February 1980

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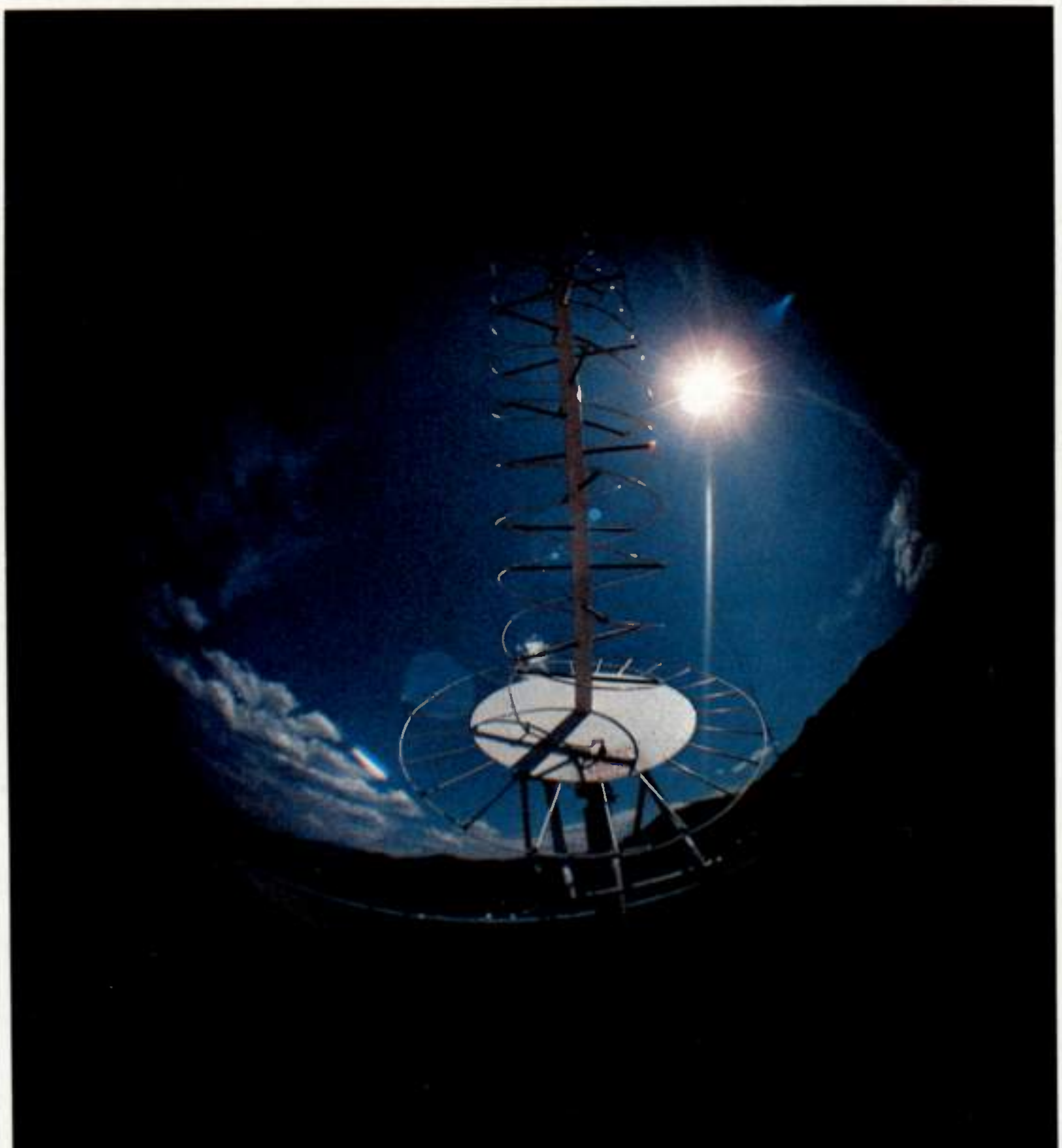
rf design



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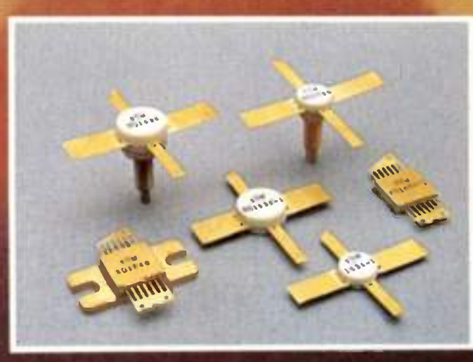

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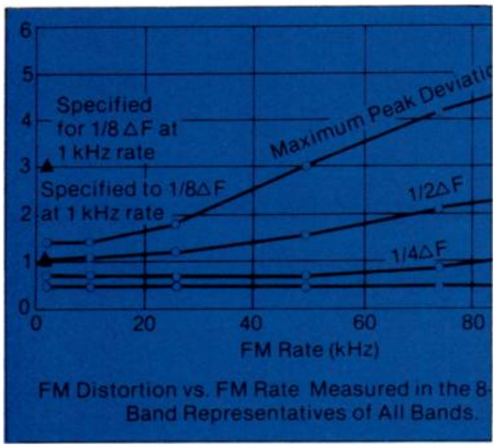

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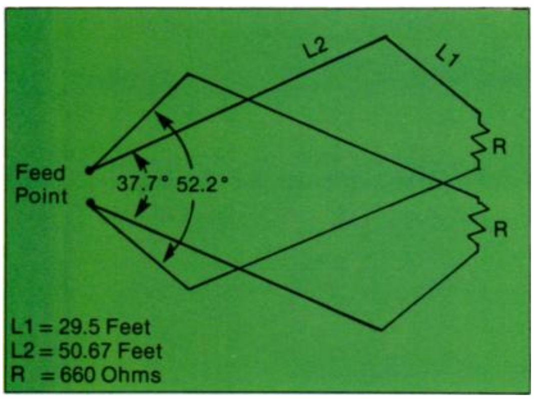

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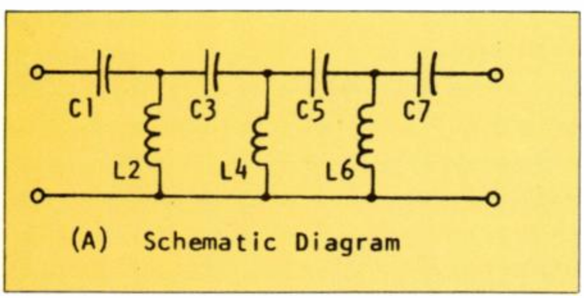

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ZHL-1A	2-500	16 Min.	±1.0 Max.	+28 Min.	11 Typ.	+38 Typ.	+24V	0.6A	199.00	(1.9)
ZHL-2	10-1000	16 Min.	±1.0 Max.	+29 Min.	18 Typ.	+38 Typ.	+24V	0.6A	349.00	(1.9)
ZHL-3A	0.4-150	24 Min.	±1.0 Max.	29.5 Min.	11 Typ.	+38 Typ.	+24V	0.6A	199.00	(1.9)

Total safe input power +20 dBm, operating temperature 0° C to +60° C, storage temperature −55° C to +100° C, 50 ohm impedance, input and output VSWR 2:1 max.
For detailed specs and curves, refer to 1979/80 MicroWaves Product Data Directory, p. 364-365 or EEM p. 2970-2971.

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Meet Rich Rosen

Did you ever wonder what kind of guy becomes the editor of a trade magazine? Well, meet Rich Rosen, the new editor of *r.f. design.* He started more than 20 years ago, as a teenager, on the road to being a first-rate RF engineer when he became interested in Ham radio (K2RR). The first piece of gear he ever built was a 2W-CW transmitter using an old 6AG7 in a haywire lashup only a 15 year old could make work. He now works 10-80 meters and has a penchant for unusual antennas. When he moved here to Denver, one of the big concerns in finding a house was making sure that a 50 foot tower would be allowed.

He says his most interesting job (till now!) was at AIL where he worked on interfacing optical systems with RF systems. The project dealt with simultaneous reception of azimuth and frequency spread signals. He also designed RF gear for a naval architect doing Trident work. He did a stint with the CBS Television Network as an audio/video development engineer.

Rich came to *r.f. design* from Hughes Microwave where he was Manager of Field Engineering. That translates into being in charge of the group that puts up TVRO earth stations for the CATV market.

Rich is an outgoing, friendly, precise engineer who has collected thousands of books and magazines dealing with RF, for which he has computer-compiled a bibliography with thousands of entries. He delights in writing new programs for his HP-67. Somewhere in all this he found and married the beautiful, Chaya. They now live in Littleton, Colorado with two children and a 50 foot tower.

We are glad Rich is now with *r.f. design* and think that you will be too.

Pat Wiesner

E. Patrick Wiesner

Editor,

I have tried without success to utilize the formulas given in the article by F.W. Hauer to be able to use it on my TI-59-calculator but either the formulas are incomplete or perhaps misprinted but I can't arrive at the same answers as Mr. Hauer.

Looking at

$$e_{ff} = \frac{er + 1}{2} + \frac{er - 1}{2}$$

on the beginning of the calculations for either formula for e_{eff}, this portion of the equation will always equal "er". What purpose does it serve?

Puzzled
C.A. Snell
Raytheon Corp. Dept. 9286
6380 Hollister Ave.
Goleta, Calif. 93017

Editor,

Mr. Snell's concern is understandable. The first two terms in these relationships would be redundant if the second term was not multiplied by the third term. The equation is not:

$$e_{eff} = \left(\frac{er + 1}{2} + \frac{er - 1}{2}\right)\left[(1 + 12\,h/\omega)^{-1/2} + .04\,(1 - \omega/h)^2\right]$$

it should be and is:

$$e_{eff} = \left(\frac{er + 1}{2}\right) + \left(\frac{er - 1}{2}\right)\left[(1 + 12\,h/\omega)^{-1/2} + .04(1 - \omega/h)^2\right]$$

The parentheses around each of the first two terms should have been included in the original manuscript for clarity.

Fred Hauer

Reader Domanski is correct in saying that reducing loop bandwidth has certain disadvantages. However, third order loop has many advantages discussed previously, and, in most cases, is necessary in RF applications. With proper design, and this is the purpose of these articles, the loop bandwidth can be made as wide as required. To design very wide bandwidth loops the additional poles have to be accounted for since they are there whether the designer wants them or not. Using the described "fifth order" loop design, PLL's with bandwidths up to 500 kHz were successfully designed and performed as predicted without any component adjustments. Neglecting the additional poles in the calculations (using a pure third order loop design) resulted in circuits which oscillated due to inadequate phase margin.

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

Gentlemen,

There is a problem in the article on pages 32-35 of the current issue of *r.f. design*. The program set forth contains a fatal error. It has 226 steps, while the HP-67 can handle a maximum of 224 steps. (Note that the steps are mis-numbered between steps 160 and 170.)

It would be appreciated if you could come up with a correction to this interesting program.

Albert E. Hayes, Jr.
Albert Hayes & Associates
Consulting Engineer

Editor,

Mr. Hayes is correct, some errors apparently crept in someplace between the manuscript and the published article. Starting with step number 168 the program reads:

g x≤y
1
+
GTO 5

Delete 1 and + so the program is:

Step	Entry Code
168	gx≤y 32 71
169	GTO 5 22 05
170	g LBLfe 32 25 15

Step #200 should read GTO 9 not GTO 0.

On page 33, the equations should read:

$$\frac{\omega e}{h} = \frac{\omega}{h} + \frac{t(\qquad)}{\pi h}$$

ω/h is not part of the numerator of the second term.

Fred Hauer

Fixed Wire Wound Resistors Described in 16-Page Catalog

A concise catalog of fixed wire wound resistors for a variety of electrical power applications is being offered by Tel Labs, Inc. of Londonderry, New Hampshire. The Tel Labs Catalog of fixed wire wound resistors provides complete specifications on the Q81 temperature sensing resistor, epoxy encapsulated SA Series, ceramic core resistor CR/CA/CL Series, silicone coated EL Series, and Type EH chassis mounts.

Fully illustrated, the 16-page Tel Labs Catalog of Fixed Wire Wound Resistors includes graphic and tabular data on values, tolerances, temperature coefficients, and dimensional characteristics for each resistor. Commercial and MIL-SPEC grades are available.

The Tel Labs Catalog of fixed wire wound resistors is available free from Tel Labs, Inc., 154 Harvey Road, Londonderry, N.H. 03053. INFO/CARD #117.

New Wire Cloth Catalog

The Woven Products Division of National-Standard Company has published a new 28-page, two-color catalog on industrial wire cloth. The catalog covers types of weaves, materials and unique characteristics of wire cloth for potential applications, in addition to listing the wide range of wire cloth available in standard and special weaves. The catalog contains a glossary of wire cloth terms, English and metric tables on the physical properties of steel wire, and other helpful data making it useful for designers, specifiers, purchasing agents, production managers and others in a wide range of industries.

Contact National-Standard Company, Woven Products Division, Industrial Blvd., Corbin, Ky. 40701. Circle INFO/CARD #116.

Connector Study

The Electronic Connector Study Group has published the second hardbound volume of "The Connectors and Interconnection Handbook."

Volume 2 — Connector Types, contains over 330 pages fully illustrated with photographs, charts and graphs. Major sections cover connector selection, materials, cable/rack and panel connectors, rigid printed wiring connectors, flat cable/flexible circuit connectors and component sockets and a terms and definitions section including Fiber Optics. Updates and revisions to Volume 1's Specifications and Standards section and Symposia Title Word Index are also included with a comprehensive listing of interconnection related Technical Periodicals, Technical Societies and Connector and Socket Vendors.

Volume 1 — Basic Technology, containing over 240 fully illustrated pages is now in its second printing and covers Interconnection Technology, Connectors, Terminations and Materials.

Each volume is available from the ECSG for $25.00 each in the U.S.A. ($30.00 each outside of the U.S.). A check or P.O. payable to ECSG (Publications), should be mailed to P.O. Box 167, Fort Washington, Pa. 19034. INFO/CARD #112.

Coaxial and Waveguide Terminations

Narda Microwave Corporation has announced a new, half-watt SMA termination, model 4380, covering DC to 26.5 GHz. Also featured are two new series of medium power SMA terminations with exceptionally low VSWR specifications to 18 GHz. Thermal design of these units is otpimized for minimum temperature rise for the unit's size. Also, maximum power ratings are based upon continuous operation at the rated power for an indefinite period.

Contact Narda Microwave Corporation, 75 Commercial Street, Plainview, NY. 11803. INFO/CARD #113.

Temperature Compensated Crystal Oscillators

A detailed brochure that describes a line of temperature compensated crystal oscillator modules is being offered by Microsonics of Weymouth, Massachusetts.

The Microsonics Brochure of temperature compensated crystal oscillator modules, series 80 provides complete specifications on TCXO modules in commercial and MIL-SPEC grades with frequencies from 250 kHz to 60 MHz and ten temperature stability options.

Contact Microsonics, 60 Winter Street, Weymouth, Mass. 02188. Circle INFO/CARD #115.

Opportunities for Engineers

Opportunities for experienced engineers to work in such state-of-the-art GTE technologies as fiber optics, digital integrated circuits, highest complexity MOS memory, bubble memory, codecs and many others are outlined in a brochure from GTE Automatic Electric Laboratories Incorporated.

Entitled "Career Opportunities in Telecommunications," the brochure describes the excitement and rewards that come with operating at the leading edge of the state-of-the-art in applying new technology effectively to large-scale systems.

Contact GTE Automatic Electric Laboratories Incorporated, 400 North Wolf Road, Northlake, Ill. 60164. Circle INFO/CARD #114.

Free Catalog From Rental Electronics

Rental Electronics, Inc. announced the publication of their new Rental Catalog, which lists the $30+-million worth of electronic equipment in the company's rental inventory in the U.S. and Canada. REI maintains 11 inventory centers in the U.S., three in Canada, and a separate used equipment sales office.

The colorful 36-page catalog includes descriptions and monthly rental charges for a variety of items in the following instrument categories: amplifiers; wave and distortion analyzers; logic analyzers; data link analyzers; network analyzers; power line analyzers; sound and vibration analyzers; spectrum analyzers; attenuators; calibrators; cameras; counters; couplers; desktop computers; filters; function generators; noise generators; pulse generators; signal source generators; swept frequency generators; generators/synthesizers; analog voltage, current and resistance meters; digital meters; power meters and thermistors; microprocessor instrumentation; modulators; and more.

Contact Rental Electronics, Inc., 19527 Business Center Drive, Northridge, California 91324. INFO/CARD #134. □

Signal Generator Specifications — Part II

By Rob Oeflein
Marketing Engineer
Hewlett-Packard

Output Level

To be useful, the signal generator *output level* must be calibrated in some unit over a specified range into a characteristic impedance. The nominal output impedance of most RF signal generators is 50 ohms. For CATV and some telecommunications applications which require 75 ohms, an impedance transformer or resistive adaptor can be used. High output levels are desirable for driving mixers or overcoming system losses. Low output levels are important for receiver sensitivity measurements and to calibrate RFI measurements.

The most common output level unit is dB referenced to one milliwatt (dBm) into a 50 ohm resistive load. In many applications it is desirable to set the output directly in volts and this capability is normally provided. Several other output units have become popular in the receiver test market. These include E.M.F. or open circuit voltage, dB referenced to one microvolt (dBuV), and dB referenced to one femtowatt (dBf).

Amplitude resolution determines the settability of the output level. For relative output settings high resolution is beneficial and .1 dB may be desired. In an automatic system one-half the resolution is the maximum amplitude correction which can be made, but it should be stressed that this resolution does not determine the absolute accuracy of the setting. Figure 8 illustrates the sources of absolute accuracy error.

Temperature, level flatness, indicator accuracy, detector linearity, attenuator accuracy, and measurement error all contribute to the signal generator's *absolute output level accuracy* specification. The accuracy specification (usually given as ± dB) should include allowances for all of these factors if an absolute value is indicated. The importance of this specification is evident in receiver usable sensitivity measurements where an error of a few dB might cause a misrepresentation of many miles in the radio's reception capability. Also, if the signal generator is used to verify ALC characteristics or squelch thresholds the output level accuracy is the key specification involved.

Temperature primarily affects the sensitivity of the detector. This error can be minimized by designing the signal generator with a large thermal mass and by maintaining a relatively constant ambient temperature. *Level flatness* is a measure of the flatness of the insertion loss of both the detector and the attenuator as well as a measure of the signal generator's ALC capability. Level flatness must be specified over the frequency range of interest. *Indicator accuracy* and *detector linearity* refer to the tracking between the actual output and the level that the detector sees and the indicator displays. The indicator is commonly a meter, but may now be a digital readout.

At a single frequency these errors (temperature, flatness, detector linearity, and indicator accuracy) can be calibrated out using a power meter and a fixed vernier setting on the signal generator. Reference 1 describes this technique.

Attenuator accuracy for step attenuators is directly proportional to the accuracy of each pad involved in setting the output level. The total error is cumulative so that if a generator has 5 pads and each is specified at ±0.3 dB, the worst case error would be ±1.5 dB. Some synthesized signal generators use their internal microprocessor to calibrate out this error. Waveguide beyond cutoff attenuators are still used in some generators because of their inherent accuracy and ability to provide a continuously variable output. The drawback is the inability to program them and their large insertion loss.

Measurement error refers to the uncertainty of the absolute error which results from the signal generator calibration. The primary source of this error is the impedance mismatch or VSWR of the signal generator and the instrument used to calibrate it. The VSWR of a signal generator is normally given for both high and low levels since the VSWR is degraded at high output levels when the attenuator is not being utilized.

When the signal generator is used in an application which requires low level signals, like measuring the sensitivity of pocket pagers, *leakage* can cause serious inaccuracies. This specification is a measure of the signal generator's EMI and RFI. It is normally sufficient to indicate whether the signal generator complies with one or more of the universally accepted standards such as MIL STANDARD 461 or VDE 0871. A qualitative and easily measured specification indicates the voltage induced in a two-turn 2.4 cm. loop held 2.4 cm. from any surface on the generator. This method provides an easily measured value for comparison among signal generators.

Reverse power protection with automatic reset (or an in-line fuse) helps prevent serious damage to the generator's attenuator or circuitry in the case of an accidental transmission from a transceiver. This type of protection should specify the reverse power handling capability and indicate any degradation of the level accuracy caused by the circuits involved.

