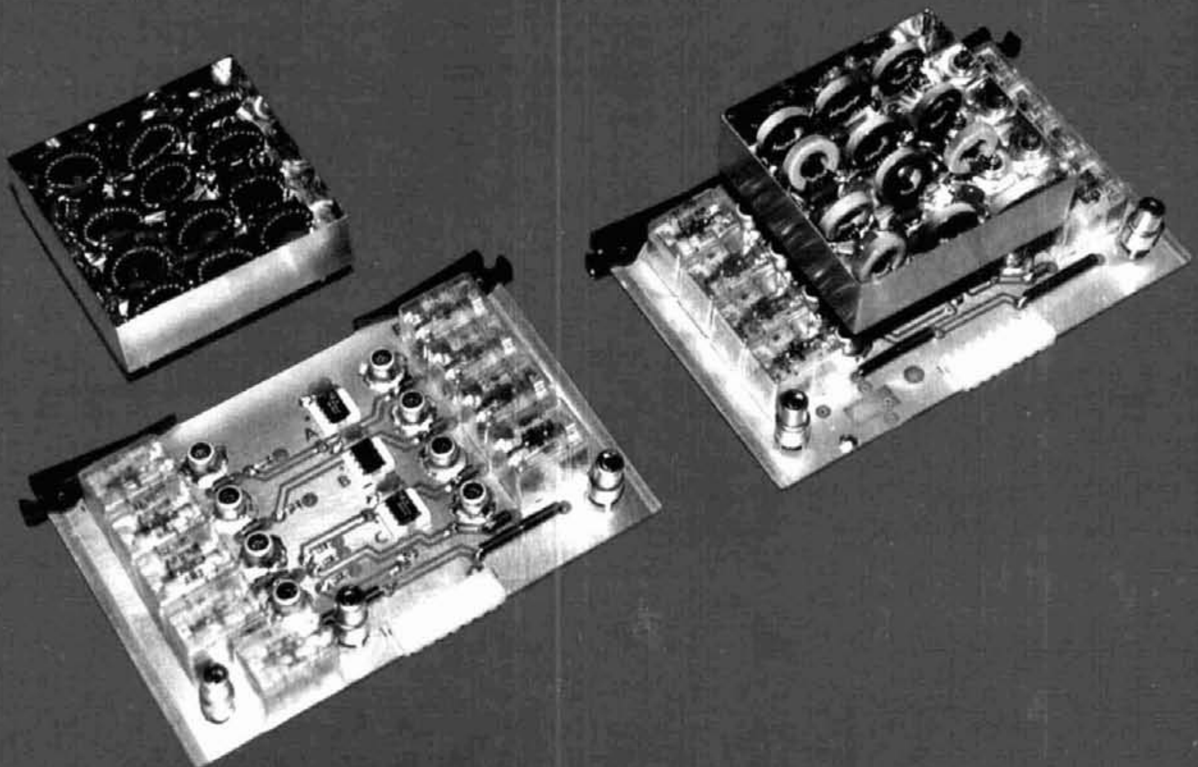
FEBRUARY 1988 / $2.50

amateur packet radio

ham radio • February 1988

ham radio magazine



hr

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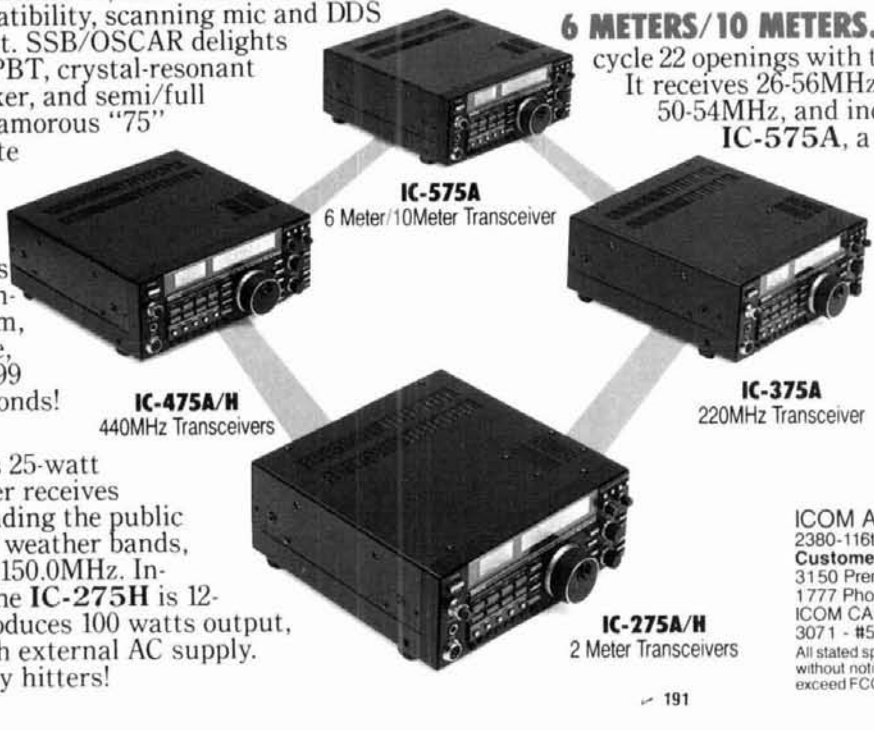

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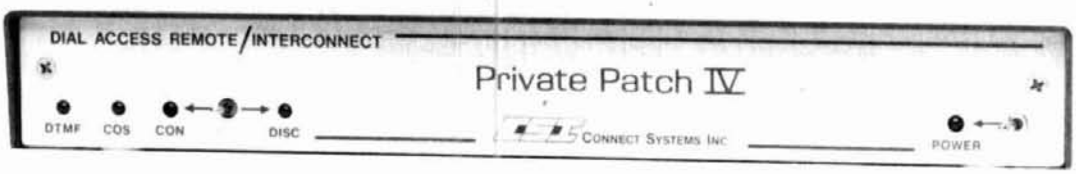

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ham radio

magazine

FEBRUARY 1988

volume 21, number 2

ham radio magazine is published monthly by
Communications Technology, Inc.
Greenville, New Hampshire 03048-0498
Telephone: 603-878-1441

subscription rates

United States:
one year, $22.95; two years, $38.95; *three years*, $49.95
Europe (via KLM air mail), $40.00
Canada, Japan, South Africa and other countries (via surface mail),
one year, $31.00; two years, $55.00; three years, $74.00

All subscription orders payable in U.S. funds, via international
postal money order or check drawn on U.S. bank

international subscription agents: page 97

Microfilm copies are available from
University Microfilms, International
Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles from *ham radio*
are available to the blind and physically handicapped
from Recorded Periodicals,
919 Walnut Street, Philadelphia, Pennsylvania 19107

Second-class postage paid
at *Greenville, New Hampshire* 03048-0498
and at additional mailing offices
ISSN 0148-5989

Send change of address to *ham radio*
Greenville, New Hampshire 03048-0498

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publisher's log

Help Wanted — Apply Within

Over 5 years ago, Rally Dennis was invited to Ham Radio for a job interview. I was immediately comfortable upon meeting him and felt his addition to Ham Radio's staff would be a perfect meld of personalities — and I was correct in my assumption. We have worked well together over the last few years and enjoyed our association. Working as a team, our advertising sales have been good; Rally deserves credit for 200% of the success. Truthfully, he has forgotten more about ad sales than many of us will ever know and he is a true master at his craft.

He has decided, however, to slow down a bit as he approaches his 70th birthday. Effective December 31, he will semi-retire from Ham Radio and become an independent advertising sales consultant for the magazine. This will allow him the best of both worlds. No more shows or early morning flights from Boston's Logan Airport — no need to hurry to work every day. Time can be spent on his auctioneering, Game Preserve Museum, and antique business. But Rally will stay in touch with his friends in the Amateur Radio business working out of our office and his home about five days per month.

Photo by Phil Alix

One project that becomes imperative is to find a new person to fill Rally's shoes. If you have a sales or marketing background and would be interested in discussing the position as advertising salesperson, please send us a resume and other information of interest about yourself. Working at Ham Radio has plenty to offer. The virtues of living and working in southern New Hampshire are almost too numerous to mention. Suffice to say — we're not fighting traffic jams as we drive home at night. Centrally located to the ocean, Boston, and the mountains, Southern New Hampshire has something for almost everyone.

Drop a note and your resume in the mail and we'll see about getting together for an interview to discuss this exciting opportunity. It's not too many jobs that require you to go to the Dayton Hamvention, all expenses paid, every year!

J. Craig Clark, Jr., N1ACH
Assistant Publisher

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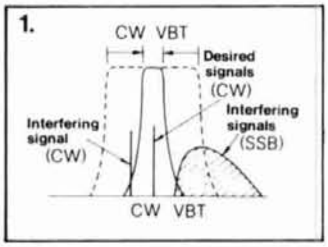

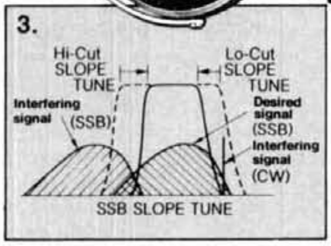

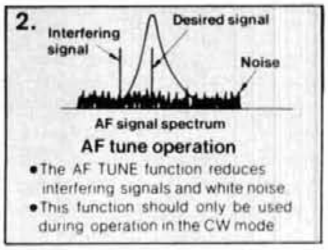

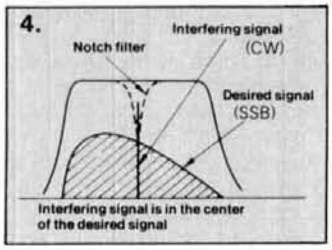

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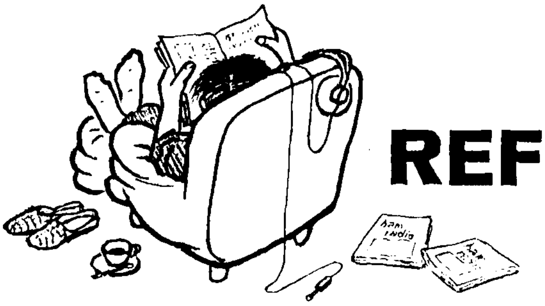


REFLECTIONS

RETURNING THE SPECTRUM TO CHAOS...Courtesy of the FCC

In the early years of wireless, radio communication was at best a chancy, chaotic business. Broadly-tuned spark-generated signals, received on primitive, unselective receivers, forced operators to wait for a break in the QRM to pass traffic or add their signal to the bedlam in hopes that the intended recipient would pick it up.

In the last 80 years, improved technology and international cooperation have brought order to the radio spectrum. With few exceptions, millions of transmitters in different services operate harmoniously under domestic and international regulations on designated frequencies or bands throughout the spectrum.

It now appears the FCC wants to turn back the clock — at least in the United States — and return much of the radio spectrum to its chaotic origins. In its recent Notice of Proposed Rule Making — General Docket 87-389 — the FCC proposes expanded use of unlicensed rf-emitting devices over most of the radio spectrum "without restriction as to bandwidth, duty cycle, modulation technique or application..." as long as their emissions do not exceed specified limits!

From 1.705-30 MHz, "Intentional Radiators" could generate signals measuring 30 μV/meter at a distance of 30 meters *with NO radiation limitation for non-digital "Unintentional Radiators"*. The change in VHF/UHF limits to 100 μV/meter from 30-88 MHz, 150 SP^{6}mV/meter from 88-216 MHz, 200 μV/meter from 216-960 MHz, and 500 μV/meter above 960 MHz, all measured at 3 meters, is minimal.

Radio Amateurs would need an S9 signal to break through the rf bedlam on 10 to 160 meters. Because VHF/UHF band (signal) limits are already high, the effect won't be as pronounced. But similar problems would eventually occur since this proposal encourages expanding the uses of unlicensed rf generators. Any repeater without CTCSS access might be keyed up continuously and the hand-held range could become severely limited.

This eventual spectrum deterioration would affect more than the Amateur Radio community. All land mobile (police, fire) services, plus paging and radio-relayed telephones, would soon find themselves fighting their way through a cacaphony of rf noise. Even the U.S. military, AM, FM, and TV broadcasting, won't escape the havoc this proposal will bring. If your neighbors think your infrequent incursions into their TV viewing are a problem now, wait until they encounter proliferating unlicensed rf generation!

The proposal has some restrictions. A few narrow QRM-free slots would be reserved for radio astronomy, maritime distress, and navigational aids. Most VHF and UHF aircraft bands would be off limits to unlicensed operations, along with a couple of VHF federal land mobile bands. But for the rest of us, it'll be a problem.

This proposal applies to "Part 15" devices — intentional radiators (unlicensed low-power transmitters) like garage door openers, wireless microphones, and cordless telephones, and unintentional radiators (electronic devices) like computers, TV receivers, cable TV equipment, and VCRs which also emit some rf radiation. Over the years, the rules governing operation of these devices have been piecemeal, complex, confusing, and contradictory. This NPRM attempts to simplify and broaden Part 15 regulations without, the FCC believes, causing undue interference to existing services.

The Commission announced General Docket 87-389 in September with a comment due date of December 4 — too late for Amateur publications to review the proposal, its band-by-band summary and errata sheets, and relay the information to their readers with enough response time. Fortunately, organizations like ARRL and ANARC (Association of North American Radio Clubs), a SWL/scanner-user group, realized its importance and petitioned the Commission for an extension which was granted to March 7, 1988.

To tell the FCC of your opposition, file a written comment emphasizing the certainty of crippling interference to most established radio services. In his Comments filed before the original deadline, George Jacobs, W3ASK, noted that the internationally recommended signal-to-noise ratio for shortwave broadcast reception is 27 dB. The FCC's proposed interference levels would demand 672 microvolts from international broadcasters. But, Jacobs said, overseas broadcast signals generally run 150 to 500 μV. Radio amateurs are now communicating with a few microvolts.

