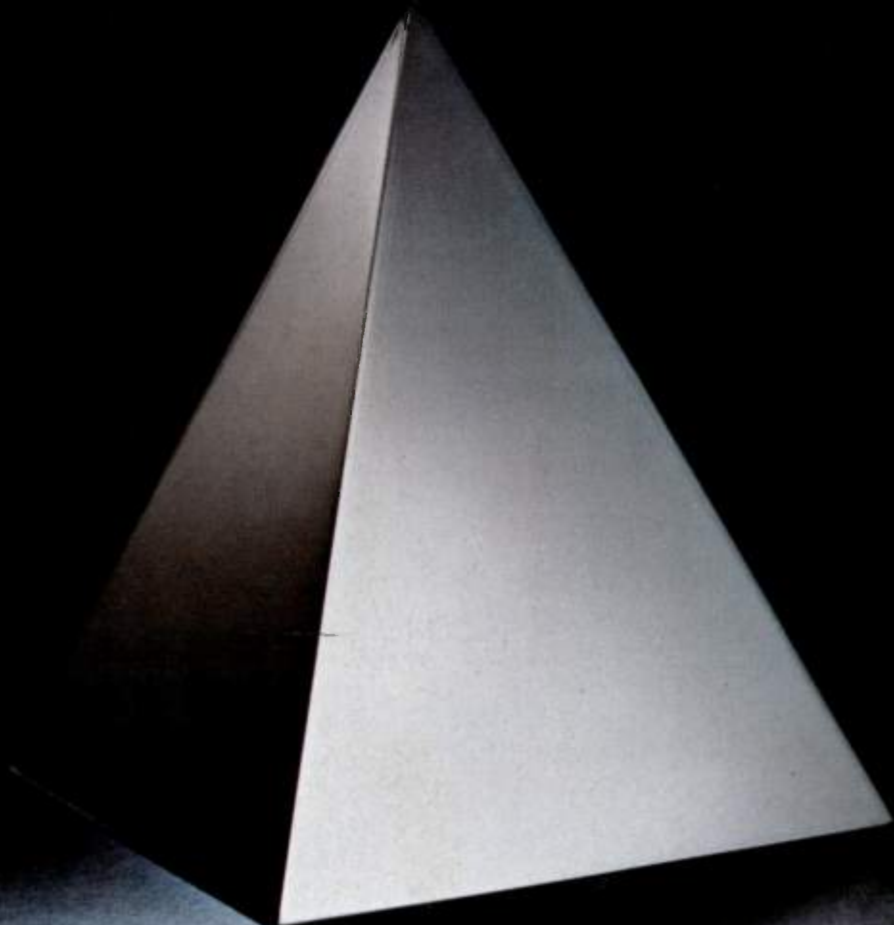
- Frequency accuracy, ± 1 Hz maximum - 40°C to + 85°C
- Tone length approximately 300 ms. May be lengthened, shortened or eliminated by changing value of resistor

Wired and tested: $79.95

COMMUNICATIONS SPECIALISTS
426 West Taft Avenue, Orange, California 92667
(800) 854-0547/ California: (714) 998-3021

INFO/CARD 13

Now...
the only RF power amplifier you may ever need.



The new ENI 550L delivers 50W, 1.5-400 MHz.

This single unit is so incredibly versatile it can replace several you may be using now. And you may never need another. It's an extremely broadband high power, solid state, Class A linear amplifier. It's rated at 50W from 1.5-400 MHz. But it can provide 100 Watts from 1.5-220 MHz. All you need with the 550L is any standard signal or sweep generator and you've got the ultimate in linear power for such applications as RFI/EMI testing, NMR, RF Transmission, ultrasonics and more.

And, like all ENI power amplifiers, the 550L features unconditional stability, instantaneous failsafe provisions, and absolute protection from overloads and transients.

The 550L represents the pinnacle in RF power versatility. There's nothing like it commercially available anywhere! And it may be the only RF power amplifier you ever need.

For more information, a demonstration, or a full line catalog, please contact us at ENI, 3000 Winton Road South, Rochester, NY 14623. Call 716/473-6900, or telex 97-8283 ENI ROC.

ENI

The advanced design line of power amplifiers

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