Modulation

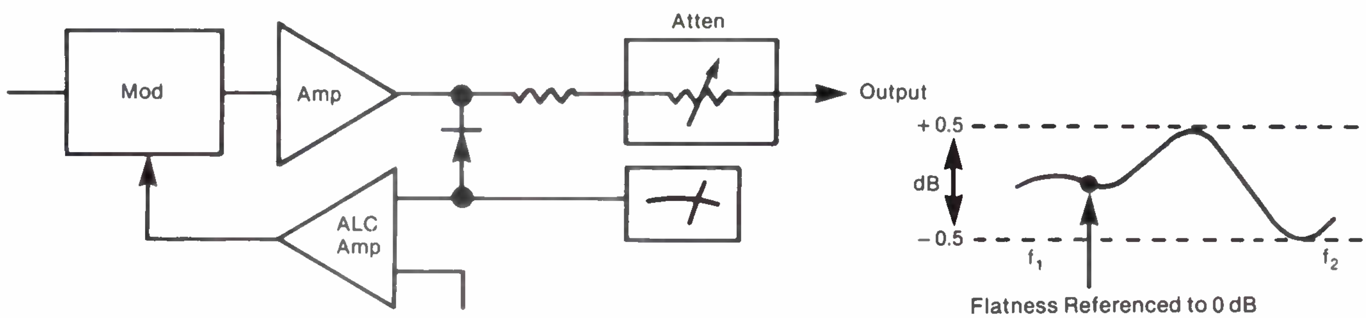
Without the ability to pass information modulated onto the carrier, an instrument cannot be classified as a signal generator. There are a variety of modulation formats in use today, but normally each method can be classified as either amplitude or angular modulation. In general, the signal generator's modulation specifications are concerned with its ability to precisely apply the internally or externally supplied information onto the carrier. Sufficient bandwidth and fidelity are the two primary concerns.

Amplitude modulation requires that the signal generator provide for a variable percentage of depth over some specified range of rates. In standard AM receiver tests the amount of depth required may be only 30 percent. However, for VOR/ILS avionics testing and other complex modulation schemes, total depths up to 100 percent AM may be necessary.

AM bandwidth available may depend on the RF carrier frequency as well as the amount of depth selected. Both internal and external rates should be specified as well as whether external DC coupling is available. DC coupling is important not only for low frequency modulation with minimal phase shifts but also for some applications where it is desired to program the output level in a continuous manner.

Figure 8. Sources of absolute output level error.

Output Level Accuracy



Sources of Error: Level Flatness
Level Variations with Temperature (Due to Detector)
Detector and Meter Linearity
Attenuator
Measurement Uncertainty During Calibration

Modern Data Sheets Should Include Total Absolute Accuracy At Any Level Setting and Include All Sources of Error.

Normally 1 kHz and 400 Hz rates are supplied internally. An optional selection of other fixed rates or a continuously variable internal oscillator may be offered.

AM distortion results from the inability of the ALC loop to track the AM envelope. For this reason, the distortion will increase with greater depths and higher rates. A specification for the internal 1 kHz or 400 Hz rates with a single depth can be misleading. Figure 9 illustrates two methods of more completely specifying AM distortion. Obviously, if the fidelity of a receiver or other system is to be measured the limiting factor is the distortion inherent in the signal generator and the audio source.

The *accuracy* of the AM depth depends primarily on the quality of the modulator utilized and the indicator used to display the setting. If a meter is used to indicate the depth it is important to know whether the accuracy is specified as percent of reading or percent of full scale. The error is constant for percent of reading but increases with lower deflections for percent of full scale.

When AM is applied to a carrier a small amount of FM generally appears at the same rate as the modulating frequency. This *incidental FM* in contrast with residual FM is not present on the CW signal. In stringent AM applications such as stereo AM incidental FM adds to the noise and distortion. The actual mechanism involved is commonly incidental phase modulation and in this case the FM deviation increases with the modulation rate. Incidental FM equals the peak incidental øM times the modulation rate.

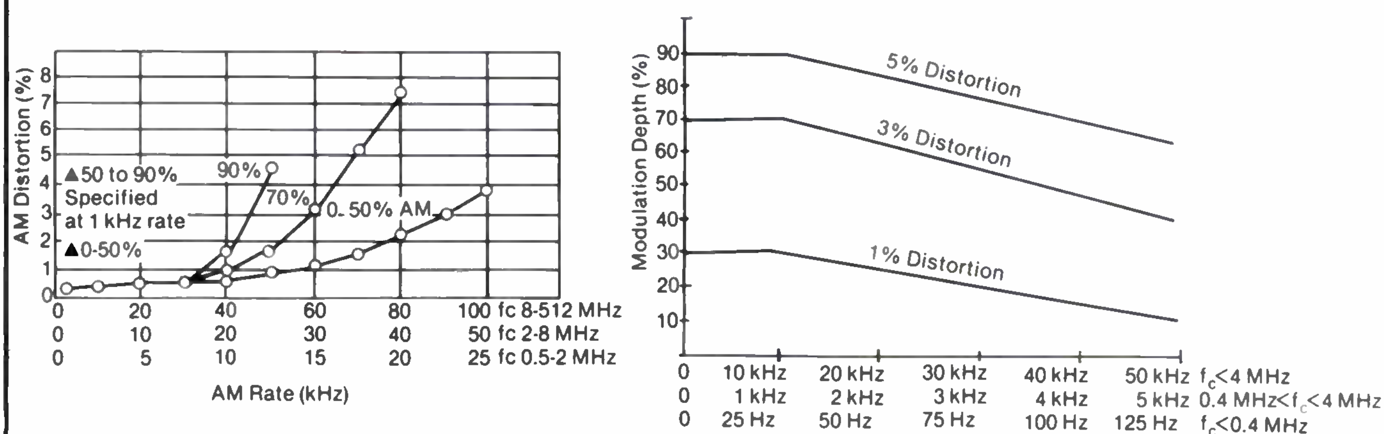
Pulse modulation (Pm) is a form of amplitude modulation. Many of the specifications are similar with the exception of the pulse parameters and the pulse on/off ratio. Fast *rise/fall times* help the input pulse to be faithfully reproduced. Pulse repetition rate and minimum pulse width further identify the type of pulses which may modulate the carrier. The *on/off ratio* defines the power present when the RF is pulsed on to the power present when it is pulsed off. High resolution radar systems may require rise/fall times less than .1μs and on/off ratios of greater than 80 dB at the IF frequencies. Other applications for pulse modulation include avionics DME radars, IF filter characterization, and EW or ECM work.

Frequency modulation requires that the signal generator provide for a variable amount of FM deviation over some specified range of rates. Although most FM mobile receivers have audio bandwidths less than 5 kHz, the entertainment FM band requires from 75 to 100 kHz of peak deviation from 88 to 108 MHz. As mentioned previously, the available peak deviation is preserved in the heterodyne process but reduced by half with each division by two. Thus the amount of available peak deviation may decrease at lower frequencies and it is important to know the specification for the carrier frequency of the application. In addition, the resolution of the FM deviation may vary depending on the carrier frequency.

FM bandwidth, like AM bandwidth, should be specified for both internal and external rates. In older generators a 3 dB bandwidth was common, but today many manufacturers are giving the 1 dB bandwidth. External DC FM is desirable to provide an analog DC sweep capability for testing IF filters or discriminators, and to allow the generator to act as the VCO in a phaselocked loop. DC FM may also be required for certain squelch formats if low rate AC coupling is not provided.

FM distortion will depend on the rate and deviation selected. Figure 10 illustrates two methods of displaying FM distortion. Although 1 percent distortion is normally sufficient for most applications, some FM stereo receivers now specify better than .05 percent distortion. Again, the distortion is commonly specified at the internal rates of 1 kHz and 400 Hz for a particular amount of deviation.

Incidental AM is the small amount of AM which occurs at the same rate as the FM. It is not present



AM Distortion vs. AM rate measured at 200 MHz and + 13 dBm, representative of all bands.

Figure 9. Two methods of specifying AM Distortion on a data sheet for various Rates and Depths.

on the CW signal, but normally becomes more pronounced with greater FM deviation. Incidental AM impairs the ability of the signal generator to make FM rejection measurements on an AM receiver, and can effect Bessel null calibrations.

The specification is normally given in percent AM, but the AM sidebands may be specified in dBc. When the power in both sidebands is measured the dBc specfication may be converted to percent AM by the formula:

$$10^{\frac{-XdBc + 40}{20}}$$

When the power in only a single sideband is indicated (as on a spectrum analyzer) a 6 dB correction factor is necessary since at 100 percent AM each sideband has 25 percent of the power in the carrier. This correction can be made by adding 46 dB instead of 40 dB.

In some synthesizers the carrier will shift in frequency when the generator is switched from the CW to the FM mode. This carrier shift is especially undesirable in narrowband applications and should be specified if the shift occurs.

Phase modulation øm is another form of angular modulation. Here the important specifications are similar to FM with the exception that the deviations are expressed in degrees or radians. Although most øM applications occur above 1 GHz, such as satellite communications, it is desirable for checking the phase characteristics of subassemblies or to analyze phase-lock loops.

Finally, if the signal generator has the ability to perform *simultaneous modulation* it should be specified. Simultaneous AM and FM can be utilized to check the FM rejection of an AM receiver or vice versa, and is required for stereo AM applications. Also, if a demodulated AM output is provided it should be noted. This feature is used to check the accuracy of percent AM settings and is desired primarily in avionics applications.

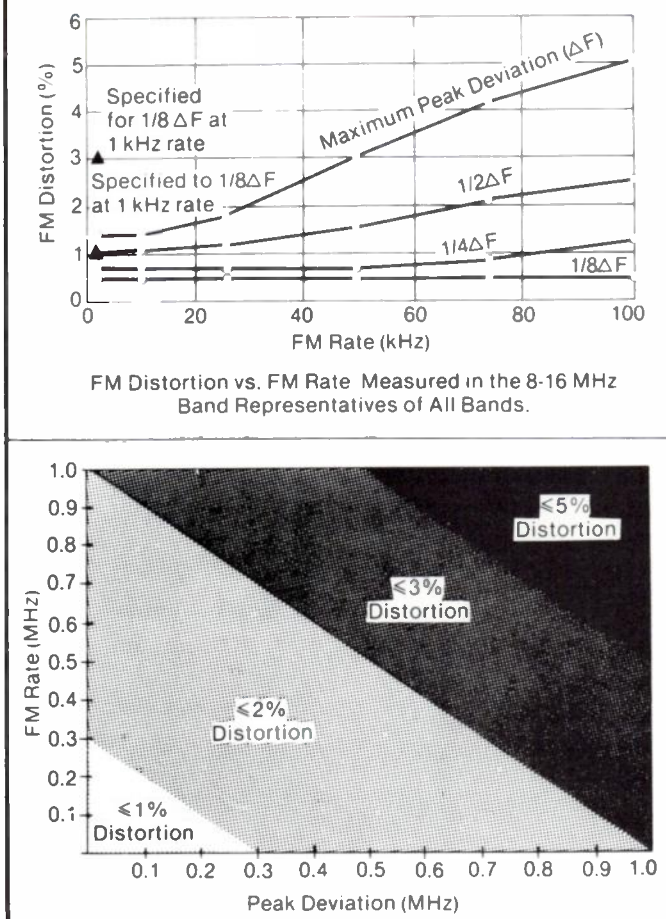


Figure 10. Two methods of illustrating FM Distortion on a data sheet for various Rates and Deviations.

Switching Speed

With the advent of synthesized signal generators, both direct and indirect, *switching speed* has become an important specification to understand. Indirect synthesizers generally switch slower than direct synthesizers due to the indirect synthesis process. Switching speed on an indirect synthesizer is comprised of three factors, control time, lock time, and settling time.

Control time is the time required for the synthesizer to receive the command, process the change information, and send out the appropriate data to the RF section. *Lock time* refers to the time required for the phase locked loops to capture a new frequency after a change is initiated by the controller. The time it takes for the phase error in the loop to decrease to within a specified value from the final frequency is *settling time.*

A direct synthesizer can change frequency essentially as fast as its internal switches. Control time is the limiting factor in their design since they do not utilize any phase-lock loops and their settling times are normally short.

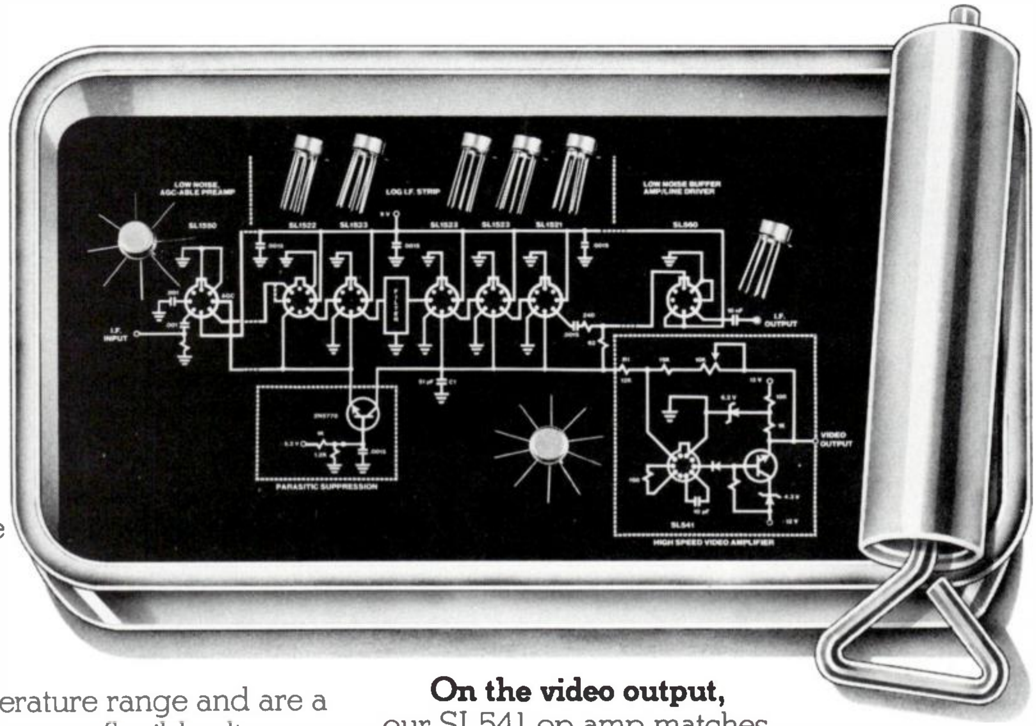
For all signal generators, the switching speed should be specified as the time from the initialization of the change to when the output is within a certain number of hertz from the desired final frequency.

Faster switching synthesizers can also incorporate a precision *digital sweep* capability into their design. The time saving advantage of this feature is readily evident during the analysis of filters, antennas, or amplifiers. With the resolution of 0.1 Hz available in some signal generators the digital point to point nature of the sweep becomes almost unnoticeable.

Control

Programmability in signal generators has led to their widespread use in Automatic Test Equipment (ATE) systems. Several types of interfaces are available today and care should be taken to select the one which offers the most flexibility for the application.

I.F. cram course



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Let's go through the diagram function by function and we'll show you exactly what we mean.

The log IF strip is based on the Plessey SL1521, the simplest, easiest-to-use and least expensive wideband amplifier you can buy. It has a gain of 12 dB and an upper cut-off frequency of 300 MHz (our less expensive SL521 is available if you're working under 100 MHz). The SL1522 is two 1521's in parallel with a resistive divider for increasing the IF strip's dynamic range, while the SL1523 is two 1521's in series. With these devices, it takes just five cans and a single interstage filter to build a log IF strip at 160±15 MHz with a logging range of 90 dB, ±1 dB accuracy, −90 dBm tangential sensitivity and a video rise time of 20 nsec or less. It's reliable, inexpensive and field repairable, which is more than can be said about the hybrids or discretes you're buying now.

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INFO/CARD 6

Binary coded decimal (BCD) interfaces have existed on signal generators for quite a few years. Since this is usually a full parallel interface it offers speed and simultaneous control of several functions. It is important to know the number of lines required by the interface and the decimal weighting of the 4 bit code. Although "8421" is the most commonly used today, "4221" and "1248" have been popular in the past.

The *IEEE 488-75* interface is quickly gaining in popularity. This 16 line bit-parallel, byte-serial interface features standardized interface connectors and a handshake type of control format. This interface bus cannot operate at the speed of a full parallel interface, but it is sufficient for the majority of ATE applications.

Control features have also improved the manual operation of today's signal generators through the use of internal microprocessors. Such features as full keyboard control, store/recall of front panel settings, and the ability to define increments for all functions can greatly reduce test set-up and measurement times.

Conclusion

The challenge of selecting a signal generator which will fulfill all of the critical requirements in a particular application can be simplified if the specifications involved are thoroughly understood. Once it has been determined which of the six specification categories (frequency, spectral purity, output level, modulation, switching speed, and control) are critical to the signal generator's performance, the various instruments available may be more easily evaluated. All future or anticipated requirements should be included in this analysis.

Comparison of the data sheet specifications should then be undertaken carefully. Widespread differences in measurement methods and the existence of "specmanship" can lead to the purchase of equipment unable to meet all the requirements. The contrasting specifications should be converted to units which can be easily compared, and particular attention should be paid to all of the qualifications associated with each specification. For example, a specification given as typical does not guarantee that every unit manufactured will have the stated performance.

Astute examination of the application followed by prudent selection of the signal generator will ensure that the equipment purchased will match the performance required. □

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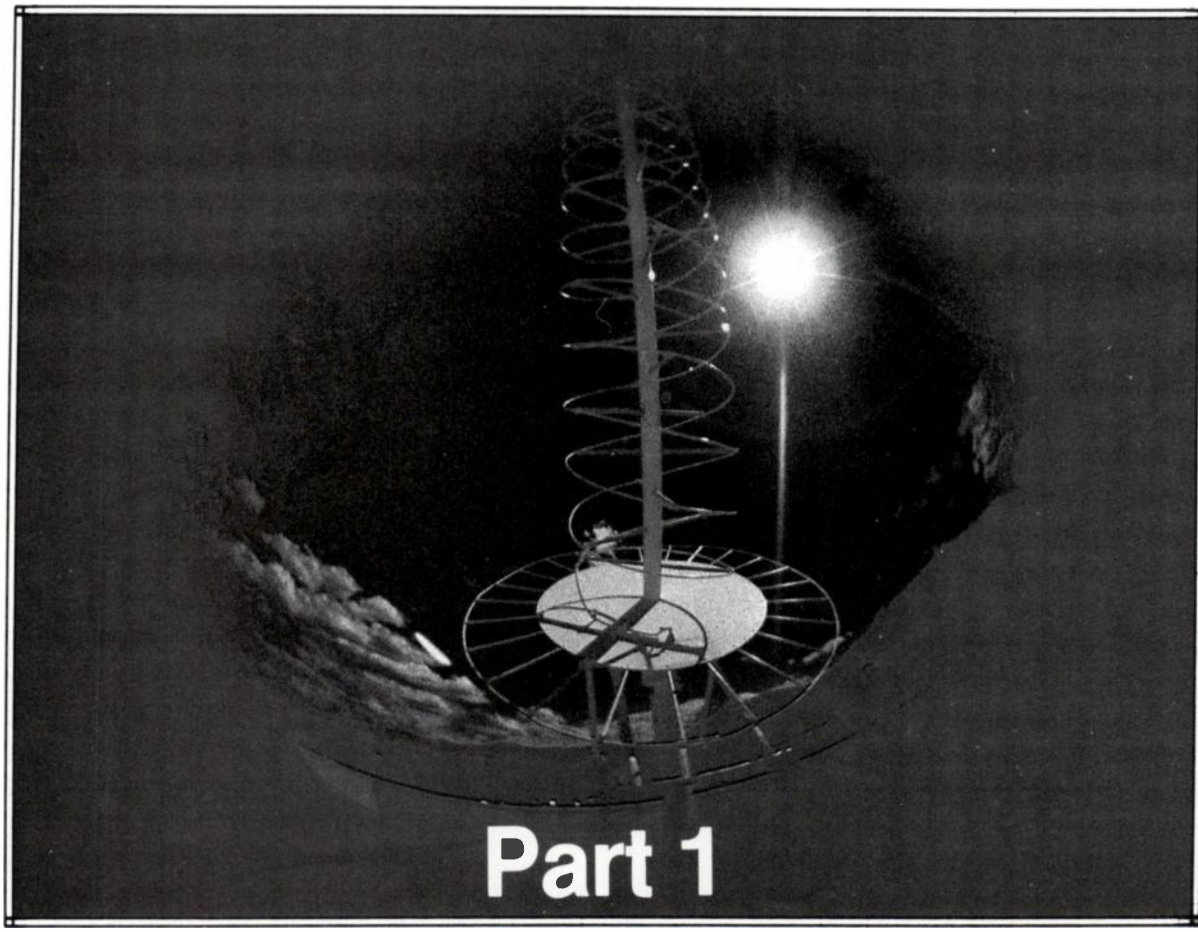
Antennas in the RF Range

. . . 10 kHz to 2000 MHz. A discussion of RF antennas used by all services.