Ask what *demonstrated* need there is for proliferation of such devices throughout the radio frequency spectrum. If a variety of frequencies for Part 15 devices is necessary, why not set aside specific frequencies or bands for them as was done with industrial, scientific and medical bands.

Point out that current techniques allow excellent frequency control and easy, economical transmission of large amounts of data in relatively narrow bandwidths. Why then, must so much spectrum be accessible to Part 15 use? Why shouldn't proper shielding be required to reduce Part 15 unintentional radiation to acceptable levels? Why should the licensed users of the spectrum be forced to carry the burden for poorly designed and manufactured consumer goods?

You'll probably think of more arguments to include with your Comments on General Docket 87-389, but whatever points you choose to emphasize, be sure the FCC knows how strongly you oppose it!

How do get your Comments to the FCC?

Type "Comments — General Docket 87-389" at the top of an 8-1/2 x 11 sheet of paper, state your argument, and end with your complete name, callsign, and address. If using more than one page, be sure to put the page number, your name, and the Docket number on each. Send it to Mr. William J. Tricarico, Secretary, Federal Communications Commission, 1919 M. Street, NW, Room 222, Washington, D.C. 20554 — preferably with the original and 11 copies (one for each commissioner). If you don't have access to a copier, your single submission will still make a difference.

Remember, *we must fight this proposal.* Your comments must reach the Commission before March 7, so why not start writing now?

Joe Schroeder, W9JUV

comments

cover

Dear HR:

The front page graphic, depicting Ham Radio in the December issue, was indeed a superb masterpiece.

That alone, is more than enough to keep a "honest-to-goodness" ham solidly attached to your fine magazine. The remaining contents, strictly a generous bonus.

Ray Ziminski, K2KC
East Meadow, New York 11554

wanted: hf amplifier

Dear HR:

In regards to the comment made by W7WRQ in the December 1987 issue of HR—he is correct. An SCR-270 radar unit was used and if I remember two antenna bays were used instead of the usual one. I worked on the composite tests on Sandy Hook, New Jersey in 1941 to get them ready for Pearl Harbor. I never knew that it was a problem to get "high powered" hf transmitters going in the '30s. They were on SSB also! While I did have a Collins transmitter with an intermittant cold solder joint that I had a heck of a time finding, Art Collins was an outstanding ham and Orr's article should be well received.

I am interested in a solid-state 500-1000W hf amplifier. The designs and/or built-up units that I can find are 10-12 years old. I know there are new 300-600W FETs. I can't seem to find any design material, kits, parts, or built-up units. If you have any pending articles, know of any source of parts, kits, boards, designs, whatever, I would like to know about the availability. I might even produce an article if I get an HF amplifier going to my satisfaction.

Wayne W. Cooper, AG4R
Miami Shores, Florida 33150

computerized Yagi beam antenna

Dear HR:

Having just received the August issue, I noted with a great deal of interest the writeup on the "computerized Yagi beam antenna" and especially the captions on the modified 205BA.

Having owned one for over a year, I have investigated a number of possible modifications, inclusive of a possible alteration to a wide-spaced four elements on a 45-foot boom, which modelled quite well.

The main attraction in Wayne Hillenbrand's design is in achieving the stated performance while maintaining the original boom length, as the 2-inch boom of the HY-GAIN would not take the extra stress very well, when extended to 45 feet.

As there are quite a few hams interested in improving the performance of their 205BA's, may I suggest that a more detailed presentation be included in one of your future issues.

Keep up the good work.

Julian S. Biermann, ZS6BJO
Germiston 1407, South Africa

great reference

Dear HR:

Just finished devouring the December issue and thought I'd send a quick note to share some thoughts.

I was mentioning last night to a friend that Ham Radio is the only (of many I receive!) magazine which I save in its entirety for reference. Seems there's always something interesting to read and learn. The December issue is a perfect example of what I meant!

I've always liked W1JR's columns. Bill Orr's column about Art Collins was touching. I never knew Art, but had a very close friend (W4MJJ, now Silent Key) who did. The stories Mel used to tell about Art and Curt Lemay during the early days of amateur SSB were always fascinating. Bill did well.

KL7AJ's article about TV didn't teach me anything new, but it was an incredibly well written review — more of what makes HR great.

I guess the bottom line is that HR is great! Please accept my thanks for HR's contribution to ham radio and to RF technology. Keep it up!

LCDR James A. Sanford, USN, PE, WB4GCS, Hampton
Virginia 23669

the good old days

Dear HR:

I wanted to let you know how much I enjoyed reading about Mr. Collins. I could not afford Collins for many years. I now own an S-line (32S-1/75S-1) and it is a treasure to me.

Thank Mr. Orr for writing about "the good old days," Mr. Collins and Collins Radio.

Robin Chestnut, WA5YGR
Perry, Oklahoma 73077

automatically switched half-octave filters: part 1

Want to exceed FCC purity-of-emissions specs? Try combination low-pass/band-pass filters.

The authors have developed a system of switched filter banks for use in equipment where contiguous, controlled bandwidth and sometimes constant delay designs are required over wide frequency ranges. Although the design presented here is intended for constant harmonic attenuation from 2 to 30 MHz, the concepts extend to more than just hf communications. — Ed.

In the good old days, radios had several bands. Mechanical band switching was accomplished by means of a multiwafer switch that extended all the way from the front panel to the back panel, with complex filter networks in between. While some of these systems are still in operation today, the introduction of digitally synthesized transceivers has brought about the need for a totally new approach to band switching. While it is a relatively simple matter to design channelized, band-switched rf equipment, the design of true broadband solid-state transceivers requiring stringent harmonic attenuation and linear phase response specifications over several hf octaves would require a multitude of band-switched filters. Using a manual band switch such as those found in old transceivers would lead to a cumbersome and impractical design.

On the other hand, one could look at a general-coverage hf synthesized radio as a device having one huge band, with an ultimate resolution of, say, 10 Hz. This would make it practical to digitally control simple, stand-alone, wide-range varactor-tuned filters which are small in size and require simple control circuits. While such circuits would be adequate in receiver environments with relatively low-level undesirable signals, the nonlinear nature of the varactor diodes can cause intermodulation distortion in the presence of higher level signals, *despite* the fact that back-to-back circuits are usually used to reduce this effect. If a transmitter is also involved, such as in the case of a transceiver, these circuits would become impractical; besides, high-power, contiguously switched filters are usually preferred. The proper selection of these filter networks requires digital intelligence in order to allow automatic coverage of the entire spectrum of interest. This, in turn, requires a unique system design.

Because there are four octaves* of bandwidth from 2 to 30 MHz, half-octave filters could be used in the system in order to keep out the second harmonic and the higher products of any of the 2,800,000 possible fundamentals (when using 10-Hz resolution) produced by the synthesizer. This concept is illustrated in **fig. 1**. Several half-octave filters will be required to meet this need. The important issue faced by a designer in such a case is not what microprocessor to use to crunch the numbers, but rather what is the best cost/complexity/performance compromise allowable by the total system design — both from a digital point of view and, more importantly, from an rf point of view. Understanding this process of design is the essence of this article.

* One octave is defined as the frequency range between a fundamental and its second harmonic.

By Cornell Drentea, WB3JZO, 7140 Colorado Avenue North, Minneapolis, Minnesota 55429, and **Lee R Watkins,** 2256 East Jaeger Street, Minneapolis, Minnesota 85213

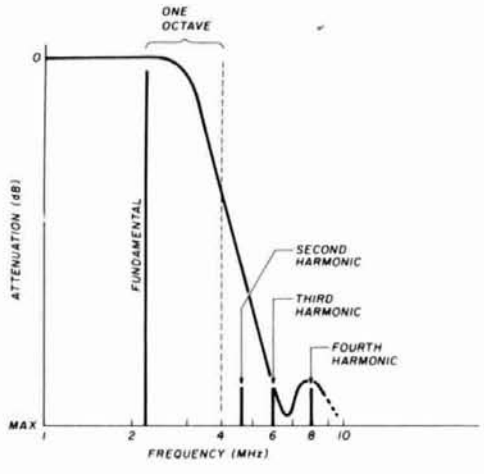


fig. 1. How a half-octave filter attenuates harmonics. Since harmonics are located at least one octave away from any fundamental, a worst case analysis for the entire range of 2 to 30 MHz indicates that several contiguous filters covering half-octaves will reject harmonics always by the same amount. An automatic switched bank of eight half half-octave filters are required to cover the 2 to 30 MHz range.

design criteria

The system described here uses eight switched filters. To eliminate any possibility of intermodulation distortion, no diodes are used in the rf switching. The filters are of a high-order bandpass design working in harmony with eight additional high-power, low-pass filters of similar design. It can be seen that because of the half-octave choice, given proper design, the filters can effectively attenuate all harmonic products at any point in the frequency coverage between the specified 2 and 30 MHz. This design calls for a composite attenuation between the filters and the linear amplifier's own harmonic response of greater than 50 dB for the specified range when used with the 120-watt linear amplifier in the WB3JZO transceiver's output. This exceeds the FCC's Part 97.73 requirement for purity of emissions for Amateur equipment.* In addition, a passband ripple requirement of 0.1 dB† was imposed on the design of all filters in order to keep receiver input impedance, and consequently noise figure and sensitivity, constant.

Switching in the appropriate set of half-octave filters for the frequency of interest can be performed automatically by the digital information available from the transceiver's frequency command input. The areas of

* The FCC requirement below 30 MHz calls for power of any spurious emission from an Amateur transmitter or transceiver to be at least 40 dB below the mean power of the fundamental, *but not to exceed 50 milliwatts.*

† With a $Q = 100$ for the inductors, a ripple of approximately 0.05 dB has been realized in practice.

. . . a word from the authors

This article, dedicated to the memory of Anatol I. Zverev, describes the design and development of a 2- to 30-MHz switched filter system as used in the front end of a modern, fully synthesized transceiver. Intended as a brief tutorial, as well as a construction article reaching beyond this single application, it enables one to appreciate the complexity of modern equipment and filter design, and shows in detail the design and development of a complex switched filter system whose harmonic attenuation exceeds the requirements for Amateur equipment. Although the emphasis is on construction, no layouts or physical details are given in order to keep the length of the article within reasonable limits.

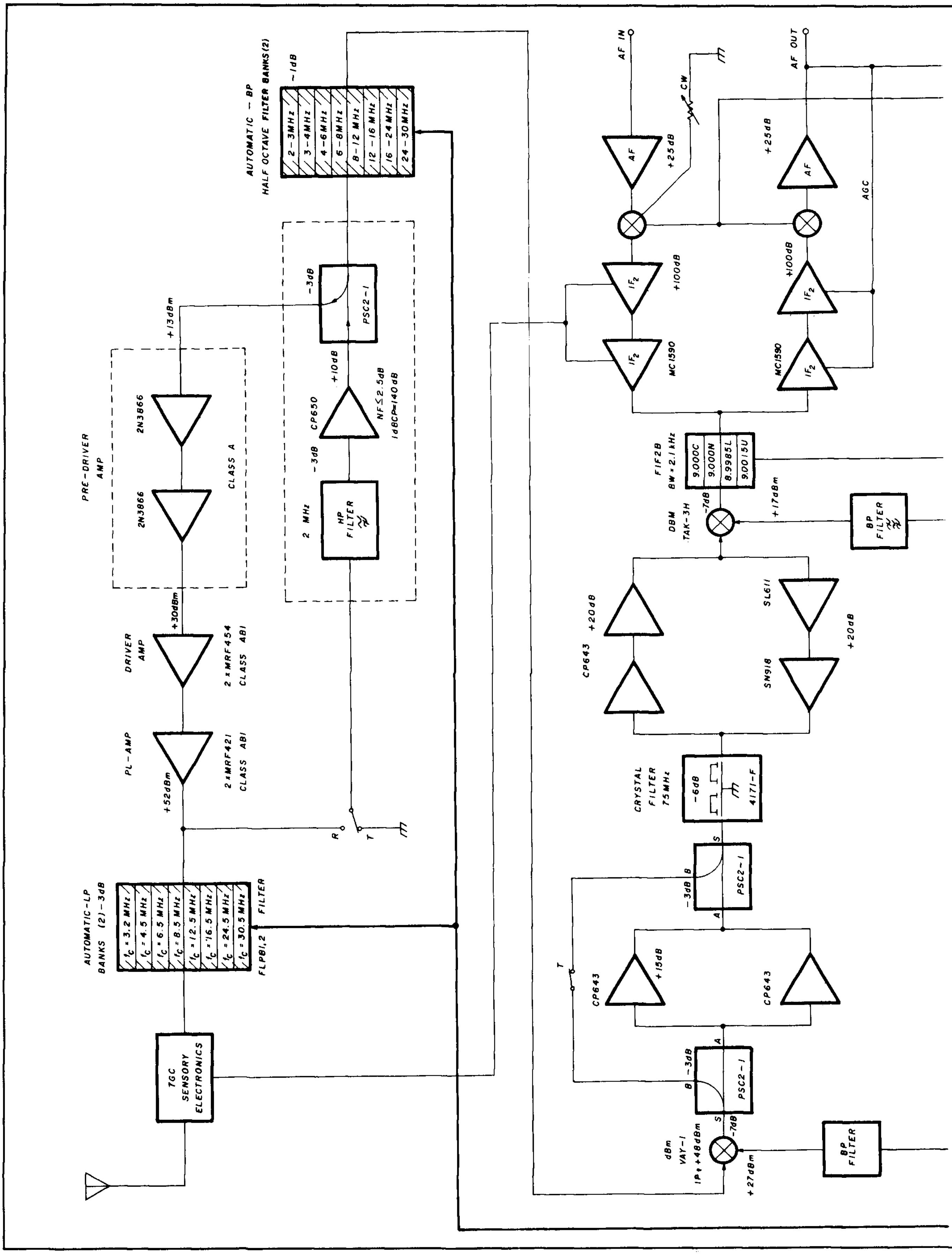
Part 1 discusses filter theory and design. Part 2 concentrates on implementation of the complex filter banks, digital control and execution, automatic high-power rf switching, and the practical aspects of the entire project as used in the WB3JZO home-built transceiver. It should be noted that this project is relatively complex and will thus not offer immediate gratification; it should be viewed, therefore, in the context of a larger project such as the design of a transceiver, receiver, or transmitter.