Most antennas in the RF range fall into one of following categories:

A. Traveling wave
B. Driven
C. Vertical
D. Leaky Waveguide
E. Reflector

An antenna, as a transducer, converts electro-magnetic energy into RF currents for receiving applications or launches an electro-magnetic wave from an RF source(s).

Though there is the appearance of five separate categories there is some degree of overlap just as there is overlap in the frequency domain and usage of all antennas.

Vertical antennas, especially when combined in arrays are not really a distinct group. They fall into the traveling wave or driven antenna category. However, especially in the lower end of the RF spectrum (10 kHz-30 MHz), vertical antennas are widely used and are discussed as a separate group.

Leaky wave antennas, such as the Directional Discontinuity Ring Radiator, or DDRR for short, though they launch and receive a vertically polarized wave and consequently could be grouped as "C", due to their unique construction are considered separately.

Traveling Wave Antennas

Type 1. This is a group of endfire antennas utilizing an array of parasitic elements or wave guiding structure with energy directed towards or from a primary driven element. The categories are:

1. Yagi-Uda Parasitic Array
2. Backfire
3. Quad or closed loop array

4. Quagi (Quad-Yagi)
5. Helical
6. Birdcage

Once again not all categories are distinct. A backfire antenna is a Yagi and reflector combination. However since modifications are required to the Yagi when combined with the reflective screen it is at best a hybrid and will be considered separately. (This is not meant to diminish the importance of it firing off the back and not the front end of the Yagi.)

The same holds true for the Quagi. Though constructed of "Yagi type" elements and closed loops by virtue of its hybrid performance rates a separate category.

1. Yagi-Uda Parasitic Array

The Yagi-Uda antenna, invented in 1926, is a parasitic structure that supports a slow surface wave along its length. It consists of a number of directors and reflectors (the parasitic elements) that enhance radiation in one direction.

The Yagi has been utilized by the military, commercial interest and experimenters alike. It is simple, rugged, can be made of all metal construction and consequently be DC grounded for lightning protection.

The simplest Yagi consists of a driven element and either a director or reflector. On the other extreme is an 80 wavelength long, 26 dBd (dB with respect to a reference dipole) experimental Yagi.[1] One of the longest boom Yagis in present use consists of 12 elements and is 152 feet long.[10]

A typical 3 element Yagi is depicted below.[2] (Figure 1)

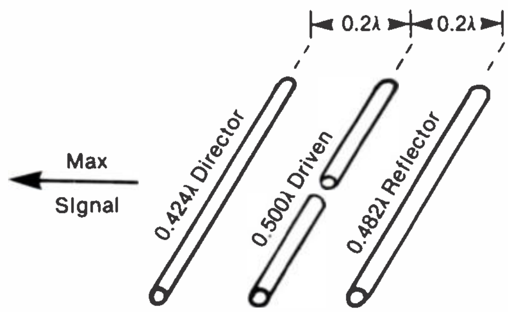


Figure 1. Typical 3 Element Yagi

Its elements can be linear, X-shaped or in the form of inverted Vees, each with its own merits.

a. Standard Design

There are several "recipes" for standardized construction of 3 or more element Yagis. One such recipe suggests that a single reflector be placed between 0.2 and 0.25 wavelengths behind the driven element and each added director of 2.5 percent diminishing length and 0.2 wavelength separation in front of the driven element.

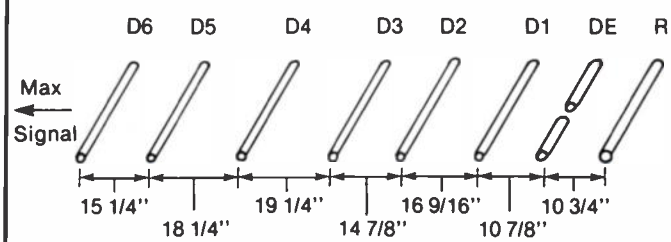
The Hansen-Woodyard condition[1] specifies that high gains are achievable as long as the difference in phase between the surface wave and the free space wave at the end of the antenna be equal to 180°. A general rule of thumb has always been "the longer it is the better it is (higher gain).

Recently several references have developed non-standard techniques for optimizing gain for a given boom length Yagi.

b. The Ascendency Curve Yagi[3]

Represents a design by formula called the ascendency — descendency curve. It recognizes the interrelationships between the high number directors (5 and 6) and their effect in increasing the overall forward gain. Improvements of the order of 10 to 20 percent are indicated by the careful adjustments of non-uniform director *spacings.* An example of an 8 element, 13.1 dBd gain Yagi is given. For actual design information the reader is referred to the original source.[3] (Figure 2).

D6 = 6th Director
DE = Driven Element
R = Reflector



D6 = 6th Director DE = Driven Element R = Reflector

Figure 2. 8 Element Ascendency Curve Yagi

Note the ascending — descending spacings between the higher number directors. This Yagi is dimensioned in inches for operation at 222.5 MHz. (Which set of element lengths to use depends upon design bandwidth requirements and are clearly delineated in the reference.)

c. The Modulated Ladder Antenna

The modulated ladder antenna is basically a multi-element Yagi with non-uniform or "modulated" element lengths and spacings. The advantages indicated by this design are very high gain (26 dBd)[1] and minimum sidelobes over a substantial bandwidth. Properly designed modulated ladder arrays should exhibit an average beamwidth (H-plane beamwidth is larger, E-plane beamwidth is less) and gain as calculated from the formulas:

Beamwidth = $55\sqrt{\ell/\lambda}$ in degrees

Gain = $10 \log_{10}(9\ell/\lambda)$ in dB above isotropic

ℓ = array length (same units as λ)

The concept relating higher achievable gains with non-uniform element lengths and spacings has been described in greater detail by David Cheng.[4,5]

d. The Trigonal Reflector

Most Yagi designs incorporate a single reflector. Experiments[2] utilizing plane reflecting surfaces, parabolas and corner reflectors were tried. The largest increase in gain (0.75 dB) occurred when a trigonal reflector was utilized as illustrated below in Figure 3.

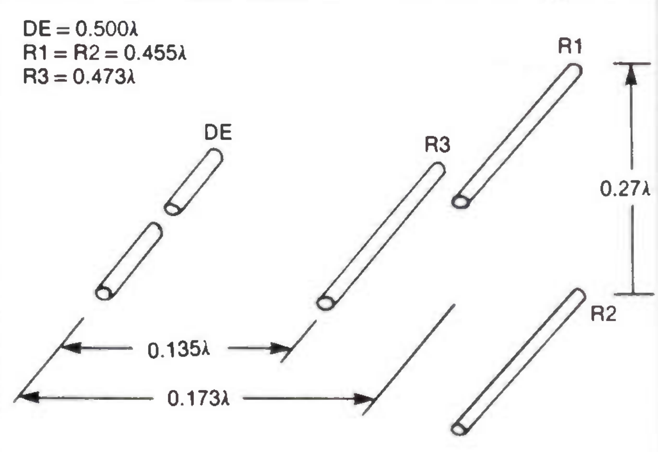


Figure 3. Yagi with Trigonal Reflector

With well over 400 U.S. references to Yagi antenna development available since 1928[6] and several definitive projects, some involving 3 or more years of research,[2] the Yagi antenna is probably one of the most interesting, useful and well studied antennas around.

2. Backfire Antenna

The Hansen-Woodyard condition, referred to previously, stated that the maximum gain of a Yagi is proportional to its length provided that the surface and free space wave 180° phase relationship is maintained. A Backfire antenna is basically a Yagi pointing towards a large reflector. (Figure 4)

LR = Large Reflector
D = Director
DE = Driven Element
R = Reflector

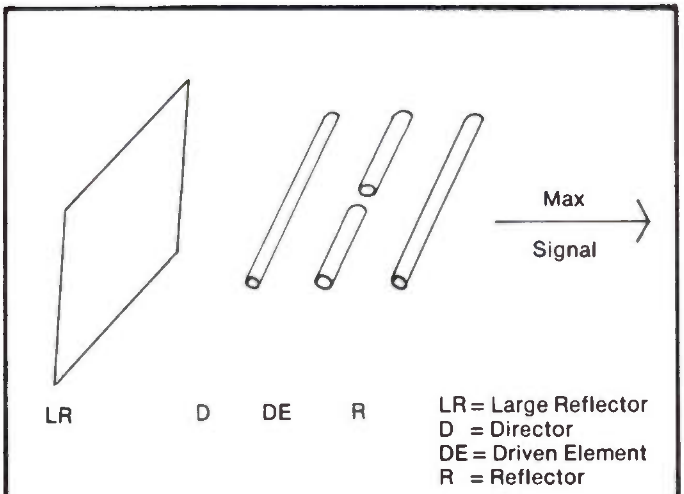


Figure 4. The Backfire Antenna

It takes advantage of the H-W condition by forcing the surface wave to traverse the physical length *twice*. This results in greater gain and frequency bandwidth than its equivalent length Yagi.[7]

Backfire antennas have been used at the high end of the RF range with gains in excess of 23 dB, at UHF with gains of 10.5 to 14 dB[7] and at VHF with gain improvements of 4.5 dB over the isolated Yagi.[8]

3. Quad or Closed Loop Array

The cubical quad, or quad for short, was developed in 1942 by Clarence Moore.[9] It is a parasitic traveling wave structure with loop elements approximately one wavelength in circumference. It is similar to the Yagi in several respects yet its differences are more widely known. These are:

a. Approximately 2 dB greater gain than the equivalent length Yagi.
b. "Quieter operation" — Less sensitive to precipitation static
c. Feedpoint impedance closer in value to standard transmission line characteristic impedances
d. Lower mounting height for the same takeoff or elevation angle as that of a Yagi.
e. Polarization can be changed by simply changing feedpoints.

Handbook style design criteria for a 3 element quad follows from these basic formula:

1) Circumference of driven element = 1005/fMHz feet
2) Circumference of reflector element = 1030/fMHz feet
3) Circumference of director element = 975/fMHz feet

Interelement spacings of from 0.15 λ to 0.20 λ are typical.

Power gain in dB above an isotropic of a closed loop is more a function of its perimeter and not shape. This has given use to loop designs with other than a diamond or rhombus configuration.

- **Circular Loops**

Research on large circular loop arrays[11] have indicated that the loop perimeter (or enclosed area) and relative thickness of the conductors are gain determining factors. Antenna area and gain are related as follows:

$$G = 4\pi A/\lambda^2$$

where A (area) and λ^2 are entered with the same units of measure.

Figure 5 illustrates current distribution in diamond, square and circular 1 wavelength circumference loops:

- **Delta Loops**

Another popular form of loop array has a delta or triangular shaped element. It has the added advantage, especially at the lower end of the HF spectrum, in that it only requires a single high support, per element.

Delta loops exhibit substantially different takeoff angles when inverted versus non-inverted, fed at a corner or side and elevated versus erected close to the ground or ground system.[12,13,14]

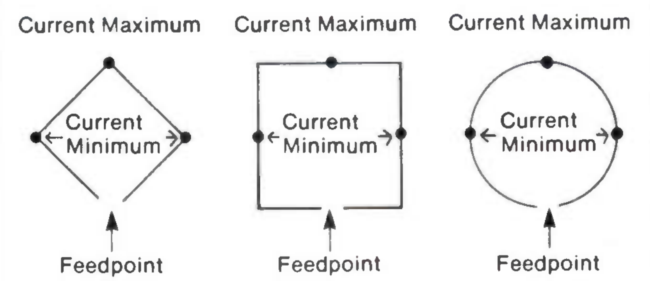


Figure 5. Closed Loop Current Distribution

- **Bi-square Array**

As mentioned earlier conventional cubical quads have a 1 wavelength perimeter. A diamond-shaped quad would then have 4 sides each $\lambda/4$ long. The Bi-square loop is similar in shape and radiation direction (broadside) to the quad with the added advantage of 3 dB more gain. This additional gain is achieved by making the sides $\lambda/2$ in length. The total perimeter of a Bi-square loop is $4 \times \lambda/2$ or 2λ. By adding a parasitic reflector or director gains over 9 dBd have been claimed.[15]

- **Alford Loop**

One means of increasing antenna efficiency is to increase its radiation resistance. The Alford loop is a large square loop, consisting of four dipole radiators for sides. The radiation resistance is much higher than that of an ordinary loop. A typical Alford loop[16] is indicated in Figure 6.

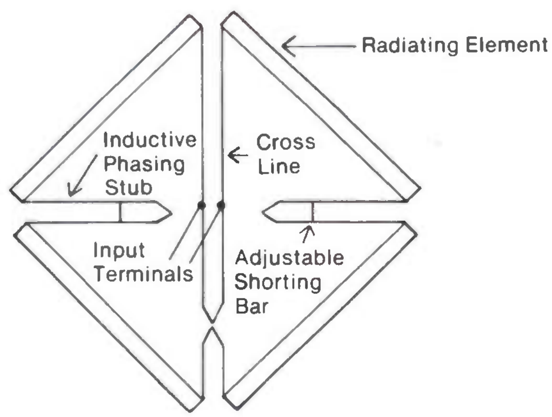


Figure 6. The Alford Loop

4. Quagis

The Quagi is a hybrid antenna that combines the best features of the cubical quad and the Yagi beam. Its directors are linear elements as in a Yagi while the driven and reflector are closed loops.

The immediately apparent advantages are greater gain over an equivalent number of element Yagi and simpler and more repeatable construction techniques. (Figure 7)

D4, D3, D2, D1 = Directors
DE = Driven Element
R = Reflector

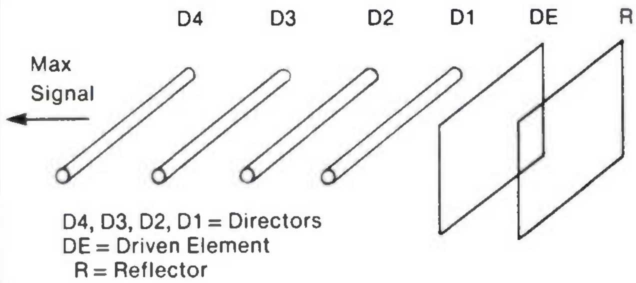


Figure 7. A 6 Element Quagi

5. Helicals

The front cover of this month's issue depicts a right-hand circularly polarized helical antenna operating at 468 MHz. Unlike dipole antennas where the electro-magnetic wave travels at the velocity of light the velocity of the wave along the axis of the helix is lower. It depends on the frequency, diameter and number of turns per unit length.

Helical antennas are categorized by their shape (cylindrical, flat or conical) and their mode of operation: normal or axial.

- **Normal Mode**

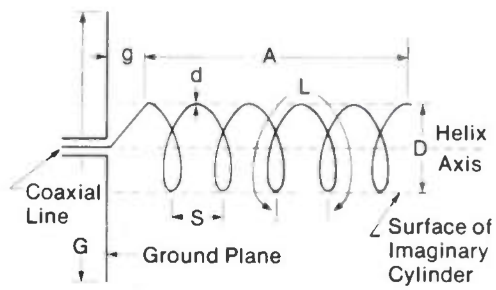
Distinguishing characteristics of a normal mode helix are their small diameters and lengths with respect to wavelength. The radiated field is elliptically polarized. Circular polarization is obtained with a resonant helix at a 0.9 height/diameter ratio.[17]

- **Axial Mode**

The axial mode helix is used as an end-fire circularly-polarized radiating beam. Gain is nearly constant over an octave. Both gain and beamwidth are a function of the number of turns — gain increases, the beamwidth decreases with increasing turns.

Axial mode helices are used for communications between terrestrial stations and orbiting satellites. Figure 8 illustrates a helix and its dimensions utilized for just such an application — radio amateur communications with OSCAR 7.[26]

D = Diameter of Helix
S = Spacing between turns
a = Pitch angle = Arctan ($S/\pi D$)
L = Length of 1 turn
n = Number of turns
A = Axial length = nS



D = Diameter of Helix
S = Spacing between turns
a = Pitch angle = Arctan (S/πD)
L = Length of 1 turn
n = Number of turns
A = Axial length = nS
d = Diameter of conductor
g = Separation between ground
G = Diameter of groundplane

Figure 8. Helix and Dimensions

d = Diameter of conductor
g = separation between ground plane and first turn
G = Diameter of ground plane

6. Birdcage

The birdcage is a unique 2 element parasitic array that borrows favorable characteristics from the cubical quad and Vee beams. Invented and patented by Dick Bird, a British radio amateur, in 1958,[18] it closely resembles a bird cage. (Figure 9)

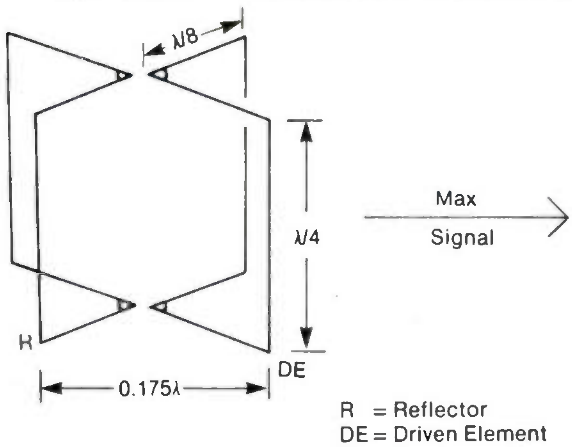


Figure 9. The G4ZU Birdcage Antenna

Several attractive features of the Birdcage are:
a. High gain — between 8.5 to 10 dBd.
b. Extremely compact — very small turning radius even at the low end of the HF spectrum.
c. Feedpoint impedance — 40 to 50 ohms, ideal for matching to standard coaxial transmission lines.

Traveling Wave Antennas

Type 2. This is a group of antennas using a wave guiding structure, primarily wire, that is long with respect to the operating wavelength. These antennas are additionally grouped by their height above ground.

- **Elevated above ground (>λ/8)**

1. Long wire antennas
2. Vee antennas, Vee beams
3. Rhombics

Unlike the previous Type 1 (Traveling Wave Antenna) Type 2 antennas evolve. That is long wires "1" make up Vee antennas "2" and Vees combine to form Rhombics "3".

1. Long-wire Antennas

The old adage "the bigger the better" applies to this type of antenna. A long wire antenna when fed at one end either directly (Fuchs antenna) or by means of a transmission line has a radiation pattern that directly relates to its length and height above ground. Radiation from end-fed long wire antennas is broadside when the overall length is 5/8 wave or less.

As the length of the long wire increases above 5/8 wave the radiation pattern forms into a cone coaxial and away from the feed point as illustrated below. (Figure 10)

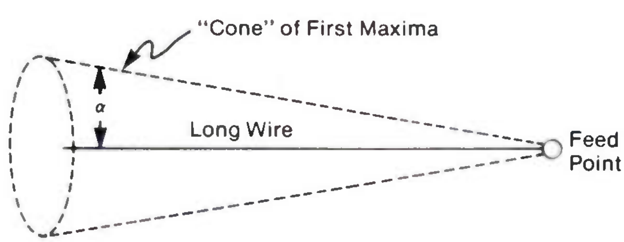


Figure 10. Long wire Antenna Pattern

A table of the locus (angle) of first maxima of the cone of radiation versus long wire length is given below: (Table 1)

Table 1

α	λ	α	λ
53°	1	23°	5
37°	2	20°	6
29°	3	18°	7
26°	4	17°	8

The advanages ot this type of antenna are its simplicity, high gain (for many wavelength long wires) and low cost to install and maintain.

2. Vee Antennas or Vee Beams

Vee antennas represent a natural progression from long wire antennas. They have the added advantages of greater gain, (3 dB versus same length long wires) sharper directivity and reduction of some sidelobes. Two long wires of specified length (λ) can be combined at one point to form a Vee beam of related angle (2α). This is to say that length and apex angle are related in optimized gain Vees.

Figure 11 illustrates spatial lobe combination and cancellation in the long wire construction of a Vee beam.

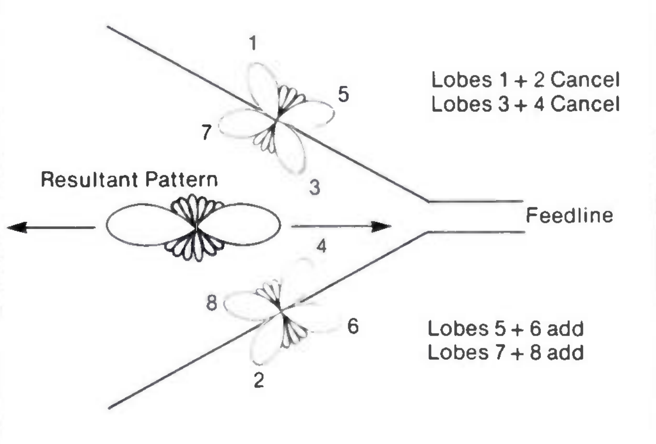


Figure 11. Vee Beam Construction from 2 Long Wires

As illustrated a Vee beam, when properly constructed is bidirectional with maximum radiation along the bisector.

When resistively terminated at the ends the Vee Beam is unidirectional and radiates towards (and receives from) the left.

3. Rhombics

a. General

A rhombic is a four-sided, diamond-shaped high gain wire antenna. It provides a high signal-to-noise ratio for reception and a relatively low take-off or elevation angle. It exhibits a nearly constant feedpoint impedance over a wide bandwidth when properly terminated at the far end. A terminated rhombic is illustrated below. (Figure 12)

Maximum radiation is indicated along the bisector towards the left. Design terminology and criteria are as follows:

ϕ = tilt angle
Δ = vertical angle
H = height above ground
L = long side length

Using readily available charts[19] and assigning values to two out of four of the parameters the remaining two are calculated.

Assume a 15° vertical angle is required. (Based on a specific propagation path and mode) and a maximum side length of 5λ is all the existing property will allow.

What tilt angle and height above ground does that calculate out to?

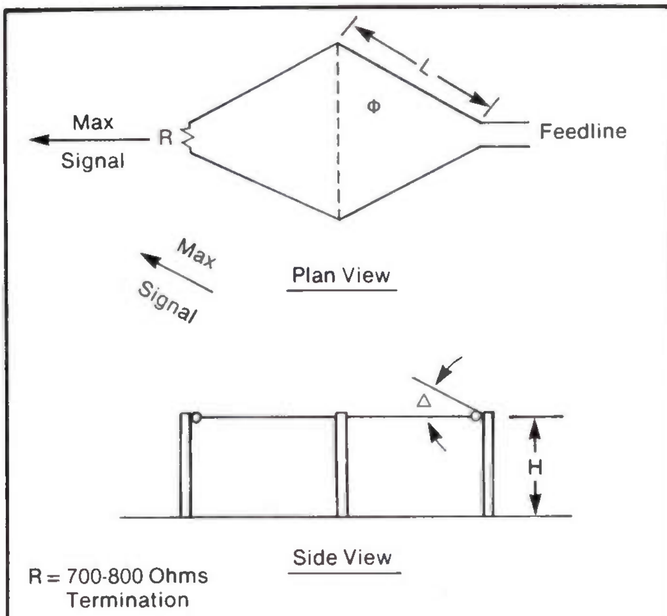


Figure 12. Rhombic Antenna

The referenced chart indicates that a tilt angle ϕ of 71° and a height of 0.95λ or 2.8λ will suffice. The lower of the two (heights) for practical considerations is chosen.

Terminations for rhombics are non-inductive resistors or in the case of very high power transmitters lossy balanced transmission line is utilized.

b. Multistrand Rhombics

Often for broadbanding purposes or for lowering the feedpoint impedance rhombics are multistranded. That is each leg consists of two or more conductors separated by several feet at the side poles and brought together at the forward and rear supports.

c. Nested Rhombics

A nested rhombic configuration is a pair of staggered coplanar rhombics with a common feedpoint. It exhibits substantial gain (27 dBd), broadbandedness, low vertical angle and lower (sidelobes) than a conventional rhombic.

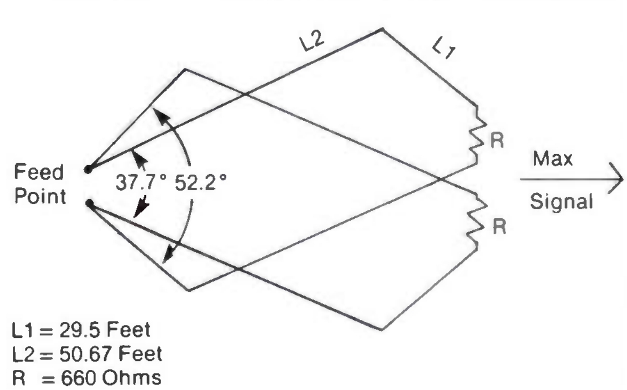


Figure 13. 144 MHz Nested Rhombic

This nested rhombic is fed by 300 ohm balanced line, is mounted 12.29 feet above ground and has a vertical and horizontal beamwidth of 5.5° and 8.5°, respectively.

- **Close to the Ground (≤ λ/8)**
- **Beverages**

A Beverage antenna is a horizontal longwire antenna utilized especially for reception of low-frequency vertical polarized waves. It normally consists of a single wire several wavelengths long erected close to the ground, or even on the ground, and terminated at the far end in its characteristic impedance.

An outstanding merit of this antenna is its ability to "pull signals out of the noise". The low end of the HF spectrum is typically atmosphere noise limited as opposed to thermal noise limiting in the higher RF ranges. Most HF antennas previously described present large signal *and* noise voltages at receiver terminals. The Beverage, on the other hand, generates a lower signal voltage while at the same time a considerably lower noise voltage. The overall effect is to produce a better or higher signal-to-noise ratio which is really the name of the game in communications intelligibility.

- **Parallel Wire Beverage Antenna**

An interesting and very useful feature related to the operation of a Beverage is the ability to reject signals not only off the back but off the sides — selectively. By tuning the complex termination impedance depth and direction of rejection can be adjusted.

A twin or parallel wire Beverage places the termination network close to the receiver (and operator). (Figure 14)

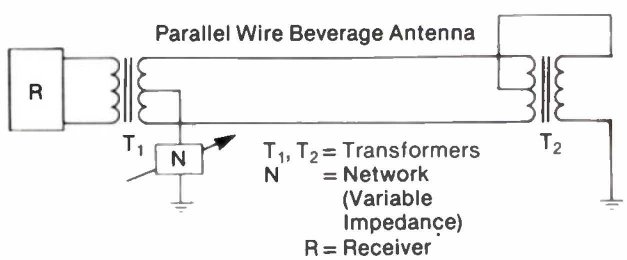


Figure 14. Parallel Wire Beverage Antenna

T_1, T_2 = Transformers
N = Network (Variable Impedance)
R = Receiver

Several comprehensive research efforts have explored many aspects of Beverage antenna design, development, deployment and measurement.[20, 21, 22, 23, 24] They are well worth reading about.

Part 2 in the series continues with "Driven Antennas."

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Acknowledgment

Cover photo made available by Mr. Kent Higgins, and Mr. Collier Smith, Public Information Specialists of the National Bureau of Standards.

7-Element 50-ohm Chebyshev Filters Using Standard-Value Capacitors

By proper selection of reflection coefficient and cutoff frequency, one can obtain a design very close to the required cutoff frequency and use only standard value capacitors.

Edward E. Wetherhold
Honeywell Inc.
Defense Electronics Division

Ocassionally, the RF engineer requires a relatively simple lowpass or highpass LC filter for non-critical applications such as for reducing oscillator harmonic levels, or for providing receiver preselection. In these cases, the exact cutoff frequency is not critical and 40 to 50 dB of attenuation per octave is adequate. The average RF engineer may spend too much time in developing a suitable design. Although the lowpass design procedure is relatively simple and straightforward and can usually be completed without error, this is not true of the highpass procedure. In this case, the normalized lowpass element values must be first transformed into highpass values before frequency and impedance scaling is done. This complication provides additional opportunities for error. Even after the design calculations have been completed, the construction invariably will be complicated because most or all of the calculated capacitor values will not be available in the standard-value series. The designer then has the choice of either accepting the design as calculated, or shifting the cutoff frequency slightly to cause a corresponding shift in the capacitor value so it will be a standard value. Of course, the non-standard design values can be achieved by paralleling capacitors or using variables. But this procedure lacks elegance, and it is aesthetically unappealing, both from economical and performance standpoints. Two capacitors are more costly than one, and paralleled capacitors may have unexpected and undesired resonances due to unavoidable internal and external stray inductances. The most suitable design is that which provides an acceptable performance with the minimum number of components. For this reason, all designs should be such that they can be constructed with only standard-value capacitors. The cutoff frequency may be slightly shifted to make the calculated non-standard capacitor values become standard values. However, this is applicable only for the equally-loaded 3-element configuration or for the 5-element inductor input/output configuration where only one capacitor value is involved. For the 5 and 7-element capacitor input/output configurations, a second capacitor value (different from the first) occurs, making it unlikely that both values could be made to become standard values by just shifting the cutoff frequency. However, if the reflection coefficient of the filter is varied, the capacitor ratio may be changed. Thus, by proper selection of both the reflection coefficient and cutoff frequency, it is possible to obtain a design that will simultaneously be close enough to the required cutoff frequency, and require only standard-value capacitors.

For most non-critical filtering applications where standard-value capacitors are desired, the usual design procedure is inconvenient. A better procedure is to first determine what standard capacitor values are available and then develop a series of designs using these values. Because a perfectly flat passband is seldom required, a Chebyshev response (with a small amount of passband ripple) can be used so the capacitor ratios can be varied to obtain ratios that are identical with those of the standard capacitor values. The number of capacitor values should be sufficiently numerous so that the increments of the cutoff frequencies will be small enough to provide any cutoff one might require (say within ± 5 percent of any desired frequency). For example, if a 1 MHz filter is desired, a filter having a cutoff anywhere between 0.95 and 1.05 MHz should suffice. The filter terminations can be fixed at 50Ω impedance. Inductors used are usually hand-wound to get the required value.

Pre-calculated filter tables of the 5-element Chebyshev type have already been published. The first set of LP and HP tables were published in 1972[1], and listed eighteen designs each of LP and HP. A more comprehensive set of tables was published in

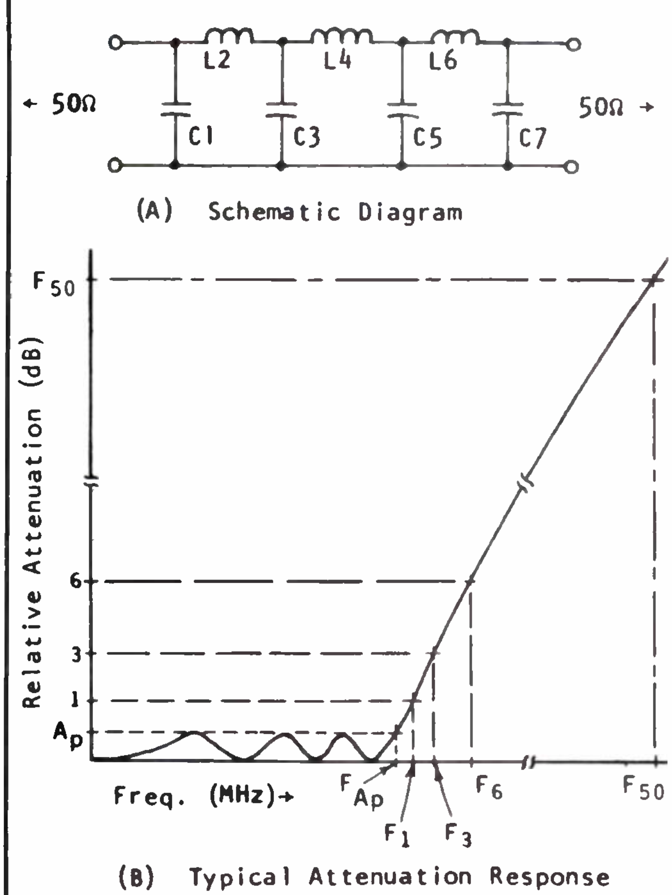

Figure 1.

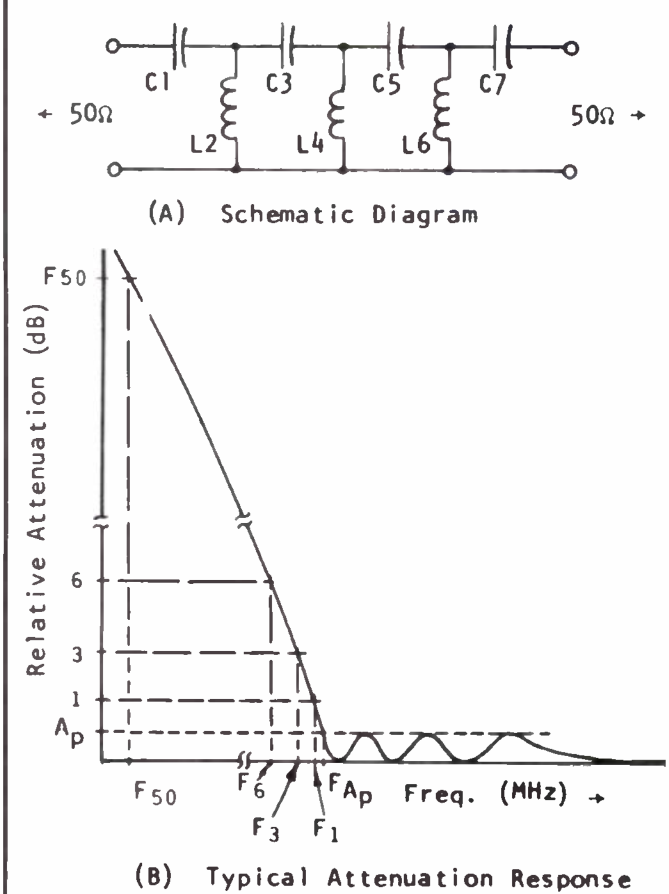

Figure 2.

1978[2], and included were attenuation versus frequency parameters in addition to the component values. The practicality and usefulness of such tables has been recognized and improvements made to the 1978 lowpass and highpass filter tables justify the publication of a new listing.

Selection of Filter Type

The selection of the filter type for the proposed pre-calculated tables is based on the fact that the attenuation response is of primary concern; that is, the signals to be filtered are primarily steady-state sine waves, and the phase and impulse responses are not important. Also, inductor Q is assumed to be high (>70) so that inductor losses may be ignored. The Chebyshev type has the design and performance characteristics best suited for non-critical filtering applications in the sense that of all possible filter networks with transmission zeroes at infinity (all-pole functions) it has the lowest complexity for providing a prescribed maximum passband attenuation (Ap) and the most abrupt rise of attenuation outside the passband [See Figures 1(B) and 2(B) for a graphical representation of filter parameter Ap.]. Also, the Chebyshev design and performance parameters are easily calculated for any given level of Ap. Of course, the elliptic type can be made to provide a more abrupt rise in attenuation than possible with the Chebyshev type, but at the expense of additional components that must be tuned to the correct frequencies. Also, the elliptic designs are restricted to those normalized values that are published in handbooks such as references 5, 7 and 8. Because of these disadvantages of the elliptic filter type, it is recommended that the inexperienced filter designers restrict themselves to the Chebyshev designs discussed in this article.

After selecting the Chebyshev filter type, additional decisions must be made as to what should be the maximum value of Ap, what filter configuration is preferable, and how many elements should be used. The Chebyshev response is especially useful in that one is free to choose any value of Ap for a corresponding attenuation slope near the cutoff frequency. Thus, an infinite number of designs are possible between Ap levels of zero dB (corresponding to a Butterworth response) and 3 dB, which is the highest possible Ap level. Other design parameters directly related to the Ap level are the filter reflection coefficient (ρ) and VSWR.* As the Ap level and the attenuation slope increase, so do the values of ρ and VSWR. Because it is desirable to minimize passband attenuation and VSWR and because designs with low Ap are less critical to construct (high levels of Ap, such as 3 dB, require closer tolerance and higher Q components and closer tolerance terminations to obtain the expected response), an arbitrary limit of Ap = .1583 dB has been selected. The corresponding values of VSWR and ρ are 1.467 and 0.1892, or 18.92 percent. The number of Chebyshev designs available between ρ = zero (Butterworth response) and 18.92 percent are quite sufficient for our purposes.

The capacitor input/output configuration [see Figures 1(A) and 1(B) was selected instead of the alternate inductor input/output configuration to minimize the number of inductors. For the engineer who must wind his own inductors, the capacitor input/output configuration is more convenient to construct, is less expensive and the filter will usually perform better (because capacitors have higher Q than inductors) as compared to the alternate configuration.

The seven-element filter was chosen because it is optimum in the sense of providing a minimum acceptable attenuation rise near the cutoff frequency (42 dB/octave) for a reasonable number of elements. In the equally-terminated 7-element filter [see Figures 1(A) and 2(A)], capacitors C1 and C7 have the same value, and capacitors C3 and C5 are also identical in value. Inductors L2 and L6 are also identical in value. The C3/C1 ratio varies from 4.0489 (for a Butterworth lowpass response) to 1.6968 (for a ρ = 18.92 percent Chebyshev lowpass response). Because the capacitor ratio is directly related to the Chebyshev design and performance parameters, it is easy to calculate a Chebyshev filter table for any number of different capacitor ratios. This is true for the 5 and 7-element capacitor input/output configuration and also for the 9-element inductor input/output configuration. Either of these three configurations could be used in the calculation of a filter table where only one capacitor ratio per design is involved; but, because the 7-element is most suitable for this purpose, it will be used. (The 5-element filter does not provide sufficient attenuation, and the 9-element L-in/out filter has too many inductors.)

Capacitor Selection And Ratio Determination

As previously explained, the C3/C1 ratio of the 7-element lowpass (LP) C-in/out Chebyshev filter automatically specifies a particular Chebyshev LP design whose component and performance parameters can be calculated. (The highpass (HP) filter C-ratio is identical to the reciprocal of the LP ratio.) The selection of capacitor values must be such that the increment of the cutoff frequencies from one design to the next will be small enough so any desired cutoff frequency will always be within several percent of one of the pre-calculated filter designs. This will be acceptable for all expected applications of these filters. Those desired cutoff frequencies above or below the tabulated cutoff frequency decade can be obtained by simply shifting the decimal point of the tabulated data.

The twelve preferred standard values of the 10 percent series** (10,12,15,18,22,27,33,39,47,56,68, and 82) were used as the basis for determining the capacitor ratios between the maximum ratio of 4.0489 (corresponding to a Butterworth response) and the minimum ratio of 1.6968 (corresponding to a 18.92 percent Chebyshev response). A computer was used to calculate the ratios of all possible combinations of capacitors in the 10 percent standard-value series, and then the reflection coefficients of the Chebyshev designs corresponding to each ratio between the maximum and minimum limits was calculated. The computer was again used to determine the parameters of all possible filter designs having the previously calculated C-ratios. The computer was programmed to provide designs only over the one to ten megahertz frequency decade. The results of these calculations for LP and HP filters are shown in Tables 1 and 2, and the corresponding schematic diagrams and responses are shown in Figures 1 and 2.

A Tabulation of Sixty 50-Ohm Lowpass Filters

Table I shows the LP filter tabulation in which each design is identified with a filter number in the first column. The next six columns list the frequencies corresponding to the attenuation levels in the column headings. This permits one to quickly sketch the attenuation response of each filter up to the 50 dB level. The next two columns specify the percentage of reflection coefficient and the maximum VSWR. The last four columns list the capacitor and inductor values of the filter. The fact that only standard-value capacitors are required in these designs can be confirmed by referring to the C1,7 and C3,5 columns where only stand-values are listed. By referring to the C3,5 column, it is obvious that all of the twelve values of the 10 percent series in one decade have been accounted for. The C1,7 values are those that provide the C3/C1 ratio within the previously mentioned limits. A consistent and orderly decrease in the values of C1,7 and C3,5 causes a gradual increase in the 3-dB cutoff frequency, and the increments are small enough so that any non-critical filtering

*See Appendix B for equations relating A_p, $|\rho|$, and VSWR. VSWR is the ratio of the absolute magnitude of the filter input impedance (measured with the filter output terminated in its design resistance, R_o) to the design termination resistance, R_o. VSWR is never less than 1, thus: VSWR = $|Z_{in}|/R_o$ or $R_o/|Z_{in}|$, for VSWR>1.