The project will apply to other automatic filtering applications of stand-alone, high-power, solid-state, hf linear amplifiers requiring stringent harmonic suppression and linear phase response over wide bands. These functions would be needed in order to reduce dynamic range requirements of co-located receivers or to provide phase-coherent data communications in installations requiring frequency diversity. In linear amplifier applications, a modified design (not described here) could contain sensory circuits to read the frequency of the rf present at the amplifier's input and quickly select the proper network at the output.

The article will also apply to receivers. The development of an off-board preselector which combines the techniques presented here with other techniques could be added to the front end of existent wide-band receivers to improve their in-band and out-of-band dynamic range performance.

We hope that this material is informative and useful for both beginners and advanced Radio Amateurs.

—Cornell Drentea, WB3JZO, and Lee R Watkins


AUTOMATIC – BP
HALF OCTAVE FILTER BANKS (2)
−1dB
2 – 3MHz
3 – 4MHz
4 – 6MHz
6 – 8MHz
8 – 12 MHz
12 – 16 MHz
16 – 24MHz
24 – 30MHz
PRE-DRIVER
AMP
2N3866
2N3866
CLASS A
+13dBm
−3dB
PSC2-I
+10dB
CP650
NF≤2.5dB
1dBCP=140dB
−3dB
2 MHz
HP
FILTER
+30dBm
DRIVER
AMP
2×MRF454
CLASS ABI
PL-AMP
2×MRF421
CLASS ABI
+52dBm
R
T
AUTOMATIC-LP
BANKS (2)-3dB
FILTER
FLPBI,2
f_c = 3.2 MHz
f_c = 4.5 MHz
f_c = 6.5 MHz
f_c = 8.5 MHz
f_c = 12.5 MHz
f_c = 16.5 MHz
f_c = 24.5 MHz
f_c = 30.5 MHz
TGC
SENSORY
ELECTRONICS
AF IN
AF OUT
CW
AF
+25dB
AF
+25dB
AGC
IF_2
+100dB
IF_2
MC1590
IF_2
+100dB
IF_2
MC1590
FIF2B
BW=2.1kHz
9.000C
9.000N
8.9985L
9.0015U
DBM
TAK-3H
−7dB
+17dBm
BP
FILTER
CP643
+20dB
SL611
+20dB
SN918
CRYSTAL
FILTER
75MHz
−6dB
4171-F
S
B
−3dB
PSC2-I
A
T
CP643
+15dB
CP643
A
B
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PSC2-I
S
−7dB
dBm
VAY-1
IP_3 +48dBm
+27dBm
BP
FILTER

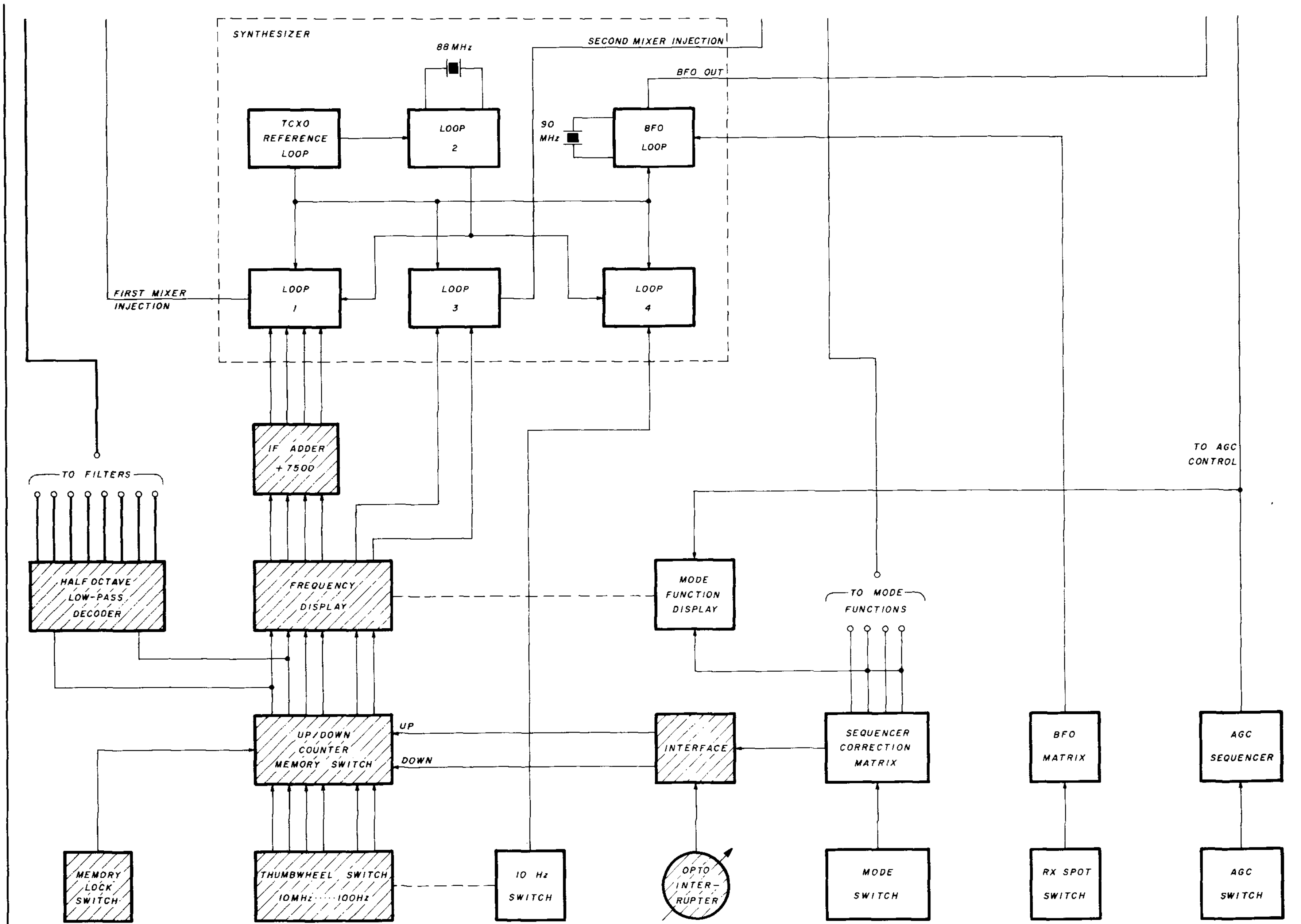

fig. 2. Areas of the transceiver discussed in this article. The block diagram shows the half-octave automatic filter system, which provides attenuation for harmonics generated by the wide-band transmitter. Low-pass filters are switched in along with bandpass filters. Total passband ripple does not exceed 0.1 dB.

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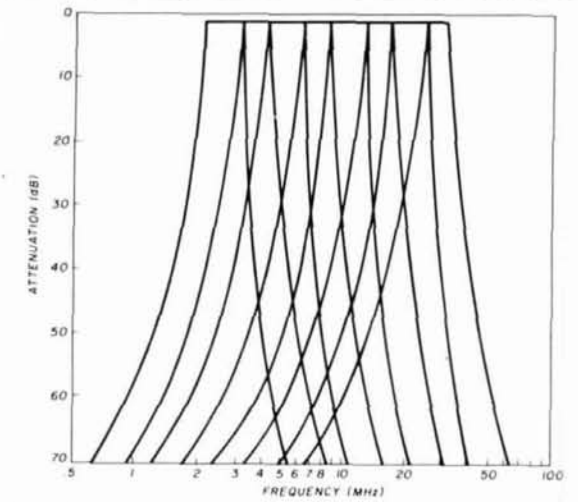

fig. 3. Composite frequency response for all lowpass and bandpass, half-octave filters. No matter where the fundamental frequency is generated by the transmitter, total harmonic content will always be at least 50 dB below it.

the transceiver affected by this article have been outlined in the block diagram shown in **fig. 2**, which illustrates that both the low-pass and the bandpass banks are always in the circuit regardless of whether the transceiver is in receive or transmit. Only one filter set and its corresponding pair are in the path at any given time.

A class A/AB amplifier has been specifically designed to keep the second harmonic down and therefore simplify the filter design requirement. The level of the third harmonic which is the worst-case product will determine the overall attenuation required by the filters and consequently their order and design type. However, since the third harmonic is further away from the corner frequencies of the half-octave bands, a less complex filter is required. This criterion is valid for all the frequencies and their corresponding harmonics in each of the eight filters, as affected by the linear amplifier's harmonic characteristics.

This principle also applies in receive, where the image, a third-order product, and higher order products are kept out of the receiver's input in any of the eight selected ranges through a combination of bandpass and low-pass filters. The image rejection specification for a receiver with a first i-f of 75 MHz, using the front-end filters designed here, has been calculated to be 70 dB.

So far we have determined the overall characteristic response of our filters as matching at least a 50-dB harmonic attenuation requirement when used with a specifically designed amplifier. This composite frequency response is shown in **fig. 3**. We will now examine the design of the filters more closely.

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table 1A. Bandpass. Design requirements for all half-octave filters used in the transceiver.

Filter No.	Passband (MHz)	Stopband (MHz)	Attenuation (dB)
1	2-3	<.625; >5.25	70
2	3-4	<.875; >7.75	70
3	4-6	<1.25; >10.5	70
4	6-8	<1.75; >15.5	70
5	8-12	<2.50; >21.0	70
6	12-16	<3.50; >31.0	70
7	16-24	<5.00; >42.0	70
8	24-30	<6.75; >61.0	70

table 1B. Lowpass. Design requirements for all half-octave filters used in the transceiver.

Filter No.	Cutoff (MHz)	Stopband (MHz)	Attentuation (dB)
1	>3	6	60
2	>4	9	60
3	>6	12	60
4	>8	16	60
5	>12	24	60
6	>16	36	60
7	>24	48	60
8	>30	72	60

parts required for the construction of the low-pass filters, it was decided that the lowest order filter that satisfied the above criteria would be considered in order to conserve space. First, the linear amplifier chain was designed and developed in order to realize a second harmonic attenuation of 30 dB. As mentioned before, the rejection requirements have been met by complementing these specifications with the linear amplifier's own harmonic attenuation.

During amplifier design, a nonlinear transfer function analysis was performed with the help of the Volterra series.* The data obtained was then compared with the harmonic content information available from Motorola's *RF Data Manual* and the actual spectrum analyzer data obtained from the designed amplifier. After several design iterations, the second harmonic emission was brought within the specification.† This placed a 20-dB attenuation requirement on all low-pass half-octave filters at the second harmonic in order to achieve the required 50 dB. The worst-case third harmonic emission was verified at 10 dB below the carrier. This placed a 60-dB attenuation requirement on the low-pass filters at the third harmonic frequencies and beyond in order to maintain a specified 70-dB total rejection. **Table 1** lists all the resulting electrical requirements for the bandpass and low-pass filters.

A more stringent requirement intended for linear amplifiers with a lesser second harmonic rejection was

* Unlike the classic model to Fourier analysis, the Volterra series is a much more precise modeling tool used in the study of nonlinear effects in transistor amplifiers and receivers.

† The 39-dB gain amplifier uses a combination of class A/AB designs.

worked out but was not implemented because of its complexity (it required a minimum of eight poles). However, the performance of the filters described here in standard solid-state, push-pull power amplifiers of class AB or B with matched characteristics and with a typical second harmonic rejection of 20 dB below the carrier or better, will also meet or exceed the 40-dB FCC requirement for Amateur service.

the ideal theoretical filter

The following is a discussion of the basic theory required for understanding of the filter design part of the article. Because filter articles can become very involved, we chose to emphasize only limited theoretical aspects of the subject. Only the bandpass design of one filter will be treated in detail. However, references to the low-pass equivalent design and detailed construction data are provided.

There are many ways to design filters. An ideal filter has no insertion loss in its passband (0 dB attenuation from $\omega = 0$ to $\omega = 1$), and infinite attenuation everywhere else. ($\omega > 1$!) as illustrated in **fig. 4A**. Mathematically, the value of the magnitude response function $|H(j\omega)|$ in such a filter would be infinite, which in turn would require an infinite number of poles. Finally, the filter would not be practical since an infinite number of poles would create an infinite insertion loss and an infinite delay for the waveform at its output. In addition, the ideal filter would require a linear phase response over the entire passband, as shown in **fig. 4B**. Other important elements affected in such a design would be the impulse and step response characteristics as shown in **figs. 5A** and **5B**. It can be seen that for the two responses, the output of the filter starts at $t = -\infty$ while the input is only applied at $t = 0$, a real-world impossibility.* This brief theoretical discussion is important because it makes us realize the imperfection of any filter design by comparison.

practical filters

Because actual filters can be no more than approximations of ideal filters, tradeoffs of performance characteristics are inevitable. Knowing what is important in each particular application determines the type of approximation. Designers have many filter types to choose from: Butterworth, Chebyshev, Legendre, Gaussian, Least-Square, Laguerre, Hermitian and Bessel; each is an approximation of the ideal filter, and each has its own positive characteristics. Chebyshev filters have an equal-ripple passband and steepest out-of-band attenuation (i.e., monotonically increasing attenuation). Since a 0.1-dB passband ripple specifi-

* The linear phase and constant group delay requirements imply that the response be of an anticipatory nature. The impulse response is obtained from the Fourier inversion integral. This integral produces a response which starts at $-\infty$ and ends at $+\infty$.