**A series of values based on a step multiplier of $(10)^{1/12}$, adopted by the EIA, and now a USA Standard (C83.2).

FLTR	FREQ. (MHZ) AT AP,1,3,6,30 & 50 DB						R.C.	VSWR	C1,7	C3,5	L2,6	L4
NO.	AP-DB	1-DB	3-DB	6-DB	30DB	50DB	(%)	(MAX.)	(PF)	(PF)	(UH)	(UH)
	MEGAHERTZ											
1	.430	.951	1.038	1.12	1.65	2.3	.4550E-02	1.000	1500	5600	10.11	15.26
2	.714	.991	1.063	1.13	1.60	2.2	.2555E 00	1.005	1800	5600	10.92	14.77
3	.886	1.038	1.095	1.15	1.56	2.1	.1782E 01	1.036	2200	5600	11.36	14.04
4	1.037	1.117	1.162	1.21	1.58	2.1	.6629E 01	1.142	2700	5600	10.90	12.57
5	1.255	1.289	1.326	1.36	1.71	2.2	.1891E 02	1.466	3300	5600	8.86	9.69
6	.336	1.119	1.227	1.32	1.99	2.8	.2063E-03	1.000	1200	4700	8.28	12.94
7	.842	1.180	1.265	1.34	1.91	2.6	.2328E 00	1.005	1500	4700	9.14	12.41
8	1.034	1.229	1.299	1.36	1.86	2.5	.1466E 01	1.030	1800	4700	9.52	11.88
9	1.208	1.313	1.369	1.42	1.87	2.4	.5600E 01	1.119	2200	4700	9.27	10.78
10	1.451	1.498	1.542	1.59	2.00	2.6	.1665E 02	1.400	2700	4700	7.75	8.53
11	.429	1.350	1.480	1.60	2.39	3.3	.3140E-03	1.000	1000	3900	6.89	10.73
12	.954	1.411	1.518	1.61	2.31	3.1	.1379E 00	1.003	1200	3900	7.49	10.37
13	1.250	1.483	1.567	1.65	2.25	3.0	.1514E 01	1.031	1500	3900	7.90	9.85
14	1.440	1.572	1.641	1.71	2.25	3.0	.5160E 01	1.109	1800	3900	7.73	9.04
15	1.714	1.775	1.830	1.89	2.39	3.1	.1519E 02	1.358	2200	3900	6.60	7.30
16	.208	1.581	1.740	1.88	2.84	3.9	.5540E-06	1.000	820	3300	5.73	9.14
17	1.094	1.661	1.790	1.91	2.74	3.7	.1078E 00	1.002	1000	3300	6.30	8.80
18	1.404	1.729	1.834	1.93	2.67	3.6	.9490E 00	1.019	1200	3300	6.64	8.47
19	1.682	1.847	1.929	2.01	2.65	3.5	.4710E 01	1.099	1500	3300	6.58	7.72
20	1.957	2.041	2.108	2.18	2.77	3.6	.1280E 02	1.294	1800	3300	5.83	6.50
21	.452	1.940	2.132	2.30	3.47	4.8	.3260E-04	1.000	680	2700	4.72	7.46
22	1.344	2.032	2.188	2.33	3.35	4.6	.1119E 00	1.002	820	2700	5.16	7.19
23	1.746	2.122	2.248	2.37	3.26	4.4	.1112E 01	1.022	1000	2700	5.45	6.89
24	2.022	2.238	2.341	2.44	3.24	4.3	.4112E 01	1.086	1200	2700	5.41	6.40
25	2.436	2.531	2.611	2.69	3.42	4.4	.1406E 02	1.327	1500	2700	4.66	5.18
26	.685	2.389	2.621	2.83	4.25	5.9	.1479E-03	1.000	560	2200	3.87	6.06
27	1.705	2.503	2.692	2.86	4.10	5.6	.1478E 00	1.003	680	2200	4.23	5.84
28	2.155	2.608	2.763	2.91	3.99	5.3	.1174E 01	1.024	820	2200	4.44	5.61
29	2.524	2.770	2.894	3.01	3.98	5.2	.4710E 01	1.099	1000	2200	4.39	5.15
30	2.935	3.062	3.162	3.27	4.16	5.4	.1281E 02	1.294	1200	2200	3.88	4.33
31	1.123	2.940	3.216	3.46	5.16	7.1	.1235E-02	1.000	470	1800	3.21	4.93
32	2.108	3.063	3.293	3.50	5.00	6.8	.1629E 00	1.003	560	1800	3.47	4.78
33	2.667	3.198	3.384	3.56	4.87	6.5	.1312E 01	1.027	680	1800	3.64	4.57
34	3.090	3.389	3.540	3.69	4.87	6.4	.4776E 01	1.100	820	1800	3.59	4.21
35	3.654	3.797	3.917	4.04	5.13	6.6	.1406E 02	1.327	1000	1800	3.11	3.45
36	1.298	3.524	3.857	4.15	6.20	8.6	.9369E-03	1.000	390	1500	2.67	4.11
37	2.560	3.682	3.954	4.20	5.99	8.1	.1806E 00	1.004	470	1500	2.90	3.97
38	3.166	3.827	4.053	4.26	5.86	7.8	.1190E 01	1.024	560	1500	3.03	3.82
39	3.694	4.059	4.241	4.42	5.84	7.7	.4640E 01	1.097	680	1500	2.99	3.52
40	4.314	4.498	4.645	4.80	6.11	7.9	.1295E 02	1.298	820	1500	2.64	2.94
41	2.282	4.469	4.861	5.21	7.67	10.5	.1195E-01	1.000	330	1200	2.19	3.25
42	3.386	4.638	4.966	5.27	7.45	10.1	.2939E 00	1.006	390	1200	2.35	3.16
43	4.125	4.841	5.109	5.36	7.29	9.7	.1[illegible]42E 01	1.035	470	1200	2.43	3.01
44	4.719	5.134	5.354	5.57	7.32	9.6	.5502E 01	1.116	560	1200	2.37	2.76
45	5.597	5.792	5.968	6.15	7.78	10.0	.1554E 02	1.368	680	1200	2.02	2.23
46	2.516	5.339	5.819	6.25	9.24	12.7	.6280E-02	1.000	270	1000	1.81	2.72
47	4.150	5.583	5.971	6.33	8.92	12.0	.3533E 00	1.007	330	1000	1.97	2.62
48	4.933	5.804	6.126	6.43	8.75	11.[illegible]	.1687E 01	1.034	390	1000	2.03	2.51
49	5.694	6.180	6.442	6.70	8.79	11.5	.5730E 01	1.122	470	1000	1.97	2.29
50	6.632	6.879	7.094	7.32	9.27	12.0	.1464E 02	1.343	560	1000	1.71	1.89
51	2.966	6.500	7.089	7.62	11.3	15.5	.4870E-02	1.000	220	820	1.48	2.23
52	5.046	6.806	7.279	7.71	10.9	14.7	.3442E 00	1.007	270	820	1.61	2.15
53	6.173	7.138	7.518	7.88	10.7	14.1	.2130E 01	1.044	330	820	1.66	2.04
54	7.007	7.577	7.892	8.20	10.7	14.0	.6137E 01	1.131	390	820	1.61	1.86
55	8.295	8.565	8.820	9.08	11.5	14.[illegible]	.1646E 02	1.394	470	820	1.36	1.49
56	3.295	7.811	8.532	9.18	13.6	18.8	.2640E-02	1.000	180	680	1.22	1.86
57	5.937	8.176	8.758	9.29	13.2	17.8	.2777E 00	1.006	220	680	1.33	1.79
58	7.363	8.575	9.041	9.48	12.9	17.1	.1930E 01	1.039	270	680	1.38	1.70
59	8.583	9.224	9.595	9.96	13.0	17.0	.6874E 01	1.148	330	680	1.32	1.52
60	10.01	10.33	10.64	10.96	13.8	17.8	.1651E 02	1.396	390	680	1.12	1.24

Table 1

FLTR	FREQ. (MHZ) AT AP,1,3,6,30 & 50 DB						R.C.	VSWR	C1,7	C3,5	L2,6	L4
NO.	AP-DB	1-DB	3-DB	6-DB	30DB	50DB	(%)	(MAX.)	(PF)	(PF)	(UH)	(UH)
	MEGAHERTZ											
1	1.027	.988	.958	.93	.73	.57	.1406E 02	1.327	2700	1500	5.43	4.89
2	1.217	1.108	1.061	1.02	.77	.59	.4710E 01	1.099	3300	1500	4.70	4.01
3	1.386	1.168	1.106	1.05	.77	.58	.1514E 01	1.031	3900	1500	4.63	3.71
4	1.707	1.218	1.136	1.07	.75	.56	.2330E 00	1.005	4700	1500	4.82	3.55
5	2.803	1.268	1.162	1.08	.73	.53	.4550E-02	1.000	5600	1500	5.19	3.44
6	1.307	1.253	1.213	1.18	.92	.71	.1281E 02	1.294	2200	1200	4.25	3.81
7	1.546	1.397	1.336	1.28	.97	.73	.4112E 01	1.086	2700	1200	3.74	3.16
8	1.823	1.480	1.395	1.32	.96	.72	.9490E 00	1.019	3300	1200	3.73	2.92
9	2.270	1.535	1.427	1.34	.94	.69	.1379E 00	1.003	3900	1200	3.91	2.82
10	5.346	1.605	1.464	1.36	.90	.65	.2063E-03	1.000	4700	1200	4.26	2.72
11	1.540	1.483	1.437	1.39	1.10	.85	.1406E 02	1.327	1800	1000	3.62	3.26
12	1.825	1.663	1.592	1.53	1.16	.88	.4710E 01	1.099	2200	1000	3.14	2.67
13	2.149	1.768	1.669	1.59	1.15	.86	.1112E 01	1.022	2700	1000	3.10	2.45
14	2.806	1.848	1.715	1.61	1.12	.82	.1078E 00	1.002	3300	1000	3.28	2.34
15	6.052	1.924	1.755	1.63	1.09	.78	.3140E-03	1.000	3900	1000	3.54	2.27
16	1.909	1.831	1.773	1.72	1.35	1.04	.1295E 02	1.298	1500	820	2.91	2.61
17	2.222	2.026	1.939	1.86	1.41	1.07	.4776E 01	1.100	1800	820	2.57	2.19
18	2.605	2.153	2.033	1.93	1.41	1.05	.1174E 01	1.024	2200	820	2.54	2.01
19	3.406	2.252	2.091	1.96	1.37	1.00	.1119E 00	1.002	2700	820	2.68	1.92
20	18.02	2.368	2.152	1.99	1.32	.95	.5540E-06	1.000	3300	820	2.95	1.85
21	2.218	2.144	2.080	2.02	1.60	1.24	.1554E 02	1.368	1200	680	2.53	2.29
22	2.689	2.447	2.342	2.25	1.70	1.29	.4640E 01	1.097	1500	680	2.13	1.81
23	3.104	2.588	2.446	2.33	1.70	1.27	.1312E 01	1.027	1800	680	2.10	1.67
24	3.973	2.706	2.516	2.37	1.65	1.22	.1478E 00	1.003	2200	680	2.21	1.60
25	12.21	2.844	2.589	2.40	1.59	1.15	.3260E-04	1.000	2700	680	2.43	1.54
26	2.728	2.630	2.550	2.47	1.95	1.51	.1464E 02	1.343	1000	560	2.05	1.85
27	3.195	2.937	2.816	2.71	2.06	1.57	.5502E 01	1.116	1200	560	1.77	1.52
28	3.809	3.151	2.976	2.83	2.06	1.54	.1190E 01	1.024	1500	560	1.73	1.37
29	4.769	3.281	3.053	2.87	2.01	1.48	.1629E 00	1.003	1800	560	1.82	1.32
30	12.00	3.443	3.138	2.91	1.94	1.40	.1479E-03	1.000	2200	560	1.99	1.27
31	3.169	3.070	2.981	2.89	2.29	1.78	.1646E 02	1.394	820	470	1.77	1.61
32	3.787	3.489	3.347	3.22	2.45	1.87	.5730E 01	1.122	1000	470	1.49	1.28
33	4.355	3.711	3.517	3.35	2.46	1.85	.1742E 01	1.035	1200	470	1.45	1.17
34	5.614	3.904	3.634	3.42	2.40	1.77	.1806E 00	1.004	1500	470	1.52	1.11
35	10.67	4.074	3.724	3.46	2.32	1.68	.1235E-02	1.000	1800	470	1.65	1.07
36	3.817	3.697	3.590	3.49	2.76	2.15	.1651E 02	1.396	680	390	1.47	1.34
37	4.521	4.182	4.015	3.86	2.95	2.26	.6137E 01	1.131	820	390	1.24	1.07
38	5.266	4.476	4.240	4.04	2.97	2.23	.1687E 01	1.034	1000	390	1.20	.970
39	6.393	4.668	4.360	4.11	2.91	2.15	.2939E 00	1.006	1200	390	1.25	.926
40	13.35	4.915	4.490	4.17	2.79	2.02	.9369E-03	1.000	1500	390	1.37	.889
41	4.368	4.252	4.136	4.02	3.21	2.50	.1891E 02	1.466	560	330	1.30	1.19
42	5.261	4.895	4.706	4.53	3.48	2.66	.6874E 01	1.148	680	330	1.06	.924
43	6.065	5.244	4.979	4.75	3.51	2.65	.2130E 01	1.044	820	330	1.02	.830
44	7.399	5.499	5.142	4.85	3.44	2.55	.3533E 00	1.007	1000	330	1.05	.786
45	11.21	5.725	5.263	4.91	3.34	2.43	.1195E-01	1.000	1200	330	1.13	.761
46	5.502	5.332	5.178	5.03	3.99	3.10	.1665E 02	1.400	470	270	1.02	.930
47	6.462	6.002	5.767	5.55	4.25	3.25	.6629E 01	1.142	560	270	.867	.752
48	7.495	6.435	6.104	5.82	4.29	3.23	.1930E 01	1.039	680	270	.832	.676
49	9.070	6.724	6.287	5.93	4.21	3.11	.3442E 00	1.007	820	270	.859	.643
50	14.92	7.029	6.449	6.01	4.06	2.95	.6280E-02	1.000	1000	270	.931	.620
51	6.891	6.653	6.454	6.26	4.95	3.84	.1519E 02	1.358	390	220	.812	.735
52	8.114	7.466	7.160	6.88	5.24	4.00	.5600E 01	1.119	470	220	.697	.599
53	9.280	7.921	7.508	7.16	5.26	3.96	.1782E 01	1.036	560	220	.678	.548
54	11.41	8.283	7.733	7.29	5.15	3.80	.2777E 00	1.006	680	220	.704	.522
55	18.94	8.641	7.922	7.37	4.98	3.61	.4870E-02	1.000	820	220	.761	.505
56	8.717	8.357	8.091	7.84	6.15	4.76	.1280E 02	1.294	330	180	.637	.572
57	10.02	9.179	8.795	8.45	6.42	4.89	.5160E 01	1.109	390	180	.567	.486
58	11.58	9.742	9.217	8.77	6.42	4.82	.1466E 01	1.030	470	180	.556	.445
59	14.08	10.14	9.459	8.91	6.29	4.64	.2555E 00	1.005	560	180	.577	.427
60	25.12	10.60	9.702	9.02	6.07	4.40	.2640E-02	1.000	680	180	.627	.412

Table 2

requirements can be satisfied. The reflection coefficient percentages are listed in exponential form for ease in listing very small values; thus, R.C. = 3260E-04 percent (Design #21) means $.326 \times 10^{-4}$ percent. Of course, the 5 percent series could also have been used in calculating the tables, but the sixty designs that are derived from the more common 10 percent values is quite adequate.

Table 2 shows the HP filter tabulation, and the explanation of the LP column headings also applies to the HP table, except the capacitor groups of C1,7 and C3,5 are exchanged. Columns C1,7 and C3,5 start and end with values that are different from those in the LP table, and this was done so the HP 3-dB cutoff frequencies will occur in the 1-10 MHz decade similar to that for the LP table. Note that for the same capacitor ratio in the LP and HP tables the R.C. and max. VSWR values are identical, but the other tabulated values are quite different.

To use the tables, select the appropriate table (LP or HP) and find a listed 3-dB cutoff frequency nearest the desired cutoff frequency (assume the desired f_{co} is in the 1-10 MHz range). Note the attenuation response and VSWR, and if these parameters are acceptable, construct the filter in accordance with the appropriate schematic diagram using the listed component values. If the desired cutoff frequency is in a lower frequency decade, simply divide all the listed frequencies by a factor which is obtained by raising ten to the power which is equal to the number of decades below the 1-10 MHz decade. Multiply all component values by the same factor. For example, if the desired cutoff frequency is in the 10-100 kHz decade (two decades below the 1-10 MHz decade), the scaling factor is 10^2, or 100. Designs in the 10-100 MHz range are obtained by multiplying the frequency values by ten and dividing the component values by ten. (It is suggested that the tabulated designs not be used above 100 MHz because of the difficulty in obtaining the desired response from lumped L-C filters.) An example will demonstrate the application of a pre-calculated filter to a typical RF design problem.

Lowpass Filter Design Example

Receiver testing is conveniently performed at fixed frequencies using one or more inexpensive transistorized crystal oscillators to provide stable RF test signals. Unfortunately, the high levels of the oscillator harmonics sometimes makes the unfiltered oscillator output unsuitable for some applications such as for receiver intermodulation testing where two oscillators are required. For this application, the harmonic levels of each oscillator must be reduced by more than 55 dB if the intermodulation test results are to be accurate. This filtering requirement provides a perfect application for one of the

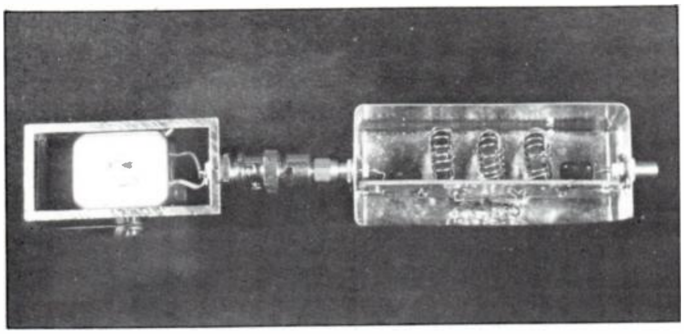

Figure 3. Crystal Oscillator and 7-Element Lowpass Filter, Design #10.

pre-calculated 50-ohm 7-element lowpass Chebyshev filters of Table 1.

Figure 3 is a photograph of an International Crystal Mfg. Co., Type OE-10 14.060 MHz oscillator and a lowpass filter that was assembled to demonstrate how the oscillator harmonics can be reduced to an acceptable level. Figure 4 shows the relative output of the unfiltered 14.06 MHz oscillator and its harmonics. Note that the second and third harmonic levels are only 3 and 8 dB below the fundamental level. Assuming that a harmonic level reduction of at least 55 dB is desired, it is obvious that a filter having a high reflection coefficient (for maximum rise of attenuation) and a cutoff frequency as near as possible to 14.06 MHz is required. Referring to Table 1, and after scaling to the next higher frequency decade, Filter No. 10 appears to be most suited for this application. This design has one of the highest values of R.C. (16.65 percent) and its attenuation is less than 1 dB below 14.98 MHz so the oscillator fundamental will not be significantly attenuated. The C1,7 and C3,5 capacitor values are 270 and 470 pF, respectively.

Filter No. 10 was constructed and its insertion loss (I.L.) was measured using the test setup shown in Figure 4, except the crystal oscillator was replaced by the Hewlett-Packard Tracking Generator, Model 8443A. The measured filter insertion loss is shown in Figure 5, and it compares favorably with the theoretical I.L. calculated from the data in Appendix A, Table A-3 (see data for Filter No. 10, 2nd line from bottom). The loss at the oscillator 2nd harmonic (28.1 MHz) was 57 dB which is quite adequate for this application. The filter I.L. was greater than 70 dB up to 100 MHz. Figure 6 shows the oscillator output level after filtering. The 2nd harmonic level is now 60 dB below the fundamental (a drop of 57 dB), and the 3rd harmonic (at 42.2 MHz) is undetectable.

In order to obtain insertion loss levels in excess of 55 dB above 30 MHz, it is important that all capacitors have very low impedance paths to the case ground. If any common ground impedance exists between the capacitors it will provide a coupling path between the filter input and output with a corresponding reduction in insertion loss. When the

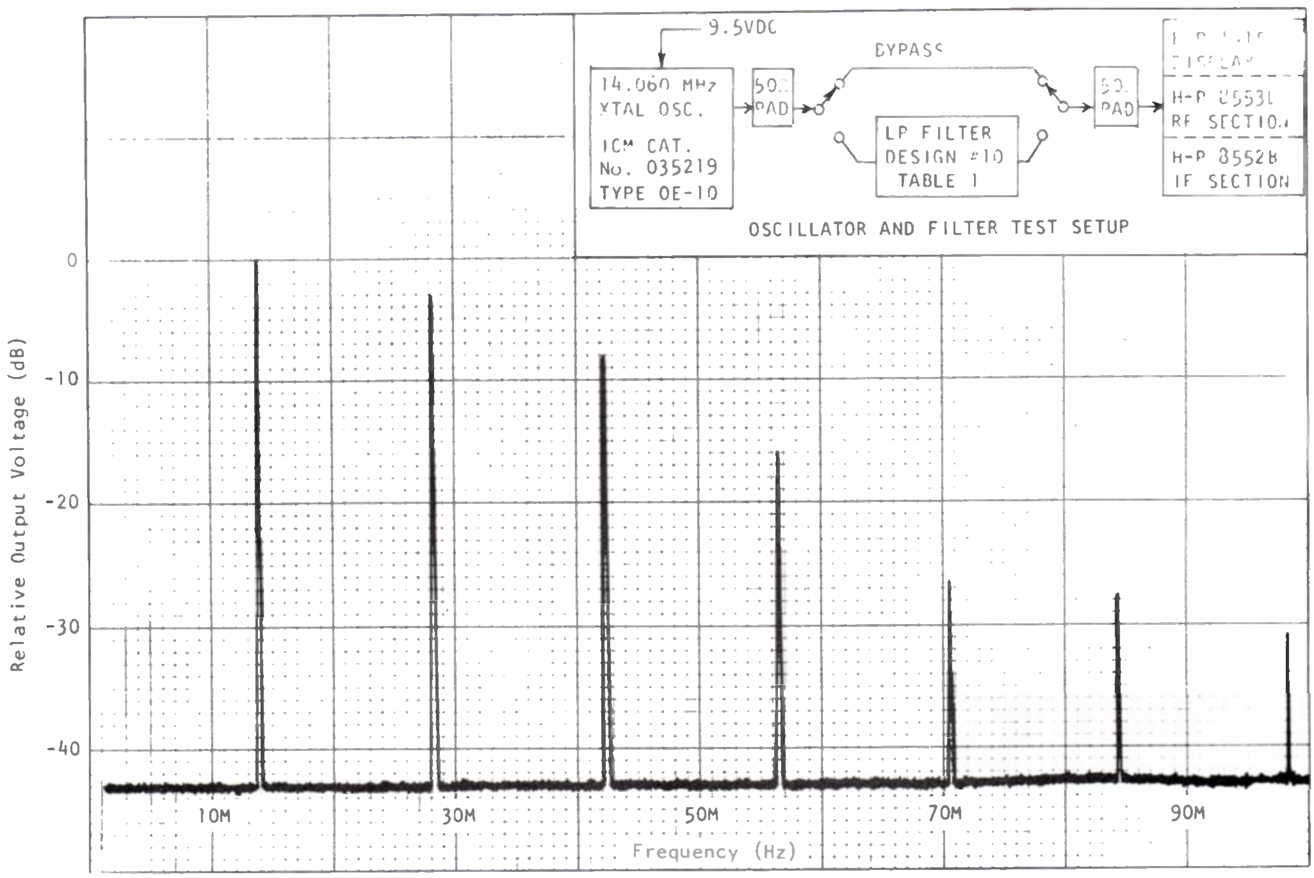

Figure 4. Output Levels

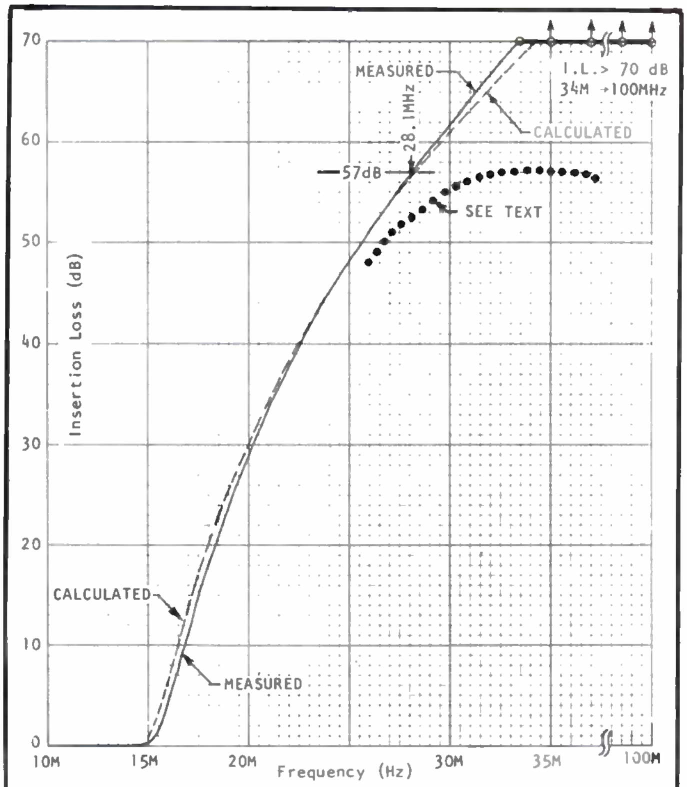

Figure 5.

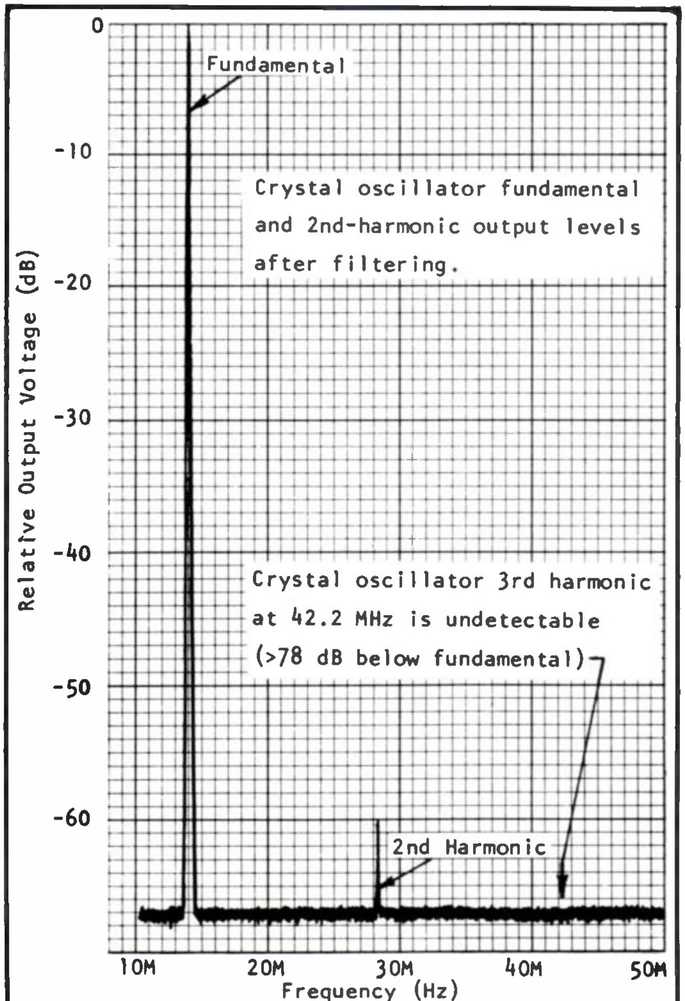

Figure 6.

filter was first constructed, the copper side of the single-sided p.c. board was used as the common ground plane for all four capacitors. With this arrangement, the I.L. above 25 MHz showed a significant drop-off from the theoretical I.L. curve (see dotted line, Figure 5). After eliminating the common ground paths and after providing C3 and C5 with separate low impedance grounding straps connected directly to the tinned-metal case, the expected insertion loss was obtained. The previously detectable 3rd harmonic became undetectable.

The measured insertion loss above 30 MHz (see Figure 5, solid line) is slightly greater than the theoretical loss, and this is no doubt due to the self-resonance of the inductors introducing an unintentional (but useful) parallel resonance somewhere above 50 MHz. The presence of this parallel resonance in one or more of the series inductors causes their impedance to increase considerably, and since they are in series with the signal path, the signal loss increases correspondingly. If more than 57 dB of attenuation is needed at the second harmonic, the parallel resonance of one of the series inductors (caused by the inductor winding capacity) can be lowered by the deliberate addition of capacitance across one of the inductors, thereby enhancing the I.L. at a lower frequency. If this is done, however, the passband ripple probably will be increased, but in this particular application it probably won't matter.

The filter was constructed with dipped mica and chip ceramic capacitors for C1,7 and C3,5, respectively. Polystyrene capacitors (470 pF) were originally used for C3,5, but were replaced with the chip ceramic capacitors in an attempt to improve the deficient I.L. response before the true cause (common ground impedance coupling) was found and corrected. The original polystyrene capacitors probably would have been satisfactory if their ground leads had been connected directly to the case. The .775 and .853 μH inductors were wound on Micrometals T68-10 toroidal cores with 14 and 15 turns, respectively, of #20 magnet wire. The Q of these inductors was about 165 at 14 MHz.

Most design difficulties concerning simple lowpass and highpass 50-ohm passive filters are eliminated by the use of the pre-calculated filters in Tables 1 and 2; however, the problems in realizing these designs still remain. Those not well acquainted with the selection and purchasing of filter components and the assembly and housing of the filter are advised to obtain a copy of the Measurements and Control Filter Series Home Study Course which is available for $12 prepaid from M&C, 2994 W. Liberty Ave., Pittsburgh, Pa. 15216. The subjects of passive and active filters, and component selection and test procedures are discussed in five separate courses which will be useful to those responsible for the occasional design and construction of filters.

Verification of Tabulated Designs

Whenever any computer-derived design data is published, such as shown in Tables 1 and 2, the reader should be provided the opportunity to independently verify the calculations associated with the tables. Because the computer performs all the calculations for each design in an identical manner, the verification of only one or two designs should suffice to establish the validity of the entire table. This is essentially true for Tables 1 and 2, except different calculations are used in determining the normalized component values for filters having reflection coefficients less than and greater than one percent.

Table A-1 is a listing of about half of the filters in Table 1, normalized to a 3-dB cutoff frequency of 1 radian/second and 1-ohm terminations. The "G" heading is used to distinguish the normalized tabulations from the scaled values. Also included are designs for the Butterworth (indicated by a single asterisk) and a 20 percent Chebyshev (indicated by a double asterisk). These two designs are included only for comparison purposes.

Table A-1 can be used to independently calculate and verify the correctness of the lowpass and highpass component values in Tables 1 and 2. For example, the lowpass design of No. 16, Table 1, can be checked by scaling the normalized data given in Table A-1 where G1,7 and G2,6 = .4482F and 1.2525H, respectively. The component values of design #16 will be calculated as a demonstration of the scaling process. The normalized values will be scaled to produce the lowpass design of filter No. 16, which has a 3-dB cutoff of 1.74 MHz. Since all the filters of Table 1 are designed for 50 ohms, the normalized values must be scaled from one to 50 ohms. The C_s and L_s scaling factors are calculated in the following manner:

$C_s = 1/R\omega$ where $R = 50$ and $\omega = 2\pi f_3 = 2\pi(1.74 \cdot 10^6) = 10.9327 \cdot 10^6$,

$C_s = 1/(546 \cdot 637 \cdot 10^6) = 1829.4 \cdot 10^{-12}$.

$L_s = R/\omega = 50/(10.9327 \cdot 10^6) = 4.5734 \cdot 10^{-6}$.

The normalized G-values are now scaled to the desired 3-dB cutoff frequency and 50-ohm level by multiplying the G-values by the appropriate scaling factors. Thus:

$C1,7 = G1(C_s) = .4482(1829.4)$pF = 819.9pF or 820pF;

$C3,5 = G3(C_s) = 1.8039('')$pF = 3300.05pF or 3300pF;

$L2,6 = G2(L_s) = 1.2525(4.5734)\mu$H = 5.728$\mu$H or 5.73$\mu$H;

$L4 = G4(L_s) = 1.9992('')\mu$H = 9.143$\mu$H or 9.14$\mu$H.

Note the exact agreement between the manually and computer-calculated scaled values of lowpass design #16 in Table 1. By applying the proper lowpass-to-highpass transformation and scaling procedure, the highpass designs in Table 2 may similarly be confirmed.

To confirm that the Chebyshev normalized data of Table A-1 are correct, the values for R.C. = 20 percent should be divided by the F3/F-Ap ratio to convert the G1-G7 values to a cutoff frequency of F-Ap. The converted values will then be seen to be identical with those values in previously published tables (such as in references 5,6,7,8 and 9) which are normalized for an F-Ap cutoff frequency of 1 rad/sec. For example, the G1,7 value of the 20 percent Chebyshev filter in Table A-1 is 1.4059. If this is divided by 1.0531 (the F3/F-Ap ratio), the new normalized G1,7 value is 1.335. Referring to Geffe's Simplified Modern Filter Design, Table A2-3, the normalized values for G1,7 will be seen to be identical.

Since tables for reflection coefficients less than one percent are not normally published, it is not possible to make a table-to-table comparison for those Chebyshev designs having a reflection coefficient less than 1 percent. However, it is possible by referring to Table A-1 to see that the G1-7 values approach the Butterworth values as a limit. This is to be expected as all the Chebyshev designs are normalized to the same cutoff frequency (1 rad/sec at 3 dB) as is the Butterworth. The first twelve designs in Table A-1 have such small reflection coefficients that, for all practical purposes, they provide a Butterworth response. The Ap value is so small that the computer calculation indicates it as zero. An alternate and less convenient method of checking those designs with less than 1 percent reflection coefficient is to use a network analysis program to computer-calculate the attenuation versus frequency response of one of the filters. This was done and the frequencies at the 1,2,6,30 and 50 dB levels of the calculated response were found to agree with the tabulated frequencies of Tables 1 and 2.

Of particular interest is the G3/G1 ratio for each design which was computer-calculated from the G3 and G1 values (see second last column, Table A-1). These ratios should be equal to the ratios of the C3 and C1 capacitor values for any given design. For example, design #10 in Table 1 has a C3/C1 ratio of 4700/2700 = 1.74074, and this is the ratio given for the same design in Table A-1. In a few cases, there may be an insignificant disagreement in the last decimal place, such as in designs, 42 and 57, and a few others.

Equations for Calculating Normalized Values

Table A-2 lists the program used to calculate the normalized values of Table A-1. A computer print-out is shown below Table A-2 for five different values of reflection coefficient. The first computer run (RC = .949 percent) is for lowpass design #18, and the values calculated correspond to those in Table A-1. The second and last runs (RC = 1 percent and 20 percent) are included to permit the reader to compare the values (after dividing by the F3/F-Ap ratio) with those published by Saal, Zverev and Geffe. Run #4 (RC = 15.0874 percent) is included to permit comparison with those published tables listing normalized values for Ap = .1 dB. For example, dividing G1 through G7 (for Ap = .1 dB) by 1.068 gives G1,7 = 1.81, G2,6 = 1.423, G3,5 = 2.097 and G4 = 1.573. These values are identical with those published in Reference #9, page 101, Table 4.4 for n = 7 and ripple (dB) = 1/10. The data obtained in Run #3 (RC = 5.735 percent) is used later in the article for a design example using 10 percent-tolerance Mylar capacitors.

This same program, with minor changes and additions, was used to calculate the exact value of reflection coefficient that corresponded to a given capacitor ratio. The program was changed so the desired capacitance ratio was specified as an input. A reflection coefficient, which was known to be less than that value which corresponded to the desired ratio, was entered into the program. An iterative loop was established which incremented the R.C. in very small steps until the calculated G3/G1 ratio equaled the desired C3/C1 ratio. At this point, the iterative calculations were stopped, the final R.C. value was printed by the computer and the next desired capacitor ratio was requested. This iterative calculation procedure was easier to use rather than finding the explicit equation relating the reflection coefficient to any given C-ratio.

The iterative calculation procedure is especially useful when one is forced to use 10 percent-tolerance capacitors. This might occur when constructing 50-ohm filters for cutoff frequencies below 100 kHz where the 2.5 percent polystyrene capacitors become prohibitively large and expensive. For example, if a 15 kHz lowpass filter is desired, design #10 of Table 1 can be constructed with 10 percent Mylar capacitors having nominal values of .27 and .47 μF. But, because of the 10 percent tolerance range of this capacitor type, a considerable error can result if a random selection of the nominal values is used. The following procedure is suggested to eliminate this problem. Measure about ten capacitors of each nominal value with a digital capacitance meter (the Data Precision Model 938 selling for $139 is highly recommended), and select four capacitors so that each pair of the selected nominal values are within 2 percent of a common value. Determine the new C-ratio and find the corresponding reflection coefficient and normalized component values using the Table A-2 program with an iterative loop. By reducing the iterative R.C. loop increment, the R.C. value can be found to any desired accuracy.

For example, assume that two capacitors of the .27 μF group are within 1 percent of a common value of .243 μF, and two capacitors of the .47 μF group are within 1 percent of a common value of .517 μF. Normally, the .47/.27 ratio would be 1.7407, but because the extremes of the 10 percent tolerance range were assumed in this example, the new C-ratio of the selected pairs is 2.12757, which is significantly different than the original ratio. Using the Table A-2 program and an iterative loop of .001 percent, the R.C. corresponding to a C-ratio of 2.12757 is 5.735 percent. The normalized values are then calculated, and the new cutoff frequency (based on the new capacitor values) is determined. The new design will then exactly match the values of the capacitors selected. But, before proceeding any further, the new cutoff frequency should be calculated because it no longer may be satisfactory. Using the normalized G1 value associated with the R.C. value of 5.735 percent, calculate the new 3-dB cutoff frequency using the following equation:

$F_3 = G1/(100\pi \cdot C1)$ where $C1 = .243 \cdot 10^{-6}$ and $G1 = .9512$

$F_3 = .9512/(100\pi \cdot .243 \cdot 10^{-6}) = 12.46$ kHz.

The new 3-dB cutoff frequency is 2.96 kHz lower than the original 15.42 kHz cutoff frequency, and if this is unacceptable, then a new group of capacitors must be selected.

Calculation of Stopband Attenuation Versus Normalized Frequency

Table A-3 lists the stopband attenuation versus normalized frequencies for 18 designs selected from Table 1. This data permits the calculation of 13 values of attenuation versus frequency for any of the 18 listed filter designs. The table was used to calculate the dashed curve for the insertion loss response for filter design #10 shown in Figure 5. In this case, the 3-dB fco was 15.42 MHz (after frequency scaling design #10, Table 1) and the stopband attenuation at 1.3(15.42) = 20.0 MHz was 30.1 dB. Similarly, for 1.4(15.42) = 21.6 MHz, 1.5(15.42) = 23.1 MHz, and 1.6(15.42) = 24.7 MHz, the corresponding attenuation levels were 36.4, 41.9 and 46.8 dB respectively.

The discussion and tabulated data contained in this article should be of considerable long-term use to those RF engineers having an occasional need to design and construct simple passive LC filters.

Designing for Any Impedance Level

Although the 50-ohm impedance level is most frequently used by RF designers, there are several other levels that are also used. In the video and RF range, 75, 93 and 300 ohms are occasionally encountered, and 600 ohms is very common in the audio range. When a filter is needed for an odd impedance level, it would be advantageous if the design could be realized with only standard value capacitors as was demonstrated with the 50-ohm filter tables. Because of limited space, it is not practical to list separate filter tables for these less common values; however, by using a simple impedance scaling procedure, standard-value capacitor designs can be calculated for any desired impedance level within 5 percent of any cutoff frequency.

Tables 1, 2 and A-1 are used in the calculation of the new values, and Table A-3 is used to estimate the response. A lowpass and highpass design example will illustrate the procedure:

1) Select the filter type (lowpass or highpass) and the termination resistance, R. If R/50>10, or F_3 is outside the 1-10 MHz decade, see design example #2.

2) Select the desired 3-dB cutoff frequency, F_3.

3) Calculate the resistance ratio, R_r, relative to 50 ohms: $R_r = R/50$.

4) Calculate the desired 3-dB cutoff frequency when scaled to 50 ohms: $F_3^{50} = R_r(F_3)$.

5) Refer to Table 1 (or 2) to find the design that is closest to F_3^{50}. This is the design that will have a cutoff frequency within ±5 percent of F_3 and can be constructed with the listed standard-value capacitors for the desired resistance terminations.

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6) Using the listed value of C3 (from Table 1 or 2) and the normalized value of G3 (from Table A-1), calculate the exact F_3 frequency of the selected design.

7) Calculate C1 and the inductor values. This completes the design.

For example, assume a 75-ohm lowpass filter is desired with a 3-dB cutoff frequency of 3.00 MHz. Find a design having a cutoff frequency within about ±5 percent of 3.0 MHz (±.15 MHz). Thus, R = 75 ohms, F_3 = 3.0 MHz and $R_r = 75/50 = 1.5$.