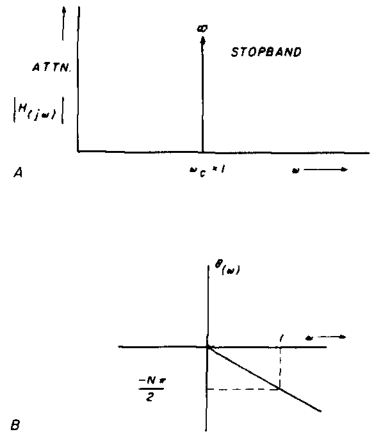


fig. 4. Ideal filter. (A) Amplitude frequency response. (B) Linear passband phase response.

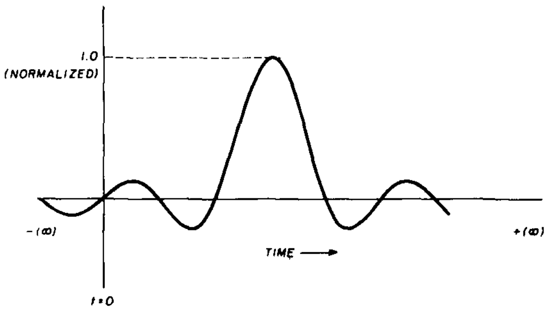


fig. 5A. Ideal filter. Impulse response.

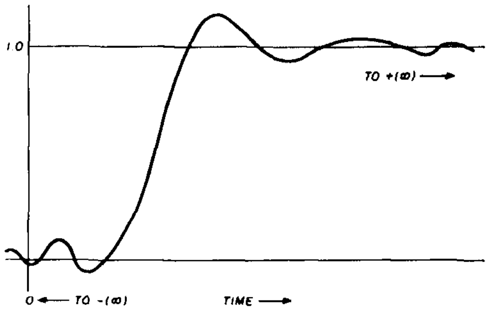
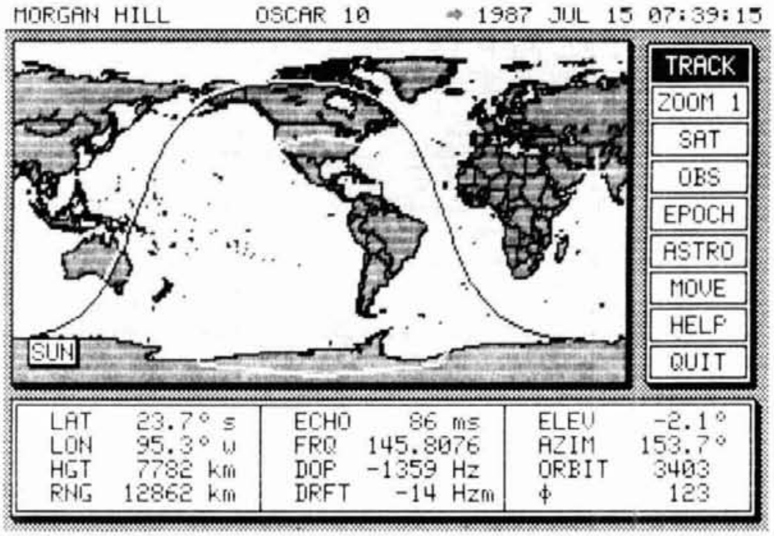


fig. 5B. Ideal filter. Step response.



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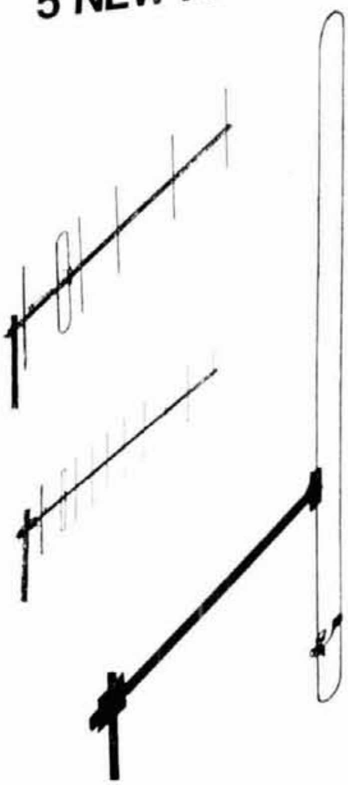
BALUN	4:1 coax
MECHANICAL:	
ELEMENT LENGTH	13½" max.
BOOM LENGTH	28"
TURN RADIUS	28"
WINDLOAD	.2 sq. ft.
WEIGHT	1 lb.
MAST	1½" o.d.
MOUNT	Rear

440-10X

ELECTRICAL:	
BANDWIDTH	420-460 MHz
GAIN	11.2 dBd
VSWR	1.5:1
F/B	20 dB
BEAMWIDTH	48°
FEED IMP	50 ohm
BALUN	4:1 coax
MECHANICAL:	
ELEMENT LENGTH	13½" max.
BOOM LENGTH	64"
TURN RADIUS	64"
WINDLOAD	.4 sq. ft.
WEIGHT	1½ lbs.
MAST	1½" o.d.
MOUNT	Rear



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VSWR	1.5:1
FEED IMP	50 ohms
NO GROUND PLANE REQUIRED	
MECHANICAL:	
HEIGHT	61"
WEIGHT	2½ lbs.
MAST	1½" o.d.

CJ220

ELECTRICAL:	
BANDWIDTH	220-224 MHz
GAIN	1.8 dBd
VSWR	1.5:1
FEED IMP	50 ohms
NO GROUND PLANE REQUIRED	
MECHANICAL:	
HEIGHT	40"
WEIGHT	2 lbs.
MAST	1½" o.d.

CJ440

ELECTRICAL:	
BANDWIDTH	420-470 MHz
GAIN	1.8 dBd
VSWR	1.5:1
FEED IMP	50 ohms
NO GROUND PLANE REQUIRED	
MECHANICAL:	
HEIGHT	19¼"
WEIGHT	1 lb.
MAST	1½" o.d.

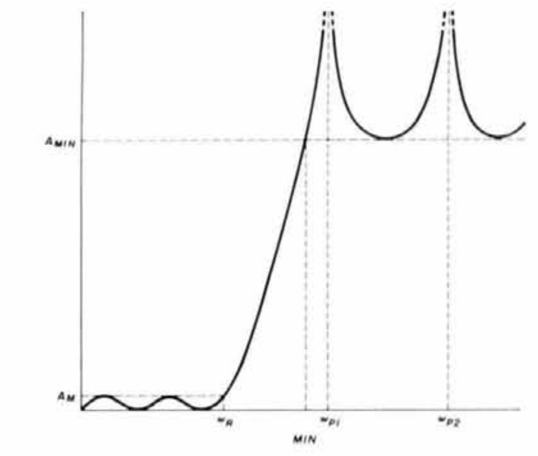


fig. 6A. Cauer elliptic filter. Low pass amplitude response.

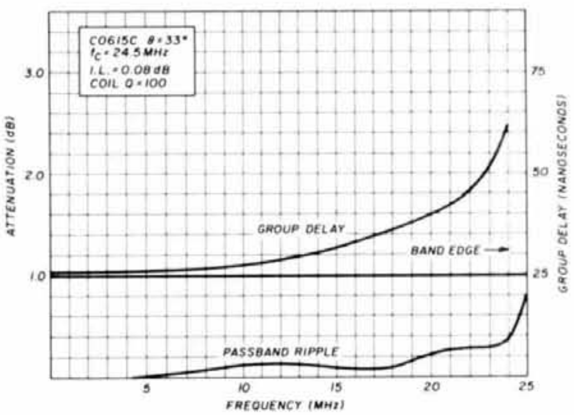


fig. 6B. Insertion loss (passband ripple) and group delay performance (Filter no. 7 from *table 1*). A "Q" of 100 was used in the calculations.

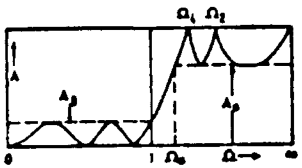
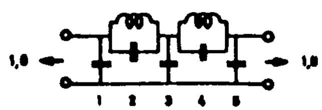
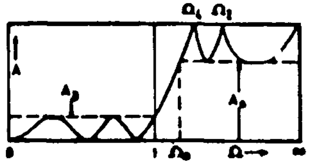
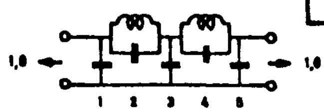
cation is needed for the entire 2- to 30-MHz range, a low-pass, half-octave, Chebyshev filter that has this characteristic is an eighth-order design and is too complex. If the ripple requirement were relaxed, a seventh-order design would be sufficient, but peaks in the ripple would make construction very sensitive to adjustment of component values and their variations with temperature. Duplicating the effort by a factor of 16 filters would be difficult.

compromise filter requirements

Any compromise filter would have to meet the following criteria:

• conform to the electrical design values shown in **table 1**;

C0515

θ	Ω_S	a_S[N]	A_S[db]	C_1	C_2	L_2	Ω_2	C_3	C_4	L_4	Ω_4	C_5
T	∞	∞	∞	1,144	0,000000	1,372	∞	1,972	0,000000	1,372	∞	1,144
06	9,566 772	14,94	129,7	1,142	0,002768	1,368	16,246 767	1,965	0,007259	1,362	10,056 467	1,137
07	8,205 509	14,16	123,0	1,141	0,003771	1,367	13,925 961	1,962	0,009893	1,359	8,624 709	1,135
08	7,185 297	13,50	117,2	1,140	0,004930	1,366	12,185 377	1,959	0,01294	1,355	7,551 553	1,132
09	6,392 453	12,90	112,1	1,139	0,006246	1,365	10,831 605	1,956	0,01641	1,351	6,717 466	1,129
10	5,758 770	12,37	107,5	1,138	0,007720	1,363	9,748 603	1,952	0,02030	1,346	6,050 729	1,125
11	5,240 843	11,90	103,3	1,136	0,009353	1,361	8,862 521	1,947	0,02461	1,340	5,505 703	1,121
12	4,809 734	11,46	99,5	1,135	0,01115	1,359	8,124 130	1,943	0,02936	1,335	5,051 963	1,117
13	4,445 411	11,05	96,0	1,133	0,01310	1,357	7,499 346	1,938	0,03455	1,328	4,668 444	1,112
14	4,133 565	10,68	92,8	1,131	0,01522	1,355	6,963 824	1,932	0,04018	1,321	4,340 101	1,107
15	3,863 703	10,33	89,7	1,130	0,01750	1,352	6,499 710	1,926	0,04627	1,314	4,055 901	1,102
16	3,627 955	10,00	86,9	1,128	0,01995	1,350	6,093 615	1,920	0,05282	1,306	3,807 568	1,096
17	3,420 304	9,70	84,2	1,125	0,02257	1,347	5,735 299	1,913	0,05984	1,297	3,588 776	1,090
18	3,236 068	9,41	81,7	1,123	0,02536	1,344	5,416 798	1,906	0,06735	1,289	3,394 603	1,083
19	3,071 553	9,13	79,3	1,121	0,02832	1,341	5,131 623	1,899	0,07535	1,279	3,221 164	1,076
20	2,923 804	8,87	77,0	1,118	0,03146	1,337	4,875 347	1,891	0,08385	1,269	3,065 351	1,069
21	2,790 428	8,62	74,9	1,115	0,03478	1,334	4,643 295	1,882	0,09287	1,259	2,924 648	1,061
22	2,669 467	8,38	72,8	1,113	0,03827	1,330	4,432 337	1,874	0,1024	1,248	2,796 996	1,053
23	2,559 305	8,15	70,8	1,110	0,04195	1,326	4,239 719	1,865	0,1125	1,237	2,680 696	1,044
24	2,458 593	7,94	68,9	1,107	0,04582	1,322	4,063 150	1,855	0,1232	1,225	2,574 331	1,036
25	2,366 202	7,72	67,1	1,103	0,04988	1,318	3,900 700	1,846	0,1345	1,212	2,476 711	1,026
26	2,281 172	7,52	65,3	1,100	0,05413	1,313	3,750 741	1,835	0,1463	1,199	2,386 829	1,017
27	2,202 689	7,33	63,6	1,096	0,05858	1,309	3,611 883	1,825	0,1588	1,186	2,303 827	1,007
28	2,130 054	7,14	62,0	1,093	0,06323	1,304	3,482 936	1,814	0,1720	1,172	2,226 971	0,9963
29	2,062 665	6,95	60,4	1,089	0,06809	1,299	3,362 873	1,803	0,1858	1,158	2,155 627	0,9856
30	2,000 000	6,78	58,8	1,085	0,07316	1,293	3,250 805	1,791	0,2004	1,143	2,089 247	0,9744
31	1,941 604	6,60	57,4	1,081	0,07845	1,288	3,145 956	1,779	0,2157	1,128	2,027 351	0,9629
32	1,887 080	6,44	55,9	1,076	0,08396	1,282	3,047 649	1,767	0,2317	1,112	1,969 523	0,9509
33	1,836 078	6,27	54,5	1,072	0,08970	1,276	2,955 288	1,754	0,2486	1,096	1,915 395	0,9386
34	1,788 292	6,11	53,1	1,067	0,09567	1,270	2,868 346	1,741	0,2664	1,080	1,864 643	0,9259
35	1,743 447	5,96	51,8	1,062	0,1019	1,264	2,786 358	1,728	0,2850	1,063	1,816 980	0,9128
36	1,701 302	5,81	50,4	1,057	0,1084	1,258	2,708 909	1,715	0,3047	1,045	1,772 152	0,8993
37	1,661 640	5,66	49,2	1,052	0,1151	1,251	2,635 631	1,701	0,3253	1,027	1,729 930	0,8853
38	1,624 269	5,52	47,9	1,047	0,1221	1,244	2,566 192	1,686	0,3470	1,009	1,690 112	0,8710
39	1,589 016	5,37	46,7	1,042	0,1293	1,237	2,500 299	1,672	0,3699	0,9901	1,652 516	0,8563
40	1,555 724	5,24	45,5	1,036	0,1369	1,229	2,437 673	1,657	0,3939	0,9709	1,616 977	0,8412
θ	Ω_S	a_S[N]	A_S[db]	L_1	L_2	C_2	Ω_2	L_3	L_4	C_4	Ω_4	L_5