From step (4), $F_3^{50} = R_r F_3 = 1.5(3.0)$ MHz = 4.5 MHz. Referring to Table 1 (lowpass), the closest design to F_3^{50} = 4.5 MHz is design #40 at F_3#40 = 4.645 MHz, with C1,7 = 820 pF and C3,5 = 1500 pF.

From step (6), calculate the exact value of F_3^{75}.

$F_3^{75} = G3/(2\pi R \cdot C3)$ where $G3^{\#40} = 2.1890$.

$F_3^{75} = 2.189/(2\pi \cdot 75 \cdot 1500 \cdot 10^{-12}) = 3.0968$ MHz. Therefore, the exact F_3^{75} is 3.0968 MHz for C3,5 = 1500 pF. Complete the filter calculation by determining the L_s and C_s scaling factors and calculating the L and C values.

$L_s = R/\omega = 75/(2\pi \cdot 3.0968 \cdot 10^6) = 3.8545 \cdot 10^{-6}$

$C_s = 1/R\omega = 1/(75 \cdot 2\pi \cdot 3.0968 \cdot 10^6) = 685.2 \cdot 10^6) = 685.2 \cdot 10^{-12}$

$C1,7 = G1 \cdot C_s$ where G1 = 1.1967 (from Table A-1, LPF #40) = $1.1967(685.2) \cdot 10^{-12}$ = 820 pF

$C3,5 = G3 \cdot C_s = 2.1890(685.2) \cdot 10^{-12}$ = 1500 pF

$L2,6 = G2 \cdot L_s$ where G2 = 1.5423 and G4 = 1.7182

$L2,6 = 1.5423(3.8545) \cdot 10^{-6} = 5.94\ \mu H$

$L4 = 1.7182(3.8545) \cdot 10^{-6} = 6.62\ \mu H$

To summarize the final design: R = 75 ohms, F_3 = 3.0968 MHz, R.C. = 12.95 percent (based on LPF Design #40), C1,7 = 820 pF, C3,5 = 1500 pF, L2,6 = 5.94 μH and L4 = 6.62 μH.

Design Example #2 for R/50>10, or When F_3 is Outside the 1-10 MHz Decade

If R is such that R/50>10, then divide R by ten and proceed with the design. After the calculations have been completed, the component values are scaled from the impedance level of R/10 to R by multiplying all inductor values by ten and by dividing all capacitor values by ten.

If F_3 is outside the 1-10 MHz decade, scale F_3 to fall within the 1-10 MHz decade by multiplying F_3 by ten raised to an integral power. After calculating the component values, scale them to the desired F_3 by multiplying them by the same factor.

For example, design a 600-ohm, 4-kHz lowpass filter: F_3 = 4 kHz and R = 600 ohms. Since 600/50>10, divide 600 ohms by ten: 600 ohms/10 = 60 ohms. Since 4 kHz is outside the 1-10 MHz range, it is scaled to fall within the 1-10 MHz range by multiplying it by 10^3. Thus, 4 kHz(10^3) = 4 MHz. A 60-ohm, 4 MHz lowpass filter will be calculated and then scaled to the desired 600 ohms and 4 kHz cutoff frequency. $R_r = 60/50 = 1.2$, $F_3^{50} = 1.2(4\text{ MHz}) = 4.8$ MHz. The closest lowpass filter F_3 from Table 1 is Design #41, F_3^{50} = 4.861 MHz and C3 = 1200 pF. From Table A-1, $G3^{\#41}$ = 1.8327. $F_3^{60} = G3/(2\pi \cdot R \cdot C3) = 1.8327/(2\pi \cdot 60 \cdot 1200 \cdot 10^{-6}) = 4.051156$ MHz. [After impedance scaling, F_3^{600} = 4.051 kHz.] Calculate the L_s and C_s scaling factors: $L_s = R/\omega = 60(2\pi \cdot F_3^{60}) = 60(2\pi \cdot 4.051156 \cdot 10^6) = 2.3572 \cdot 10^{-6}$
$C_s = 1/R\omega = 1/(60 \cdot 2\pi \cdot F_3^{60}) = 654.77 \cdot 10^{-6}$

C1,7 = .5040(654.77)pF = 330 pF
C3,5 = 1.8327(654.77)pF = 1200 pF
L2,6 = 1.3409(2.3573)μH = 3.161 μH
L4 = 1.9881(2.3572)μH = 4.686 μH

Scaled from 60 to 600 ohms	*Scaled from 600 ohms & 4.05 MHz to 4.05 kHz
33 pF	33nF = .033 μF
120 pF	120nF = .12 μF
31.61 μH	31.61 mH
46.86 μH	46.86 mH

Highpass Design Example For Any Impedance Level

The highpass design procedure is similar to that of the lowpass except that after obtaining the normalized values from Table A-1, they are transformed into highpass values so the component value calculations will be similar to those used in the lowpass procedure.

Assume the highpass filter specifications are the same as for the lowpass design example; that is, R = 75 ohms and F_3 = 3.00 MHz. Then R_r = 1.5 and F_3^{50} = 4.5 MHz. From Table 2 (Highpass), the design having a 3-dB cutoff frequency closest to 4.5 MHz is #40 with F_3^{50} = 4.490 MHz. [Note that although the filter number and the larger capacitor value are the same as in the previous lowpass design example, the R.C. is different.] From Table 2 (Highpass) #40 design listing, C1,7 = 1500 pF and C3,5 = 390 pF.

*Final scaled component values for a 600-ohm 4.051 kHz lowpass filter. The desired 3-dB cutoff frequency was 4.0 kHz, but the calculated value will be acceptable to allow the use of standard-value capacitors.

Using the highpass design R.C. value (.9369E-3) as a guide, find the normalized values for this design from Table A-1 (see LPF No. 36 in Table A-1). These listed lowpass normalized component values are transformed into highpass values by calculating their reciprocals. Thus, the highpass normalized values of LPF #36 are: G1,7 = 2.1160, G2,6 = .77333, G3,5 = .550176, and G4 = .50153. As in the lowpass design example, the value of C3 and G3 will be used to first calculate the exact cutoff frequency. The L_s and C_s scaling factors will then be determined and all the highpass filter component values will be calculated.

From step (6) of the lowpass design example, $F_3{}^{75} = G3/(2\pi \cdot R \cdot C3) = .550176/(2\pi \cdot 75 \cdot 390) = 2.993615$ MHz $= F_3{}^{75}$.

$L_s = R/\omega = 75/2\pi \cdot F_3{}^{75} = 3.987 \cdot 10^{-6}$

$C_s = 1/R\omega = 1/(75 \cdot 2\pi \cdot F_3{}^{75}) = 708.86 \cdot 10^{-12}$

C1,7 = G1$\cdot C_s$ = 1500 pF; C3,5 = G3$\cdot C_s$ = 390 pF; L2,6 = G2$\cdot L_s$ = 3.083 μH; L4 = G4$\cdot L_s$ = 1.996 μH.

Appendix A

Table A-1. Element Values for 7-Element Chebyshev Lowpass Filters, Normalized for a 3-dB Cutoff frequency of 1 rad/sec and 1-ohm Terminations.

LPF NO.	R.C. (%)	A-PEAK (DB)	F3/F-AP (RATIO)	G1,7 (F)	G2,6 (H)	G3,5 (F)	G 4 (H)	G3/G1 RATIO	LPF NO
*	ZERO	ZERO	N.A.	.4450	1.2470	1.8019	2.0000	4.0489	*
16	.55400E-06	.0000E 00	8.3758	.4482	1.2525	1.8039	1.9992	4.0244	16
21	.32600E-04	.0000E 00	4.7165	.4554	1.2648	1.8081	1.9975	3.9706	21
26	.14790E-03	.0000E 00	3.8235	.4611	1.2743	1.8114	1.9963	3.9286	26
6	.20630E-03	.0000E 00	3.6523	.4627	1.2770	1.8123	1.9959	3.9167	6
11	.31400E-03	.0000E 00	3.4480	.4650	1.2808	1.8136	1.9954	3.9000	11
36	.93690E-03	.0000E 00	2.9727	.4726	1.2931	1.8176	1.9939	3.8462	36
31	.12356E-02	.0000E 00	2.8643	.4744	1.2968	1.8188	1.9934	3.8298	31
56	.26400E-02	.0000E 00	2.5895	.4825	1.3087	1.8226	1.9920	3.7778	56
1	.45500E-02	.0000E 00	2.4114	.4891	1.3188	1.8259	1.9907	3.7333	1
51	.48700E-02	.0000E 00	2.3902	.4900	1.3202	1.8263	1.9906	3.7273	51
46	.62800E-02	.0000E 00	2.3129	.4936	1.3255	1.8280	1.9899	3.7037	46
41	.11950E-01	.0000E 00	2.1307	.5040	1.3409	1.8327	1.9881	3.6364	41
17	.10782E 00	.5177E-05	1.6358	.5624	1.4166	1.8558	1.9785	3.3000	17
22	.11190E 00	.5695E-05	1.6289	.5638	1.4182	1.8563	1.9783	3.2926	22
12	.13782E 00	.8283E-05	1.5914	.5721	1.4276	1.8593	1.9769	3.2500	12
27	.14780E 00	.9319E-05	1.5791	.5750	1.4309	1.8604	1.9764	3.2352	27
32	.16290E 00	.1139E-04	1.5622	.5793	1.4355	1.8619	1.9757	3.2142	32
37	.18060E 00	.1398E-04	1.5447	.5839	1.4404	1.8635	1.9749	3.1915	37
7	.23277E 00	.2330E-04	1.5030	.5961	1.4530	1.8678	1.9727	3.1333	7
2	.25550E 00	.2847E-04	1.4882	.6009	1.4578	1.8694	1.9719	3.1111	2
57	.27770E 00	.3365E-04	1.4752	.6053	1.4621	1.8710	1.9710	3.0908	57
42	.29390E 00	.3728E-04	1.4665	.6084	1.4651	1.8720	1.9705	3.0768	42
52	.34420E 00	.5125E-04	1.4427	.6174	1.4735	1.8752	1.9687	3.0370	52
47	.35330E 00	.5436E-04	1.4388	.6190	1.4749	1.8757	1.9684	3.0303	47
18	.94900E 00	.3909E-03	1.3066	.6914	1.5301	1.9014	1.9519	2.7500	18
23	.11129E 01	.5379E-03	1.2878	.7063	1.5390	1.9070	1.9479	2.7001	23
28	.11746E 01	.5990E-03	1.2816	.7115	1.5419	1.9090	1.9465	2.6829	28
38	.11900E 01	.6151E-03	1.2801	.7128	1.5427	1.9095	1.9461	2.6788	38
33	.13125E 01	.7482E-03	1.2691	.7228	1.5480	1.9134	1.9432	2.6471	33
8	.14651E 01	.9325E-03	1.2569	.7346	1.5539	1.9181	1.9398	2.6110	8
13	.15147E 01	.9962E-03	1.2533	.7382	1.5555	1.9194	1.9386	2.6001	13
48	.16864E 01	.1235E-02	1.2419	.7506	1.5611	1.9246	1.9348	2.5641	48
43	.17423E 01	.1318E-02	1.2385	.7543	1.5625	1.9259	1.9334	2.5534	43
3	.17820E 01	.1379E-02	1.2361	.7571	1.5638	1.9272	1.9326	2.5455	3
58	.19300E 01	.1618E-02	1.2279	.7669	1.5676	1.9313	1.9294	2.5184	58
53	.21274E 01	.1966E-02	1.2181	.7795	1.5721	1.9368	1.9251	2.4847	53
24	.41112E 01	.7346E-02	1.1579	.8827	1.5932	1.9860	1.8840	2.2500	24
39	.46375E 01	.9350E-02	1.1480	.9060	1.5945	1.9984	1.8735	2.2058	39
19	.47119E 01	.9653E-02	1.1467	.9092	1.5946	2.0002	1.8720	2.2000	19
29	.47119E 01	.9653E-02	1.1467	.9092	1.5946	2.0002	1.8720	2.2000	29
34	.47760E 01	.9918E-02	1.1456	.9119	1.5947	2.0016	1.8708	2.1951	34
14	.51600E 01	.1158E-01	1.1395	.9280	1.5948	2.0105	1.8632	2.1666	14
44	.55020E 01	.1317E-01	1.1345	.9419	1.5945	2.0184	1.8565	2.1429	44
9	.56000E 01	.1364E-01	1.1331	.9458	1.5943	2.0207	1.8546	2.1364	9
49	.57345E 01	.1431E-01	1.1313	.9512	1.5941	2.0238	1.8520	2.1276	49
54	.61370E 01	.1639E-01	1.1262	.9669	1.5931	2.0330	1.8442	2.1026	54
4	.66290E 01	.1913E-01	1.1206	.9856	1.5914	2.0443	1.8348	2.0741	4
59	.68746E 01	.2057E-01	1.1179	.9948	1.5904	2.0499	1.8301	2.0606	59
20	.12804E 02	.7178E-01	1.0774	1.1922	1.5437	2.1856	1.7208	1.8333	20
30	.12804E 02	.7178E-01	1.0774	1.1922	1.5437	2.1856	1.7208	1.8333	30
40	.12950E 02	.7345E-01	1.0767	1.1967	1.5423	2.1890	1.7182	1.8292	40
25	.14060E 02	.8671E-01	1.0720	1.2306	1.5307	2.2150	1.6985	1.8000	25
35	.14060E 02	.8671E-01	1.0720	1.2306	1.5307	2.2150	1.6985	1.8000	35
50	.14640E 02	.9409E-01	1.0697	1.2481	1.5245	2.2286	1.6883	1.7857	50
15	.15190E 02	.1014E 00	1.0676	1.2646	1.5184	2.2417	1.6786	1.7727	15
45	.15540E 02	.1062E 00	1.0664	1.2750	1.5145	2.2500	1.6725	1.7647	45
55	.16460E 02	.1193E 00	1.0632	1.3023	1.5041	2.2721	1.6565	1.7447	55
60	.16512E 02	.1201E 00	1.0631	1.3038	1.5036	2.2733	1.6556	1.7436	60
10	.16650E 02	.1221E 00	1.0626	1.3079	1.5020	2.2766	1.6532	1.7407	10
5	.18913E 02	.1582E 00	1.0560	1.3742	1.4754	2.3320	1.6143	1.6969	5
**	.20000E 02	.1773E 00	1.0531	1.4059	1.4623	2.3591	1.5959	1.6781	**

NOTES:

*This line gives the Butterworth element values. The G3/G1 Butterworth ratio cannot be exactly realized with the standard-capacitor values, and the listed values and the G3/G1 ratio are included only for comparison purposes.

**This line lists the 20 percent Chebyshev design parameters to demonstrate that all the above listed Chebyshev element values (normalized for a 3-dB cutoff frequency of 1 rad/sec) can be transformed into the commonly published values (normalized for an F-Ap cutoff frequency of 1 rad/sec) by dividing each element value by the listed F3/F-Ap ratio. For example, if the RC = 20 percent G1-7 element values are divided by 1.0531, the results will be identical with the corresponding values listed in Table A2-3 of Philip Geffe's book, *Simplified Modern Filter Design.*

Table A-2. Basic Program for Calculating Element Values Normalized for a 3-dB Cutoff Frequency and One Ohm Terminations, with Examples.

```
10 *THIS BASIC PROGRAM CALCULATES THE ELEMENT VALUES (NORMALIZED
20 *TO A 3-DB CUTOFF FREQUENCY OF 1 RAD/SEC AND TERMINATIONS OF ONE OHM)
30 *AND RELATED PARAMETERS OF A 7-ELEMENT CHEBYSHEV LOWPASS FILTER FOR
40 *ANY GIVEN REFLECTION COEFFICIENT (%) GREATER THAN ZERO.
50 PRINT "%RC?"
60 INPUT R
70 PRINT"          R.C.        A-PEAK   F3/F-AP   G1,7   G2, 6   G3, 5";
80 PRINT"    G 4     G3/G1  "
90 PRINT"          (%)            (DB)     (RATIO)    (F)      (H)      (F)";
100 PRINT"     (H)     RATIO  "
110 A0=-4.3429*LOG(1-(R/100)^2)     'CALC. OF A-PEAK (DB). SEE NOTE LN 190'
120 E = SQR((.01*R)^2/(1-(.01*R)^2))                'CALC. OF RIPPLE FACTOR, E'
130 M=(1/7)*LOG(1/E+SQR(1/E^2-1)) 'PART CALC. OF F3/F-AP (SEE NOTE LN 190'
140 D = (EXP(M) + EXP(-M))/2                  'FINAL CALC. OF F3/F-AP RATIO'
150 IF R<1.0 GO TO 190          'NOTE: SPECIAL EQUATION REQUIRED FOR R<1.0'
160 B=A0/17.37178
170 X = LOG((EXP(B) + EXP(-B))/(EXP(B) - EXP(-B)))        (SEE NOTE LN 190)'
180 GO TO 200
190 X = 2*LOG(200/R)                                NOTE: LOG IS TO THE BASE E'
200 Y=.5*(EXP(.071429*X)-EXP(-.071429*X))
210 G1=.44504/Y
220 G2=.554956/((Y^2+.188255)*G1)
230 G3=2.24698/((Y^2+.611261)*G2)
240 G4=3.603876/((Y^2+.95048)*G3)
250 PRINT,260, R, A0, D, D*G1, D*G2, D*G3, D*G4, G3/G1
260 FMT X4,  E12.6, E11.4, F8.4,X1,  F7.4, F8.4, F8.4, F8.4, F8.4
270 GO TO 50


?BASIC A2
RUN
%RC?
!.949
       R.C.        A-PEAK   F3/F-AP   G1,7   G2, 6   G3, 5    G 4    G3/G1
       (%)          (DB)    (RATIO)   (F)     (H)     (F)     (H)    RATIO
   .949000E 00  .3909E-03  1.3066   .6914  1.5301  1.9014  1.9519  2.7500
%RC?
!1.000
       R.C.        A-PEAK   F3/F-AP   G1,7   G2, 6   G3, 5    G 4    G3/G1
       (%)          (DB)    (RATIO)   (F)     (H)     (F)     (H)    RATIO
   .100000E 01  .4344E-03  1.3004   .6965  1.5334  1.9036  1.9510  2.7332
%RC?
!5.735
       R.C.        A-PEAK   F3/F-AP   G1,7   G2, 6   G3, 5    G 4    G3/G1
       (%)          (DB)    (RATIO)   (F)     (H)     (F)     (H)    RATIO
   .573500E 01  .1431E-01  1.1313   .9512  1.5941  2.0238  1.8520  2.1276
%RC?
!15.0874
       R.C.        A-PEAK   F3/F-AP   G1,7   G2, 6   G3, 5    G 4    G3/G1
       (%)          (DB)    (RATIO)   (F)     (H)     (F)     (H)    RATIO
   .150874E 02  .1000E 00  1.0680  1.2615  1.5196  2.2392  1.6804  1.7751
%RC?
!20.000
       R.C.        A-PEAK   F3/F-AP   G1,7   G2, 6   G3, 5    G 4    G3/G1
       (%)          (DB)    (RATIO)   (F)     (H)     (F)     (H)    RATIO
   .200000E 02  .1773E 00  1.0531  1.4059  1.4623  2.3591  1.5959  1.6781
```

Table A-3. Lowpass Filter Stopband Attenuation vs. Frequency Normalized for a 3-dB Cutoff Frequency and One Ohm Terminations.