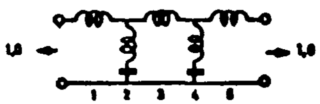
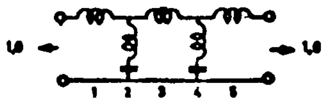
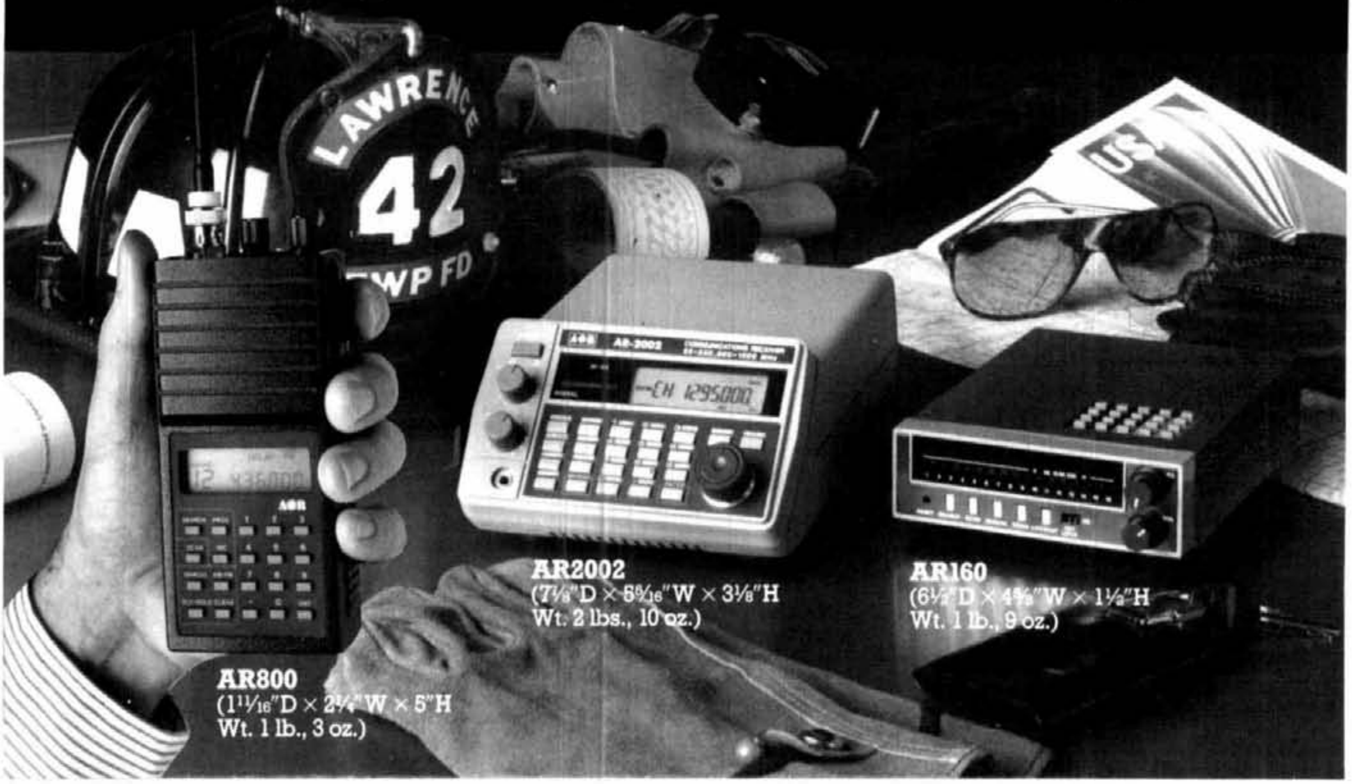



fig. 7. Fifth order Cauer (Elliptic) filter, adapted from: Von R. Seal, Der Entwurf von Filtern mit hilfe des Kataloges normierter Tiefpasse, AEG-Telefunken, 715 Backnang, West Germany, 1968, pages 86-87.

The New AOR Scanners.
Far more performance in a lot less space.



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If, for any reason, you; the original purchaser, are not satisfied with the unit purchased a full refund of the purchase price will be issued if the unit and all accessories are returned to us undamaged within 25 days of the date of original purchase. This policy excludes any additional freight that may be incurred, and in no event modifies or limits the limited warranty.

Reader Service CHECK—OFF Page 106 ✓181

- meet packaging requirements, using the minimum number of poles;
- be easy to design and economical to construct;
- be easy to tune with relatively simple laboratory equipment.

A Cauer (elliptical) filter has an in-band characteristic similar to that of a Chebyshev filter; it also has a more abrupt transition band characteristic than the monotonically increasing attenuation of the Chebyshev approach. A fifth-order Cauer is easier to tune than the more complex eighth-order Chebyshev that would have been required to meet the ripple spec.

The low-pass amplitude response of a typical Cauer filter is shown in **fig. 6A**. Insertion loss has been calculated and verified at 0.01 dB for the low-pass and approximately 1 dB for the bandpass design. The group delay remains relatively constant over the range as shown in **fig. 6B**.

Where: A_M is the magnitude of the passband ripple (expressed in nepers or dB):

$$dB = -10 \log (1 - \rho) \qquad (1)$$

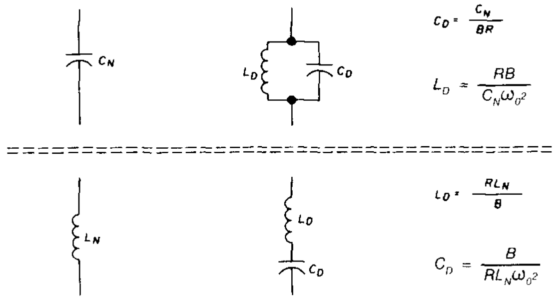
Note: while ρ is usually expressed as a percentage, the decimal value should be used in **eqn. 1**. W_R is the frequency of the ripple bandwidth normalized for ω = 1 radian. A_{MIN} is the minimum stopband attenuation in dB. ω_{MIN} is earliest frequency which has less or equal amplitude than A_{MIN} (ω_{MIN} is normalized to ω_R). ω_{P1}, ω_{P2}, are the frequencies of attenuation peaks, normalized to ω_R.

bandpass design

The 16- to 24-MHz design (filter No. 7) is analyzed according to data supplied in **table 1**. This filter is one of four (1, 3, 5, and 7) that have more stringent slope requirements on the low-frequency side of their 70-dB attenuation (a half-bandwidth ratio between the center frequency and the lower 70-dB attenuation point of 3.75:1), as can be seen from **table 1**. From the normalized tables of elliptic filters in **fig. 7**, we find that a fifth-order filter will meet the bandpass requirements, including the 0.1-dB passband ripple. The required attenuation will be determined by the conservative choice of θ = 21 degrees* which will provide more than 74.9 dB for all frequencies located 2.79 half-bandwidths away from the center frequency, more than enough to satisfy the above requirements. In addition, the design margin of 4.9 dB over the 70-dB design requirement is intended to compensate for possible theoretical vs. practical problems which may evolve during the implementation of the network. The filter has now been identified. The schematic diagram and its normalized electrical values can be extracted from **fig. 7**. This filter can also be defined as C0515 θ = 21

* θ = value of modular angle.

table 2A. Equations used for the transformations from the lowpass to bandpass Elliptic functions.



$$C_D = \frac{C_N}{BR}$$

$$L_D = \frac{RB}{C_N \omega_0^2}$$

$$L_D = \frac{RL_N}{B}$$

$$C_D = \frac{B}{RL_N \omega_0^2}$$

To transform from the normalized lowpass element values, apply the following equations: (See program B)

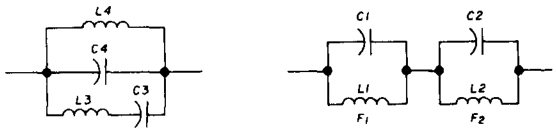
Where: B = Radian bandwidth of the design

ω_0 = Radian center frequency (geometric mean)

R = Value of input (source) resistor if elliptic or to the value of the output (load) resistor if other than an elliptic design

Table 2A

To transform from a parallel resonant-series resonant combination into two parallel resonant circuits in series with each other perform the following operations: (See program A)



$A = L_3 \cdot L_4\, C_3 \cdot C_4$

$B = C_4 \cdot L_4 + C_3 \cdot L_3 + L_4 \cdot C_3$

$E = L_3 \cdot L_4 \cdot C_3$

$P = (B + SQR(B \cdot B - 4A)) \cdot 5$

$C_1 = (A \cdot A - A \cdot P \cdot P)/(A \cdot L_4 \cdot P - E \cdot P \cdot P)$

$C_2 = (A - P \cdot P)/(E - P \cdot L_4)$

$L_1 = (A \cdot L_4 - P \cdot E)/(A - P \cdot P)$

$L_2 = (P \cdot E - L_4 \cdot P \cdot P)/(A - P \cdot P)$

The two resonant frequencies F_1 and F_2 are

$$F_1 = \frac{1}{2\pi\sqrt{L_1 C_1}}$$

$$F_2 = \frac{1}{2\pi\sqrt{L_2 C_2}}$$

table 2B. IBM-Compatible programs used for simplifying the transformation calculations.

A

```
10 DIM Q$(3)
20 PI=3.14159
30 LPRINT "THIS PROGRAM TRANSFORMS A PARALLEL RESONANT - SERIES RESONANT"
40 LPRINT "COMBINATION"
50 LPRINT "INTO  TWO PARALLEL RESONATE CIRCUITS IN SERIES WITH EACH OTHER"
60 CLS
70 PRINT "ENTER THE VALUE (IN uHENRIES) OF THE SERIES INDUCTOR"
80 INPUT L3
90 PRINT "ENTER THE VALUE (IN uHENRIES) OF THE PARALLEL INDUCTOR"
100 INPUT L4
110 PRINT "ENTER THE VALUE (IN pFARADS) OF THE SERIES CAPACITOR"
120 INPUT C3
130 PRINT "ENTER THE VALUE (IN pFARADS) OF THE PARALLEL CAPACITOR"
140 INPUT C4
150 A=L3*L4*C3*C4
160 B=C4*L4+C3*L3+L4*C3
170 E=L3*L4*C3
180 P=(B+SQR(B*B-4*A))*.5
190 C1=(A*A-A*P*P)/(A*L4*P-E*P*P)
200 C2=(A-P*P)/(E-P*L4)
210 L1=(A*L4-P*E)/(A-P*P)
220 L2=(P*E-L4*P*P)/(A-P*P)
230 F1=1/(2*PI*SQR(L1*C1))*1000
240 F2=1/(2*PI*SQR(L2*C2))*1000
250 CLS
260 LPRINT "THE INPUT VALUES ARE:"
270 LPRINT "INDUCTANCE (uHENRIES)          CAPACITANCE (pFARADS) SERIES CIRCUIT"
280 LPRINT
290 LPRINT USING "+#.######^^^^                      ";L3;C3
300 LPRINT
310 LPRINT "INDUCTANCE (uHENRIES)        CAPACITANCE (pFARADS) PARALLEL CIRCUIT"
320 LPRINT
330 LPRINT USING "+#.######^^^^                      ";L4;C4
340 LPRINT
350 LPRINT "THE VALUES FOR THE TWO RESONANT CIRCUITS ARE:"
360 LPRINT
370 LPRINT "INDUCTANCE uH        CAPACITANCE pF       FREQUENCY MHz"
380 LPRINT
390 LPRINT USING "+#.######^^^^        ";L1;C1;F1
400 LPRINT USING "+#.######^^^^        ";L2;C2;F2
410 LPRINT
420 LPRINT
430 PRINT "DO YOU WISH TO ENTER ANOTHER SET OF DATA? Y/N"
440 INPUT Q$
450 CLS
460 IF Q$="Y" THEN 60
470 IF Q$="N" THEN 500
480 CLS
490 GOTO 430
500 CLS
510 END
```

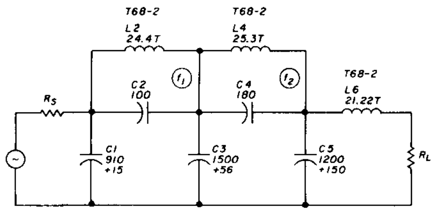
B

```
10 CLS
20 LPRINT "THIS PROGRAM COMPUTES THE NORMALIZED LOWPASS TO WIDEBAND"
30 LPRINT "BANDPASS TRANSFORMATION"
40 DIM A$(3)
50.P=3.14159
60 PRINT "ENTER THE VALUE OF THE CENTER FREQUENCY IN Hz"
70 INPUT F
80 PRINT "ENTER THE VAULE OF THE SOURCE RESISTOR"
90 INPUT R
100 PRINT "ENTER THE VALUE OF THE RIPPLE BANDWIDTH IN Hz"
110 INPUT B
120 LPRINT "THE CENTER FREQUENCY IS ";F;" Hz"
130 LPRINT "THE RIPPLE BANDWIDTH IS ";B;" Hz"
140 LPRINT "THE SOURCE RESISTOR IS ";R;" OHMS"
150 CLS
160 PRINT "ENTER THE VALUE OF THE NORMALIZED LOWPASS CAPACITOR (IN FARADS)"
170 INPUT C
180 C1=C/(2*P*B*R)
190 L1=R*2*P*B/(C*4*P*P*F*F)
200 LPRINT
210 LPRINT
220 LPRINT "NORMALIZED          L1                    C1"
230 LPRINT "LOWPASS C        HENRIES              FARADS"
240 LPRINT
250 LPRINT USING "+#.#####^^^^      ";C,L1,C1
260 LPRINT
270 PRINT "ENTER THE VALUE OF THE NORMALIZED LOWPASS INDUCTOR (IN HENRIES)"
280 INPUT L
290 C1=2*P*B/(L*R*4*P*P*F*F)
300 L1=R*L/(2*P*B)
310 LPRINT
320 LPRINT "NORMALIZED          L1                    C1"
330 LPRINT "LOWPASS L        HENRIES              FARADS"
340 LPRINT
350 LPRINT USING "+#.#####^^^^      ";L,L1,C1
360 LPRINT
370 CLS
380 PRINT "DO YOU WISH TO ENTER ADDITIONAL VALUES? Y/N"
390 INPUT A$
400 IF A$="Y" THEN 150
410 IF A$="N" THEN 440
420 CLS
430 GOTO 380
440 CLS
450 END
```

table 3. Calculated and practical (given in parentheses) values for the eight, half-octave lowpass filters used in the transceiver.