FLTR NO.	R.C. (%)	STOPBAND ATTENUATION (F=3DB FREQ.) 1.1F	1.2F	1.3F	1.4F	1.5F	1.6F	1.7F	1.8F	1.9F	2.0F	2.1F	2.2F	2.3F
		DECIBELS												
16	.554E-06	6.8	11.5	16.1	20.6	24.8	28.7	32.4	35.9	39.2	42.3	45.3	48.1	50.8
26	.148E-03	7.0	11.7	16.5	21.0	25.3	29.2	33.0	36.5	39.8	42.9	45.9	48.8	51.5
51	.487E-02	7.2	12.3	17.2	21.9	26.3	30.4	34.2	37.8	41.1	44.3	47.4	50.2	53.0
41	.120E-01	7.4	12.5	17.6	22.4	26.8	30.9	34.7	38.3	41.7	45.0	48.0	50.9	53.7
17	.108E 00	7.9	13.6	19.0	24.0	28.7	32.9	36.9	40.6	44.1	47.4	50.5	53.4	56.2
7	.233E 00	8.2	14.2	19.8	24.9	29.7	34.0	38.0	41.8	45.3	48.6	51.8	54.7	57.6
52	.344E 00	8.4	14.5	20.3	25.5	30.3	34.6	38.7	42.5	46.0	49.4	52.5	55.5	58.4
18	.949E 00	9.1	15.7	21.8	27.2	32.2	36.7	40.8	44.7	48.3	51.7	54.9	57.9	60.8
8	.147E 01	9.4	16.4	22.6	28.2	33.2	37.8	42.0	45.9	49.5	52.9	56.1	59.2	62.1
3	.178E 01	9.6	16.7	23.0	28.6	33.7	38.3	42.5	46.4	50.1	53.5	56.7	59.8	62.7
53	.213E 01	9.8	17.0	23.4	29.1	34.2	38.8	43.0	47.0	50.6	54.1	57.3	60.4	63.3
24	.411E 01	10.7	18.4	25.1	30.9	36.2	40.9	45.2	49.2	52.9	56.4	59.7	62.8	65.7
14	.516E 01	11.0	18.9	25.7	31.7	36.9	41.7	46.1	50.1	53.8	57.3	60.6	63.7	66.6
54	.614E 01	11.3	19.4	26.3	32.3	37.6	42.4	46.7	50.8	54.5	58.0	61.3	64.4	67.4
40	.130E 02	12.9	21.7	29.0	35.2	40.7	45.6	50.0	54.1	58.0	61.5	64.8	68.0	70.9
50	.146E 02	13.2	22.2	29.5	35.8	41.3	46.2	50.7	54.8	58.6	62.1	65.5	68.6	71.6
10	.167E 02	13.6	22.7	30.1	36.4	41.9	46.8	51.3	55.4	59.3	62.8	66.2	69.3	72.3
5	.189E 02	14.0	23.2	30.7	37.0	42.5	47.5	52.0	56.1	60.0	63.5	66.9	70.0	73.0

NOTES:

1. For highpass filter attenuation, the column headings are: F/1.1 . . . F/2.3.
2. The attenuation levels versus normalized frequency are listed in order of increasing value of R.C. for a selection of 18 designs. To estimate the stopband attenuation of the designs in Tables 1 or 2, find a listing above having an R.C. value nearest that of the selected design. The listed attenuation values will be within 1.5 dB of the correct value.

Appendix B

Definitions and equations related to the calculations discussed in this article.

Definitions

A_p — peak amplitude (dB) of the passband attenuation ripple.
A_s — stopband attenuation (dB).
F_{Ap} — frequency (Hz) where the passband attenuation level first exceeds the A_p level (denotes the end of the passband).
F_{AS} — frequency (Hz) corresponding to a stopband attenuation level of A_s.
R.C. — reflection coefficient (%).
ρ — absolute value of R.C. in decimal form (used in following equations).
ε — ripple factor, a parameter <1 related to the ripple amplitude.
n — number of branches in a ladder network (equal to number of reactive elements in the filters discussed in this article).
C — Chebyshev polynomial.
G1-G7 — normalized element values (1-7).
Ω_s = F-A_s/F_3 (for lowpass filter)
= F_3/F-A_s (for highpass filter)
F-A_s = filter stopband frequency (in Hz) at A_s.
F_3 = filter 3-dB cutoff frequency (in Hz).

Equations

(1) $\rho = (1 - .1^{.1Ap})^{.5}$
(2a) $Ap = -10 \cdot \log(1 - \rho^2)\,dB$
(2b) $Ap = 10 \cdot \log(1 + \varepsilon^2)\,dB$
(3a) $\varepsilon = (10^{.1Ap} - 1)^{.5}$
(3b) $\varepsilon = \rho/(1 - \rho^2)^{.5}$
(4) $VSWR = (1 + \rho)/(1 - \rho)$

Equations used in the calculation of the stopband attenuation (A_s) of a seven element lowpass or highpass filter.

(5) $A_{s(\Omega)} = 10 \cdot \log[1 + (\varepsilon \cdot C_{n(\Omega)})^2]\,dB$
where $A_{s(\Omega)}$ = *stopband attenuation in dB at* Ω,
ε = *ripple factor [from Eq.(3)],*
$C_{n(\Omega)}$ = *value of Chebyshev polynomial for "n" elements as a function of* Ω *where:*
(6) $C_{7(\Omega)} = 64\Omega^7 - 112\Omega^5 + 56\Omega^3 - 7\Omega.$
Calculation of F_3/F_{Ap} ratio:
(7) $F_3/F_{Ap} = \cosh[(\cosh^{-1} 1/\varepsilon)/n]$
Hyperbolic functions in terms of natural logs and exponents:
$\cosh^{-1}x = \ln[x + (x^2 - 1)^{.5}]$,
$\cosh y = .5[e^y + e^{-y}]$. $e = 2.718282$.

*Ω is the symbol for normalized frequency.

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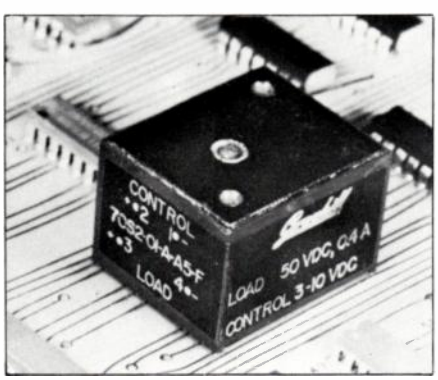
Contact Wiltron Company, 825 East Middlefield Road, Mountain View, Calif. 94043. INFO/CARD #140.

DC To DC Solid State Relay

A new family of solid state relays from Grayhill, Inc., switches a DC load with a DC input. These relays will operate over a wide load range of 3 to 50 VDC. Two package sizes with load current maximums of 400 MA and 2 amperes are offered. The inputs are logic compatible and are sensitive to 3 to 24 VDC with low drive currents. Optical isolation provides maximum protection for the driving logic circuitry. A clamped output provides maximum transient protection for the relay.



Size is another outstanding feature of this new family of relays. They have been housed in Grayhill's micro cube and mini cube packages. The micro cube measures 1" x 1.25" by less than 1/2" high. The mini cube package is 1" x 1.25" x 0.850" high. These standard packages are PC mountable with DIP compatible termiation. The larger mini cube package is also available with plug-in terminals that are accommodated in standard relay sockets.

The price of these DC to DC relays ranges from $7.50 to $9.00 each in 100 piece lots, depending upon the current range requirements and termination. Prototypes are available within two weeks.

Contact Grayhill, Inc., 561 Hillgrove Avenue, La Grange, Illinois 60525. INFO/CARD #139.

RFI and EMI Shielding

A highly conductive silver epoxy paint system for use in electronic systems as RFI/EMI shielding and static protection has been introduced by Carroll Coatings of Providence, Rhode Island.

Carroll Coatings C-608 is a conductive silver epoxy resin polyamide paint system that offers a volume resistivity of 0.01 + ohm/cm^3 in a uniform, 1 mil thick film. Easily applied by conventional paint sprayers or by brush, the forumulation resists settling and needs minimal agitation. Cured by air drying (accelerated with a low temperature bake if necessary), it has a 1 hour tack-free time, and a 12 hour pot life at 70°F.

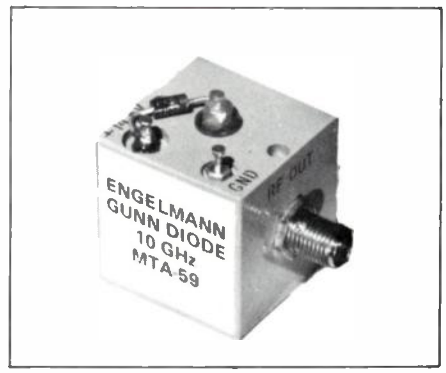
Meeting the qualitative requirements of MIL-C-22750C, Carroll Coatings C-608 adheres to a wide variety of substrates including aluminum, magnesium, polyester, acrylic, chrome, steel, bronze, brass and MIL-P-23377 zinc chromate epoxy primer. C-608 is currently used in numerous airborne, land- and sea-based military communications and weapons system programs. Carroll Coatings C-608 is supplied in 1, 3, and 6 lb kits, priced according to the current market value of silver. Literature is available on request.

Contact Carroll Coatings, Division of Spectrum Coating Laboratories, Inc., Earl T. Faria, 217 Chapman Street, Providence, R.I. 02905. INFO/CARD #138.

10 GHz Gunn Diode Oscillator

Engelmann Microwave is offering a new gunn diode oscillator that provides 10-100 MW of adjustable RF output power at 10 GHz. The model MT—A59 provides level set tuning, with mechanical frequency adjustment to provide a 5 percent screwdriver tuning range.

This gunn diode oscillator also features a frequency stability specification of 0.25 to 1 percent depending upon the output power level setting, over the temperature range of −30 to +85°C. The MT-A59 also includes an internal power supply regulator which accepts a wide range of input voltages from +12 to +28 VDC.



Over voltage and reverse voltage protection is provided, making operation of the oscillator safe even with incorrect wiring or voltage inputs. The specified spurious responses of the MT-A59 are −60 dBc and harmonies are −20 dBc.

Contact Engelmann Microwave Company, Skyline Drive, Montville, N.J. 07045. INFO/CARD #137.

SSB-Hand-held

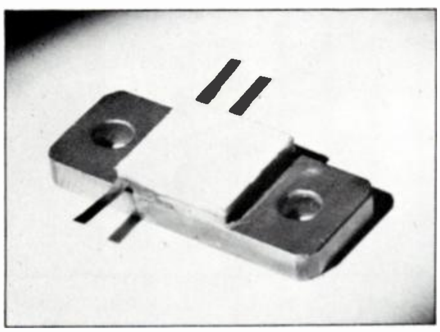
N888 single sideband (SSB) hand-held transceiver. Smallest complete H.F. communications systems available for military, police, high seas marine, government, security, forrestry, mining, pipeline, offshore drilling, exporation. Provides long distance communications where conventional FM equipment is inoperable. Ten watts PEP output power, rugged construction, rechargeable battery (AC charger),

two channels, with both USB and LSB capability, light weight 1.5 kg. size 26 cm x 5 cm x 9 cm frequency 2 to 9 MHz.

Contact Fargo Company, 577 Tenth Street, San Francisco, California 94103. INFO/CARD #136.

100 W Conduction Cooled Power Attenuator

KDI Pyrofilm introduces a new 100 watt conduction cooled power attenuator. The PAA-100 is designed to dissipate 100 watts at a heat sink temperature of 100°C. Frequency is DC to 750 MHz with a maximum VSWR



of 1.25. Attenuation values of 1.0 thru 20 dB ±0.5 dB are available. The resistor substrate is beryllium oxide ceramic with a 96 percent alumina ceramic cover. The tabs are beryllium copper. KDI Pyrofilm also offers flange mounted power attenuators in 20 watt and 50 watt versions.

Contact KDI Pyrofilm Corp., 60 South Jefferson Road, Whippany, N.J. 07981. INFO/CARD #135.

TG Spectrum Monitor

Cushman has added a unique tracking generator to the CE-50A-1 spectrum monitor. The new CE-50A-1/TG simultaneously transmits and receives as it sweeps, providing a real time look, over wide dynamic ranges (better than 70 dB) at the frequency response and return loss (VSWR) of RF and IF circuits and radio systems up to 1000 MHz.

The TG feature converts the standard synthesized signal generator into a swept tracking generator that's automatically locked to the instantaneous frequency of the spectrum monitor as it sweeps. This eliminates tuning to a spur or harmonic, so users can get accurate results, even with filters that have extremely high rejection. Because it is synthesizer-based, there is no waiting for warm-up and no drift after turn-on.

Only the CE-50A-1/TG provides a real-time display of frequency response

and return loss (SWR) of filters, cavities, duplexers, receivers, combiners, antennas and so on over a calibrated 70 dB range (100 dB total dynamic range). TG is offered with new Cushman spectrum monitors or as an option that can be added to most existing CE-50A-1 units in the field.

Contact Cushman Electronics, 2450 North First Street, San Jose, California 95131. INFO/CARD #133.

Insulation Displacing Braid Tap

Only 1/4 inch thick (installed) the AMP braid-tap connector provides a means of connecting a drain wire to a coaxial or shielded cable without pre-stripping insulation from either the drain wire or the cable. Insulating displacing terminals establish electrical connection between the braid and the drain wire as the two halves of the flame retardant thermoplastic housng are pressed together. Three color-coded styles are available to connect RG 174, RG 178, and other cables with a maximum outside diameter of .110" to drain wires ranging from AWG #20 to #24 with a maximum insulation diameter of .054".

Drain wires can be at the ends of the cable or in mid-span with this new product.

Contact AMP Inc., 3901 Derry St., Harrisburg, Pa. 17105. INFO/CARD #132.

Reduce High Frequency Noise Errors

A new two-page technical data sheet from Hewlett-Packard describes the use of their low-pass filter kit and how it can be used to reduce high-frequency noise errors in low frequency measurements. Designed specifically with frequency counters in mind, the kit provides a selection of four filters with low attenuation. They can also be used with oscilloscopes and spectrum analyzers.

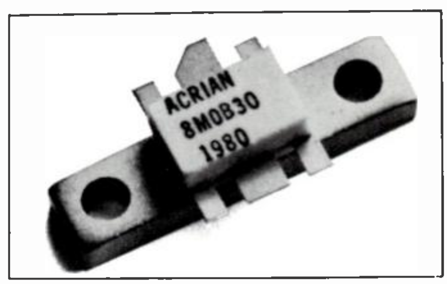
Four typical applications for this HP model 10856A low pass filter kit are described. They show how to filter out unwanted signals, how reducing bandwidth reduces noise error, how to set up the filters for programmable filtering and how they can be used for instrument calibration.

Contact Hewlett-Packard Company, 1507 Page Mill Road, Palo Alto, California 94304. INFO/CARD #131.

Land Power Transistors

Acrian Inc. has available its 806-866 MHz NPN power transistors, designed specifically for land mobile communications equipment.

Offered in power outputs of 1, 5, 15, 30 and 45 watts, the devices feature exclusive use of gold thin-film metalization for maximum protection from operational degradation and proven longest lifetime. Gold controlled-loop wire bonding is used to provide consistent RF performance, and thin film Nichrome emitter ballast-



ing offers improved current distribution and load VSWR tolerance.

The low thermal-resistance packages and eutectic die attach permit junction temperatures which are the lowest in the industry, contributing to maximal MTTF. Extended operational life also results from passivation of device surfaces to eliminate contamination.

RF performance is 100 percent tested and guaranteed in a wide-band fixed tuned test fixture. Operating conditions are: F = 806-866 MHz, V_{CC} = 12.5 volts.

Contact Acrian 10131 Bubb Rd., Cupertino, California 95014. Circle INFO/CARD #130.

Controller for Attenuators And Switches In Microwave Systems

This new Hewlett-Packard model 11713A attenuator/switch driver combines a relay actuator with a power supply in one package and interfaces with the HP-IB (IEEE-488). In microwave systems it can be used to actuate one or two step attenuators of the HP 8694/95/96 or 33320/21/22 series, plus one or two electromechanical switches such as the HP 8761B or 33311 series.

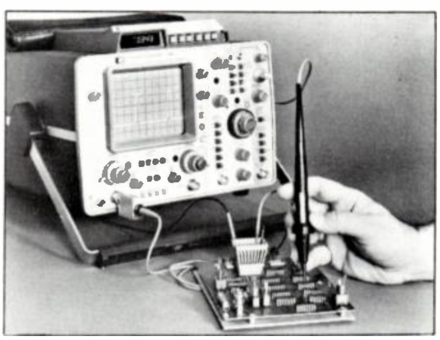
Two sets of four front panel pushbuttons can be used to manually control the attenuators; two additional pushbuttons control the switches. Because the ten front panel pushbuttons each control a transistor switch, they can also be used to control up to 10 external 24-volt relays.

If desired, full remote HP-IB control can be selected. Because the 11713A is easy to program, a simple subroutine can be incorporated into a general program.

Contact Inquiries Manager, Hewlett-Packard Company, 1507 Page Mill Road, Palo Alto, California 94304. INFO/CARD #110.

Temperature Probe

Fast and accurate temperature measurements such as are required in thermal design, diagnostic and testing applications are obtained using this new temperature probe from Hewlett-Packard. Called the HP model 10023A, it reads surface temperatures directly in degrees Celsius on any general purpose digital multimeter (DMM) with



an input of ≥ 10 megohms. Operation is easy with its convenient pencil-like probe tip and press-to-read switch.

The sensor diode in each probe is individually characterized in a precision thermal reference bath to obtain a calibrated, linear output of 1mV/°C. Additionally an integrated circuit resistor network is laser trimmed to match each diode to its electronic compensating network. The result of this type of construction is a factory calibrated probe with no internal adjustments and no bother of re-calibration.

Measurement accuracy, which is traceable to the National Bureau of Standards, is ±2°C from 0°C to +100°C decreasing linearity to +2°C, −4°C at −55°C and to +4°C, −2°C at +150°C. For applications requiring relative rather than absolute measurements of similar temperatures, the short-term repeatability is ±0.3°C. The temperature sensor has a very low thermal mass which is positioned very close to the measurement surface to assure fast temperature measurements. This design also assures measurements with very low thermal gradient errors.

High thermal isolation reduces the tendency for the probe tip to act as a heat sink and change the measured surface temperature of small electronic components. An electrically isolated probe tip to 600V also allows measurements of nongrounded components such as power transistors with the collector common to the case.

Contact Hewlett-Packard Company, 1507 Page Mill Road, Palo Alto, California 94304. INFO/CARD #129.

LSI Synthesizer Family

Motorola has introduced two new frequency synthesizer integrated circuits. They are the first of a series of seven devices, planned to change the architecture of much of the equipment currently used for RF communications. The new devices use large-scale-integration (LSI) technology to reduce synthesizer costs, thereby fostering the implementation of phase-locked loop (PLL) circuitry in equipment that still relies on tuned circuits and multiple crystals.

The new silicon gate CMOS ICs provide a high level of versatility and performance in conjunction with low DC power drain requirements. The typical frequency range extends to over 25 MHz and, when required, can be expanded to over 500 MHz with the addition of existing high frequency dual modulus prescaling ICs. Three family members are compatible with this frequency extension technique.

The MC145155 and the MC145156 are the two devices that are introduced. They are programmed with a clocked, serial input bit stream. Other members of the 7-device family are still under development but are due for introduction in 1980. These include the MC145151 and MC145152 for parallel programming and the MC145144, MC145145 and MC145146 which provides a 4-bit data bus programming approach. Ease of interfacing with microprocessor/microcomputer controllers is provided by the serial and 4-bit data bus units.

Conact Motorola, P.O. Box 20912, Phoenix, Ariz. 85036. INFO/CARD #128.

SIP Ceramic Capacitors

Varadyne Industries, Inc., introduced a line of single in-line monolithic capacitors in a wide capacitance range.

These SIP capacitors are particularly well suited for applications on high-density printed circuit boards as their design and compact size allows improved packaging density as well as reduced assembly time. The units come standard in an 8-pin configuration, and are made of flame retardent conformal epoxy to provide strength and durability.

They are available in a wide capacitance range of 15pf to .01omFd in both COG (NPO) and X7R ceramic formulation. They withstand a wide temperature range of −55°C to 125°C and come in 5 percent, 10 percent, and 20 percent tolerances.

Contact Varadyne Industries, Inc., 1520 Cloverfield Blvd., Santa Monica, Calif. 90404. INFO/CARD #124. □

One Great Value Three Great Ways

Programmable Direct Synthesizer: 0.1 to 160 MHz

THE VALUE

Rockland Series 5600 Programmable Frequency Synthesizers employ the *direct* synthesis technique – no slow and noisy phase-locked loops – yet cost less than many PLL designs in this range! Resolution is *constant*: 1 Hz across the entire 0.1 to 160 MHz range. That's a *single* range, too; no range switching, no multipliers. Spectral purity is outstanding: −70 dB phase noise: −35dB harmonics; −70 dB spurious. Stability is exceptionally high: 1×10^{-9}/day, with a very low T.C. (1×10^{-8} from 0°C to 50°C). Or inject your own external reference. Output levelling is exceptionally tight: ±0.5 dB throughout the frequency range.

Digitally programmable at much higher speed than conventional PLL designs: 20μsec switching time, negligible switching transient. All functions are remotely programmable (*including* level).

Applications unlimited: satellite communications, NMR source, spectrum analysis, HF surveillance receivers, radar testing, frequency-agile/automated test systems, manual testing, crystal manufacturing and calibration, and as a true secondary transfer standard of frequency.

The greatest value: Rockland engineering and manufacturing experience. Superb quality. Maximum applications support.

THE WAYS

Model 5600 has manual front-panel controls plus full remote digital programmability.

Model 5610A has blank front panel, no manual controls, but the same full digital programmability. Considerably lower in price than Model 5600. Ideal for OEM Systems.

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