#1

f_C = 3.2 MHz BAND = 2 – 3 MHz
Reject 6 MHz by 60 dB
f_C: Reject = 1.875:1
Use A C0615C θ = 35° Filter
Minimum Attenuation At 6 MHz = 67.8 dB
Normalized values are:



R_S = 1	R_L = 1
C_1 = 0.9316	C_2 = 0.1082
L_2 = 1.368	C_2 = 1.564
C_4 = 0.1880	L_4 = 1.466
C_5 = 1.371	L_6 = 1.033
f_1 = 2.5989	f_2 = 1.9051

the denormalized values are:

R_S = 50	R_L = 50
C_1 = 926.7* pF (910 + 15)**	C_2 = 107.6 pF (100)
L_2 = 3.40 μH	C_3 = 1556 pF (1500 + 56)
C_4 = 187 pF (180)	L_4 = 3.65 μH
C_5 = 13.64 pF (1200 + 150)	L_6 = 2.568 μH
f_1 = 8.316 MHz	f_2 = 6.0963 MHz

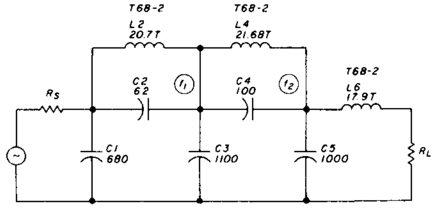
tune L_2, C_2 for an attenuation peak at f_1
tune L_4, C_4 for an attenuation peak at f_2

*Design value
**Actual value used

#2

f_C = 4.5 MHz Band = 3 — 4 MHz
Reject 9 MHz
f_C: Reject ratio = 2:1

Use a C0615C θ = 32° filter
Minimum attenuation at 9 MHz = 72.8 dB
Normalized values are:



R_S = 1	R_L = 1
C_1 = 0.9492	C_2 = 0.08912
L_2 = 1.393	C_3 = 1.601
C_4 = 0.1540	L_4 = 1.518
C_5 = 1.395	L_6 = 1.034
f_1 = 2.8385	f_2 = 2.06813

The denormalized values are:

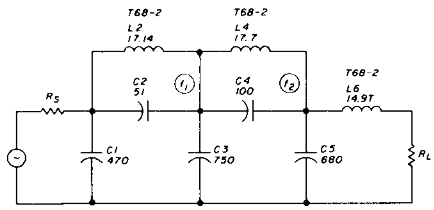
R_S = 50	R_L = 50
C_1 = 671.4 pF (680)	C_2 = 63 pF (62)
L_2 = 2.46 μH	C_3 = 1132.4 pF (1100)
C_4 = 108.9 pF (100)	L_4 = 2.68 μH
C_5 = 986.7 pF (1000)	L_6 = 1.83 μH
f_1 = 12.773 MHz	f_2 = 9.3066 MHz

tune L_2, C_2 for an attenuation peak at f_1
tune L_4, C_4 for an attenuation peak at f_2

#3

F_C = 6.5 MHz Band = 4 – 6 MHz
Reject 12 MHz
f_C: Reject = 1.846:1

Use a C0615C θ = 35° filter
Minimum attenuation at 12 MHz = 67.8 dB
Normalized values are:



R_S = 1	R_L = 1
C_1 = 0.9316	C_2 = 0.1082
L_2 = 1.368	C_3 = 1.564
C_4 = 0.1880	L_4 = 1.466
C_5 = 1.371	L_6 = 1.033
f_1 = 2.59897	f_2 = 1.9051

The denormalized values are:

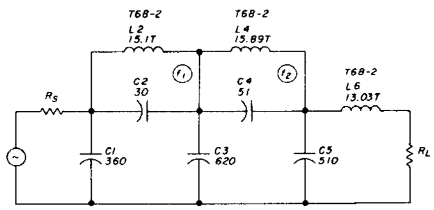
R_S = 50	R_L = 50
C_1 = 456.2 pF (470)	C_2 = 53 pF (51)
L_2 = 1.675 μH	C_3 = 766 pF (750)
C_4 = 92 pF (100)	L_4 = 1.79 μH
C_5 = 671.4 pF (680)	L_6 = 1.265 μ
f_1 = 16.893 MHz	f_2 = 12.383 MHz

tune L_2, C_2 for an attenuation peak at f_1
tune L_4, C_4 for an attenuation peak at f_2

#4

f_C = 8.5 MHz Band = 6 – 8 MHz
Reject 18 MHz
f_C: Reject = 2.11:1

Use a C0615C θ = 31° filter
Minimum attenuation at 18 MHz = 74.5 dB
Normalized values are:



R_S = 1	R_L = 1
C_1 = 0.9547	C_2 = 0.08325
L_2 = 1.400	C_2 = 1.612
C_4 = 0.1436	L_4 = 1.535
C_5 = 1.402	L_6 = 1.034
f_1 = 2.9287	F_2 = 2.1298

The denormalized values are:

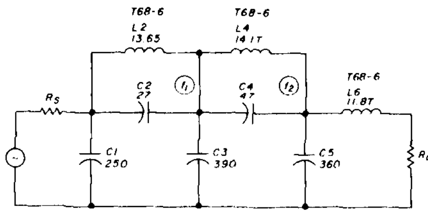
R_S = 50	R_L = 50
C_1 = 357.5 pF (360)	C_2 = 31.2 pF (30)
L_2 = 1.31 μH	C_3 = 603.7 pF (620)
C_4 = 53.8 pF (50)	L_4 = 1.44 μH
C_5 = 525 pF (510)	L_6 = 0.968 μH
f_1 = 24.894 MHz	f_2 = 18.1033 MHz

tune L_2, C_2 for an attenuation peak at f_1
tune L_4, C_4 for an attenuation peak at f_2

#5

f_C = 12.5 MHz Band = 8 – 12 MHz
Reject 24 MHz
f_C: Reject = 1.92:1

Use a C0615C θ = 34° filter
Minimum attenuation at 24 MHz = 69.4 dB
Normalized values are:



R_S = 1	R_L = 1
C_1 = 0.9377	C_2 = 0.101
L_2 = 1.376	C_3 = 1.577
C_4 = 0.1761	L_4 = 1.484
C_5 = 1.379	L_6 = 1.034
f_1 = 2.6741	f_2 = 1.9561

The denormalized values are:

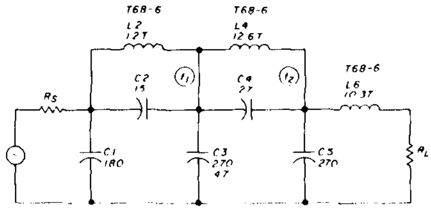
R_S = 50	R_L = 50
C_1 = 238.8 pF (250)	C_2 = 25.9 pF (27)
L_2 = 0.876 μH	C_3 = 401.6 pF (390)
C_4 = 44.8 pF (47)	L_4 = 0.945 μH
C_5 = 351.1 pF (360)	L_6 = 0.658 μH
f_1 = 33.426 MHz	F_2 = 24.451 MHz

tune L_2, C_2 for an attenuation peak at f_1
tune L_4, C_4 for an attenuation peak at f_2

#6

f_C = 16.5 MHz Band = 12 – 16 MHz
Reject 36 MHz
f_C: Reject = 2.18:1

Use a C0615C θ = 30° filter
Minimum attenuation at 36 MHz = 76.3 dB
Normalized values are:



R_S = 1	R_L = 1
C_1 = 0.96	C_2 = 0.0776
L_2 = 1.408	C_3 = 1.623
C_4 = 0.1337	L_4 = 1.551
C_5 = 1.410	L_6 = 1.034
f_1 = 3.02499	f_2 = 2.19586

The denormalized values are:

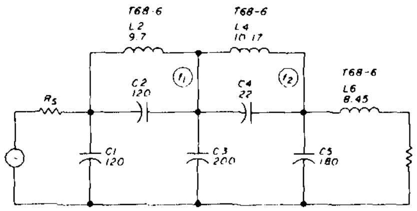
R_S = 50	R_L = 50
C_1 = 185.19 pF (180)	C_2 = 14.9 pF (15)
L_2 = 0.679 μH	C_3 = 313.1 pF (270 + 47)
C_4 = 25.8 pF (27)	L_4 = 0.748 μH
C_5 = 272 pF (270)	L_6 = 0.4987 μH
f_1 = 49.9123 MHz	f_2 = 36.2317 MHz

tune L_2, C_2 for an attenuation peak at f_1
tune L_4, C_4 for an attenuation peak at f_2

#7

f_C = 24.5 MHz Band = 16 – 24 MHz
Reject 48 MHz
f_C: Reject = 1.96:1

Use a C0615C θ = 33° filter
Minimum attenuation at 48 MHz = 71.1 dB
Normalized values are:



R_S = 1	R_L = 1
C_1 = 0.9436	C_2 = 0.09523
L_2 = 1.385	C_3 = 1.589
C_4 = 0.1648	L_4 = 1.501
C_5 = 1.387	L_6 = 1.034
f_1 = 2.75377	f_2 = 2.0103

The denormalized values are:

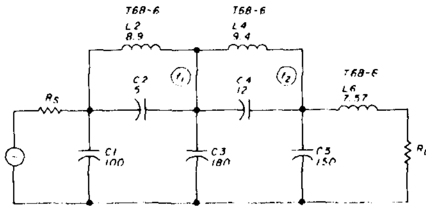
R_S = 50	R_L = 50
C_1 = 122.6 pF (120)	C_2 = 12.37 pF (12)
L_2 = 0.449 μH	C_3 = 206.4 pF (200)
C_4 = 21.4 pF (22)	L_4 = 0.487 μH
C_5 = 180.2 pF (180)	L_6 = 0.336 μH
f_1 = 67.467 MHz	f_2 = 49.252 MHz

tune L_2, C_2 for an attenuation peak at f_1
tune L_4, C_4 for an attenuation peak at f_2

#8

f_C = 30.5 MHz Band = 24 – 30 MHz
Reject 72 MHz
f_C: Reject = 2.36:1

Use a C0615C θ = 27° filter
Minimum attenuation at 72 MHz = 82 dB
Normalized values are:



R_S = 1	R_L = 1
C_1 = 0.9747	C_2 = 0.06212
L_2 = 1.429	C_3 = 1.654
C_4 = 0.1066	L_4 = 1.595
C_5 = 1.430	L_6 = 1.035
f_1 = 3.3568	f_2 = 2.4244

The denormalized values are:

R_S = 50	R_L = 50
C_1 = 101.7 pF (100)	C_2 = 6.5 pF (5)
L_2 = 0.373 μH	C_3 = 172.6 pF (180)
C_4 = 11.1 pF (12)	L_4 = 0.416 μH
C_5 = 149.2 pF (150)	L_6 = 0.270 μH
f_1 = 102.38 MHz	f_2 = 73.944 MHz

tune L_2, C_2 for an attenuation peak at f_1
tune L_4, C_4 for an attenuation peak at f_2

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degrees, which means that it is a Cauer (C, elliptic) design of a fifth order (05), with ρ = 15 percent (see **fig. 6**), and θ = 21°, as described above. Its normalized design parameters are shown respectively in **fig. 8A** for the normalized low-pass element values and in **fig. 8B** for the denormalized bandpass element values. The transformation is performed with the help of several equations, which are listed in **table 2A**. The design for the remaining seven bandpass filters is performed in a similar manner.

Information about applying denormalizing equations is provided in the references, which will follow at the end of Part 2 of this article. To make the job easier, the transformation equations from the low-pass to the bandpass elliptic filter have been applied to computer programs for the IBM and compatibles, and are included in **table 2B**.

The final transformation for the C0515 θ = 21-degree filter is shown in **fig. 9**. Note that f_1, f_2, f_3, and f_4 are frequencies where the attenuation characteristic peaks. Knowledge of where they occur is very important for tuning the filters. This is accomplished by adjusting the individual inductances or capacitors for maximum attenuation at the respective frequencies.

Reasonable care should be exercised in choosing components as close as possible to the theoretical values. For example, actual capacitor values were chosen near the theoretical values and inductance tuning was accomplished by spreading or compressing the windings on the toroidal cores. This method worked well for all the networks.

low-pass design

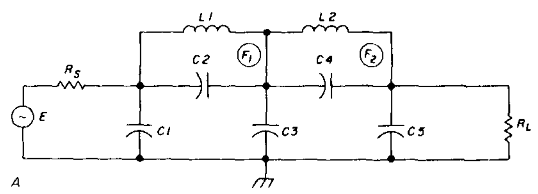
We have seen how a bandpass, half-octave Cauer (elliptic) filter, C0515, can be designed to meet stringent requirements and yet be easy to build. The design of the equivalent low-pass filters from **table 1** is performed in much the same way. Since a sixth-order filter is required for the low-pass bank to achieve the design requirements, the identification for this filter will be C0615c. (The "c" at the end indicates an equal source and load impedance.) The elliptical tables mentioned earlier (**table 2**) and **eqn. 2** and **3** below are used to denormalize the low-pass filter.

$$C_D = \frac{C_N}{R_S\,\omega_C} \tag{2}$$

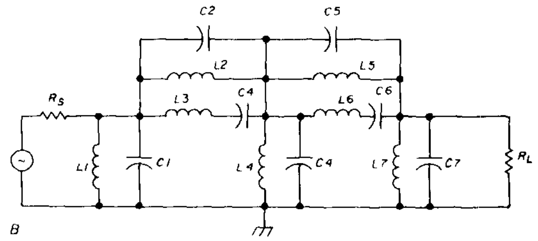
where C_D and L_D are the denormalized values. Where C_N and L_N are the normalized values,

$$L_D = \frac{L_N\,R_S}{\omega_C} \tag{3}$$

R_S is the value of the source resistor, and ω_C is the cutoff frequency in radians.

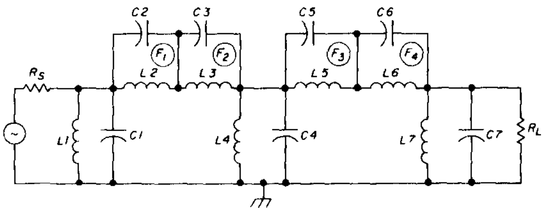


R_S = 1.000 ohms	R_L = 1.000 ohms
C_1 = 1.115 F.	C_2 = 0.03478 F.
C_3 = 1.882 F.	C_4 = 0.09287 F.
C_5 = 1.061 F.	
L_1 = 1.334 H.	L_2 = 1.259 H.
F_1 = 4.643 Radians	F_2 = 2.925 Radians



R_S = 50.000 ohms	R_L = 50.000 ohms
C_1 = 412.69pF	C_2 = 53.79pF
C_3 = 12.87pF	C_4 = 696.58pF
C_5 = 56.99pF	C_6 = 34.37pF
C_7 = 392.71pF	
L_1 = 0.161μH	L_2 = 1.23μH
L_3 = 5.16μH	L_4 = 0.0953μH
L_5 = 1.16μH	L_6 = 1.93μH
L_7 = 0.169μH	

fig. 8. Schematic diagram and values for (A) normalized lowpass element, and (B) denormalized bandpass element of the bandpass filter.



R_S = 50.000 ohms	R_L = 50.000 ohms
C_1 = 412.69pF	C_2 = 15.01pF
C_3 = 90.27pF	C_4 = 696.58pF
C_5 = 44.60pF	C_6 = 149.88pF
C_7 = 392.71pF	
L_1 = .160μH	L_2 = 0.736μH
L_3 = 4.42μH	L_4 = 0.0953μH
L_5 = 0.443μH	L_6 = 1.49μH
L_7 = 0.169μH	
F_1 = 47.89MHz	F_2 = 7.966MHz
F_3 = 35.8MHz	F_4 = 10.65MHz

fig. 9. Final transformation for CO515 θ = 21° filter. Knowing where the peaks in attenuation are will greatly help in tuning the filter.

The final practical model for C0615c θ = 31 degrees is shown in **fig. 10**. The inductors were wound by hand on Micrometals™ toroidal cores, as we will see later.

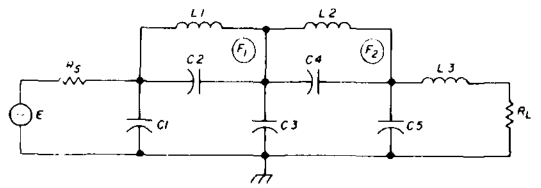


fig. 10. Realization of the practical lowpass, half-octave, Cauer (Elliptic) filter model. This model has been identified as CO615c. See text.

Table 3 shows the calculated and practical values for all eight half-octave, low-pass filters used in the transceiver. The numbers in parentheses are practical values as used in the implementation. Silver-dipped mica capacitors with a tolerance of ±5 percent were used throughout the networks. Breakdown voltages have been chosen at 250 volts.

Part 2 of this article will deal with the final implementation of the filters in the transceiver. In addition, automation and switching for the entire system will be discussed in sufficient detail to allow readers to design their own circuits.

ham radio

amateur packet radio networking and protocols, part 1

Introduction to a viable communications mode

Much has been written about Amateur packet radio over the past few years. Although the features and advantages of packet radio are well known by most active Amateurs, remarkably few operators understand how packet radio really works.

This three-part series describes the workings of packet radio. A basic reader understanding of the subject and some knowledge of the components and operating procedures are assumed.

I will cover two related areas that are integral to packet radio, networking and protocols. These concepts are viable when used with hardware systems (such as terminals, Terminal Node Controllers, and radios). All three components (hardware, networks, and protocols) work together, and a failure in one can result in total breakdown. The series introduces the subject of networks and protocols, explores the various options in each, and describes the common systems in use today.

network basics

A single packet station is useless for communications; two or more stations are needed. In terms of digital communications, a network can be defined as a collection of devices linked together so that one station can talk with *any other* station in the network. It is difficult to decide on and implement a system allowing for maximum flexibility and throughput while minimizing complexity and cost.

In the simplest case, a packet network consists of a few stations within direct communications range of each other on a single frequency (see **fig. 1**). A more complex network involves digipeating (simplex packet repeaters) to extend a station's communication range and gateways for accessing stations with different capabilities as shown in **fig. 2**.

This situation is not ideal because of congestion, range limitations, and other problems. Before more advances can be made in packet networking, additional work is needed in the area of protocols. Present day packet has stretched the current protocols to their limit, and much is being done to develop new ones.

multiplexing

Since packet operation occurs on agreed upon single frequencies, a method is needed that allows stations to access the frequencies in an orderly manner. Without this, operators using the frequency would collide with other users.

The method used is *multiplexing.* Multiplexing lets a group of users share a communication medium. In its ideal form, each user should be unaware that he is sharing the channel. The two forms of multiplexing that concern packet radio operators are Time Division Multiplexing (TDM) and Frequency Division Multiplexing (FDM).

FDM

FDM allows each transmitting user to have a separate channel for communications. The radio stations on fm stereo are a good example of this. Each has its own frequency and occupies it continuously. This would be wasteful in packet radio which uses the channel only for brief periods. There are usually a set number of frequencies allocated for communications (such as 145.01 MHz, 145.03 MHz, and 14.103 MHz), and the user selects one before beginning.

Once a station starts transmitting over a certain channel, it usually stays on it until the communications session is over. This is known as static FDM because the stations do not switch between different frequencies during connection. Pure FDM operation does not provide a very versatile network.

TDM

TDM allows users to share a common channel without interfering with others. Each station transmits one after the other while users with no traffic stand by. The frequency is allotted by time to users with traffic to send; one station will transmit for a time and then

By **Jonathan L. Mayo, KR3T**, 3908 Short Hill Drive, Allentown, Pennsylvania 18104

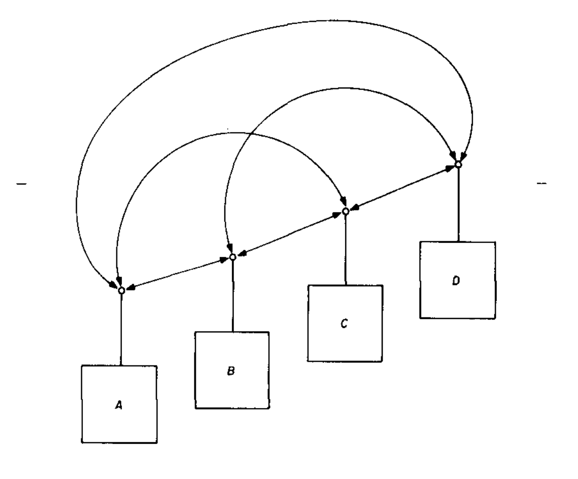


fig. 1. A diagram of a simple network in which all stations can communicate with each other directly.

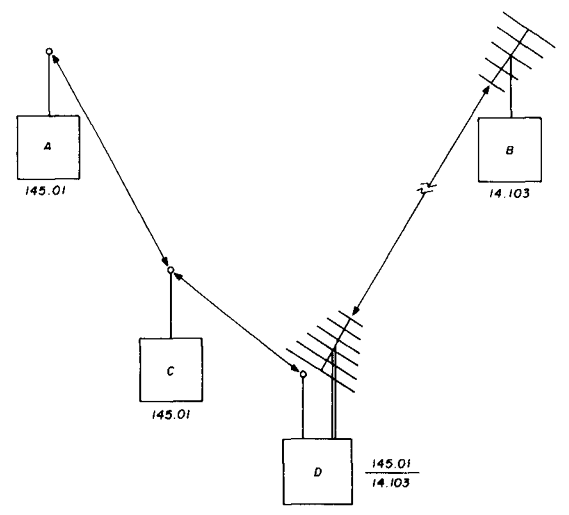


fig. 2. A diagram of a more advanced network in which a digipeater (Station C) and HF Gateway (Station D) are utilized to allow Station A on two meters to communicate with Station B on twenty meters.

be followed by another station. But, how does a station know when to transmit?

Three systems control the access of individual stations to a channel: random access, polling, and token passing.

random access

Packet radio uses random access. Individual stations follow specific rules to enter the system and must be able to determine if the channel is clear. The method used is CSMA/CD (Carrier Sense Multiple Access with Collision Detection). The station monitors (senses) the channel and checks to see if it is clear when it has traffic to send. If it is clear, the station transmits. A successful transmission is acknowledged by the destination station. If the channel is not clear, the station waits and transmits when it is. If two or more stations transmit at the same time a collision might occur. If this happens, the stations involved will receive no acknowledgment and must wait a random length of time to retransmit. Ideally, one station should have a shorter random wait, capture the channel first, and avoid another collision.

Packet radio uses both FDM and TDM, permitting operators to transmit and receive simultaneously. The channel (frequency) selected by FDM affects both the range and speed of data transmission. For example, channels in the 20-meter band have a large range but limited speed, and channels in the 2-meter band have a limited range but support much higher speeds.

TDM lets many operators share the same channel using CSMA/CD. For this to work, all stations on the channel must be within *hearing* range of each other.

Two Amateurs can occupy a higher frequency channel at the same time without interference as a result of Space Division Multiplexing or SDM. Stations in both California and Pennsylvania can transmit on 145.01 MHz simultaneously because the two signal paths do not cross. Factors in SDM are propagation, radiation patterns, and physical obstructions. Such effects are fairly constant and predictable on the VHF/UHF bands.

polling and token passing

In a polling system, a master station asks the others on the network channel if they have traffic to send. The channel is cleared by the master station for each to transmit in turn. Other stations must wait for clearance before they may begin transmission. Token passing is a similar form of TDM. In a token passing system, a single electronic *token* (a special binary sequence) is passed from station to station until it arrives at one with traffic to send. The station holds onto the token and transmits. The frequency stays clear because only the station with the token is allowed to transmit. When the station has finished transmitting, it passes the token to the next station on the network. Depending on network configuration, individual stations may communicate with each other directly or via the master station.

Polling systems lack popularity in packet radio because a master station with a fairly powerful computer and reliable communication throughout the network is needed to track users and their status. Packet users tend to drop in and out quickly and the radio links between stations vary in quality. To work effectively, polling network conditions must be regimented beyond what most Amateurs can provide. Another draw-

back is the amount of *overhead* (information that must be added to the basic data) required for destination routing.

protocols

Now that networking concepts have been covered, we can take a look at what makes packet radio work. How do individual stations know how to communicate with each other? What happens if data is sent but for some reason is not received? What if the data arrives garbled? How does a station know who data is for? How do digipeaters know which data to retransmit? The answers to these and other questions are found in the protocol.

Protocols define how data is packaged, what actions are taken under certain conditions, and when the actions are to be executed. The goal is to get data from its source to its destination as quickly, efficiently, and accurately as possible. Steps involved in using packet protocols follow.

Assuming a station can access the network, it must communicate its intention to transmit with a connect request. If the selected station is available, it will acknowledge the request and the two stations will connect. Once this happens, the information they send to each other is received error-free. When transmission is complete, the stations disconnect and are ready to contact others.

These and other processes are handled by protocols. *A protocol is a predefined series of steps followed to accomplish a task.* An illustration of a random access packet radio protocol is a normal 2-meter fm phone contact.

After first listening to see if the frequency is clear, call the station you want to contact. (AA3F AA3F this is KR3T. Do you read me?) Keep calling until a response is received or you decide to stop your tranmission. If he responds (KR3T this is AA3F. Go ahead.) you have established a connection. You would then transmit your information. (AA3F this is KR3T. Meet me at the mall in 5 minutes.) Give the receiving and sending stations' callsigns so AA3F knows the message is for him (The FCC likes this.). If AA3F acknowledges (KR3T this is AA3F. Roger.) the message was received. If AA3F doesn't respond in a reasonable length of time or asks for a repeat, send it again. End your transmission with a disconnect request (AA3F this is KR3T. 73.) and AA3F will respond (KR3T this is AA3F. 73.). You have just ended the connection and can place another call.

The same system applies to packet radio. A connection is established, information is transferred (or retransmitted when not received properly), and the connection is canceled. Keep in mind that this is a generalization. The protocol must be able to determine when information is received incorrectly, keep track of the connection status, translate data, assure device compatibility, and much more. A detailed look at protocol organization follows.

OSI/RM

Any network, packet or not, consists of many different components and functions: terminals, codings, voltages, error checking, connecting, relaying, and disconnecting. Networks become complex as their capabilities increase. The International Standards Organization (ISO) developed a reference model for networks. Known as the Open Systems Interconnection Reference Model (OSI/RM), it is designed to aid the information exchange between systems. They can be as simple as a current loop teletype or as complex as a worldwide network. *Open Systems* are those open for communications (such as a packet radio station). The OSI/RM separates network functions into levels based on their purpose.

Each OSI/RM level transfers data between the one directly above and below it. It interfaces at the point where data is transferred between levels. Data originates at the highest level and is passed down serially through each one. The information is processed by each level's protocol until it reaches the lowest implemented level. When the data is received, the path is reversed and it is sent back up the levels. Each level removes any additional information added by its equivalent at the sending station. When the data reaches the point of origin, it looks exactly as it did when entered into the network.

Each level operates independently. The only exchange of information between levels occurs at interface points. Every level has a protocol that may be changed without affecting the rest. The set of levels and associated protocols form the network architecture.

The flexibility and structure of the OSI/RM help to maintain compatibility between packet systems. The OSI/RM is divided into seven levels, each responsible for particular tasks and named according to its function. To refer to a level by number, use the word *level*, and when referring to a level by name, use the word *layer*. The seven OSI/RM layers are: physical, data link, network, transport, session, presentation, and application (see **fig. 3**).

Level 1 is the physical layer. It is responsible for the transparent transmission of bit streams across the physical interconnection between systems. This connection can be operated in either simplex, half duplex, or full duplex. The bits must arrive in the same order in which they were sent. Specifications for this layer include mechanical (plugs and dimensions), electrical (voltage and current levels), functional (the meaning of different voltage levels), and procedural (rules and sequences).

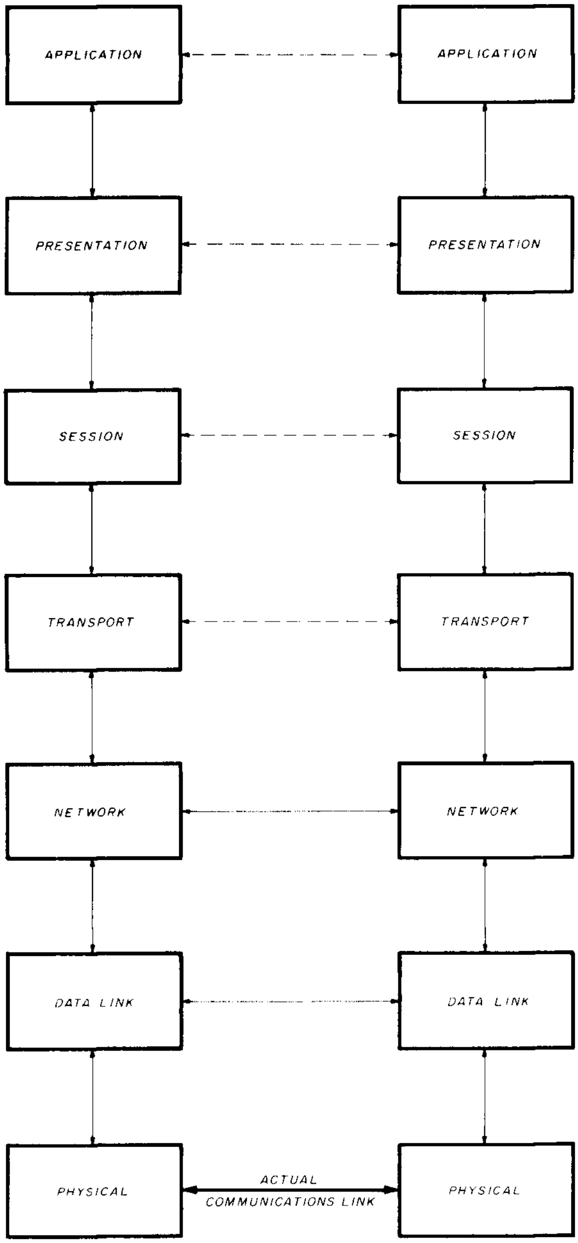


fig. 3. The levels of the OSI/RM. The solid arrows indicate layers that must be implemented at each node. The layers indicated by dashed arrows must be implemented only at the originating and destination nodes.

Level 2, the data link layer, shields the higher levels from the characteristics of the physical layer. It provides reliable transmission of data and should contain some form of error detection and correction. This layer must be independent of the data sent and may not alter it in any way. It must accept data and break it into segments for transmission. When the segmented data is combined with protocol information, a frame is formed. The frame must be delimited (allow for recognition of the beginning and end of the frame) and also be transparent (to be looked at only as a series of bits).

The frame is checked for accuracy upon reception and, if an error is found, retransmitted by the last station. Frames must be delivered in the same order they were sent. The standard level 2 protocol is HDLC (High-level Data Link Control). A subset of HDLC is used in most packet radio data link layer protocols.

The network layer, Level 3, provides transparent transfer of all data submitted by the transport layer, Level 4. Hence Level 4 is not directly involved in the connections between communicating systems. The systems may be connected point to point (direct) or have many nodes in the path. The network layer provides the routing functions needed to transfer data from one system to another; each system may act as a relay. Routing methods are not covered in the OSI/RM. Level 3 of the X.25 standard is one standard protocol for the network layer.

The transport layer arranges the information in the correct order if packets arrive out of sequence. It handles only communications between the origination and destination, and not relay stations that might be used by the network layer.

The session layer or Level 5 is responsible for initiating and terminating communications between stations on the network.

Level 6 is the presentation layer and handles data transformation (converting ASCII to Baudot), data and display formatting (a graphics terminal communicating with a hardcopy teleprinter), and syntax selection. If two systems are using incompatible devices, this layer handles the conversions necessary for data transfer.

In Level 7, the application layer, provision is made for proper operation of *application entities* or user-oriented software. Programs or computer functions controlled by the connected system are located here and fall under its protocol(s).

conclusion

We have looked into the operation of Amateur packet radio which uses a random access networking system. You have been introduced to protocols and their function. In Part 2 we will discuss protocols used in packet radio.

If you have any questions or comments, write to me at the address listed or leave a message on CompuServe; my User ID is 72276,2276.

Portions of this series are from my book, *The Packet Radio Handbook*, available from the *ham radio* Bookstore for $14.95 plus $3.50 shipping and handling.

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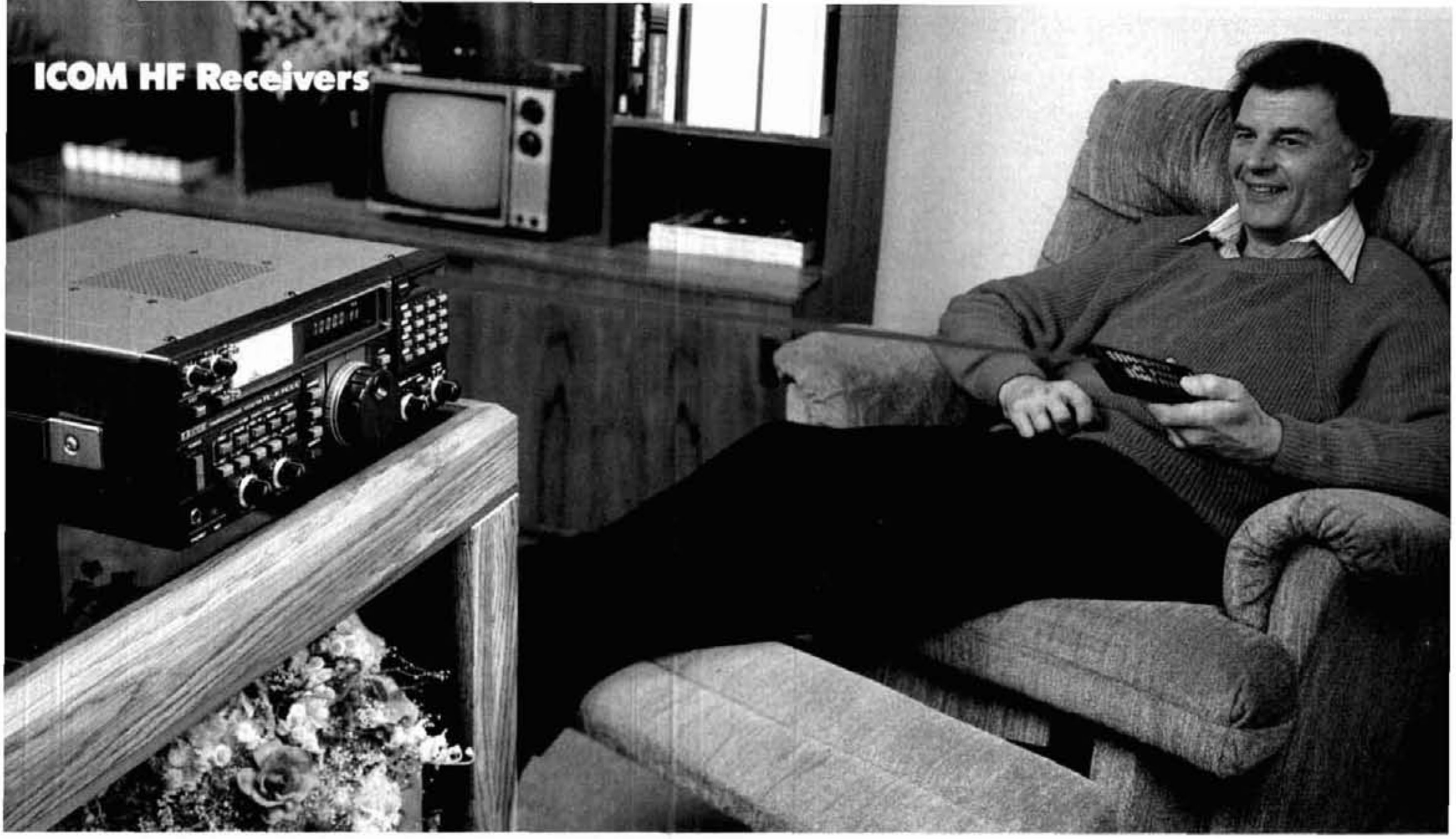

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designing a station for the microwave bands: part 1

Easing your way into a new frontier

In many respects, the microwave bands represent a frontier for Amateurs; above the more familiar regions of VHF and UHF, they're often thought of as a magic domain. Like any physical frontier, they hold promise of untapped possibilities along with seemingly insurmountable barriers. Fascinating in attraction, they hold out hope for open space as the lower bands become full and, ultimately, overcrowded.

Although the microwave bands are often considered useful only for short range contacts where both stations are within line of sight of each other, microwave DX contacts are possible, often over paths which are anything but visual. One of the first surprises I encountered on 10 GHz was being able to clearly copy a 10-milliwatt transmitter by means of scatter off a mountain visible to both receiver and transmitter. The mountain was about 6 miles from both stations, and the direct path was blocked by another mountain. Antennas were a 4-foot dish on one end and a palm-sized horn on the other. Try that on VHF or below with similar sized antennas and equivalent power!

Such contacts aren't limited to scatter from mountains. Microwave communications over extremely long distances are possible via tropospheric and marine ducting *even when lower frequencies are not*. The current 10-GHz world's record of over 1000 miles bears testimony to this. Commercial jetliners offer a large enough (scattering) cross section to be usable for beyond-the-horizon contacts. Even more generally useful, the same tropospheric scattering that provides over-the-horizon communication at VHF is in effect at microwave frequencies. Moonbounce echoes with only a few tenths of a watt of transmitter power have recently been reported at 10 GHz.

At least part of the secret of these seemingly impossible modes of propagation lies in the high antenna gains available with physically smaller, and therefore realizable, antennas. Virtually all of the available transmitter power may be focused into a pencil-like beam. This can provide more signal to a distant receive antenna than would be possible with antennas of the same size at lower frequencies. This may mean, for example, that it would require 100 watts at 144 MHz to provide the same received signal level in 4-foot antennas that 0.1 watt can provide on 10 GHz!

Another part of the secret is narrowband operation. In the past, most Amateur microwave operation was accomplished with free-running, unstable oscillators. This method required wide receiver bandwidths to accommodate the transmitted signal. In addition, the receive local oscillator was often the same one used on transmit, which further increased the necessary bandwidth. While this approach was simple, signal-to-noise ratios were degraded by the additional noise present in the wider bandwidths. Reducing communications bandwidth from, say, 300 kHz to 3 kHz provides the same improvement in signal-to-noise ratio as increasing transmitter power from 1 watt to 100 watts — i.e., 20 dB.

Admittedly, there are difficulties in building and operating an Amateur station in the microwave bands,

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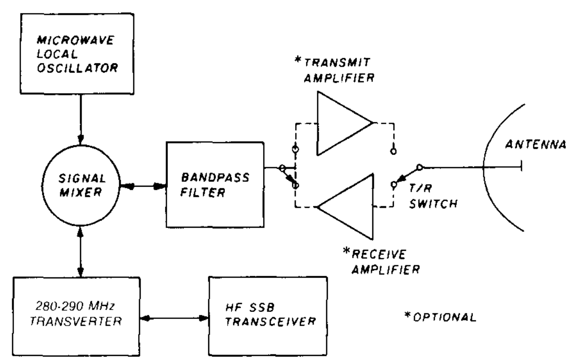

fig. 1. Microwave SSB is generated by converting a 20-30 MHz SSB transceiver first to 280-290 MHz and then to the microwave band using a microwave LO and mixer. A microwave bandpass filter removes unwanted mixing products.

but the excitement of these new regions and the chance for new records and discoveries certainly should draw some of us to further exploration.

As one goes higher in frequency, the number of components that provide gain and power is limited. In addition, maintaining stable and accurate frequencies as well as pointing high-gain antennas accurately are problems to be overcome. The point-to-point nature of microwave DX tends to reduce the likelihood of random contacts. "Round table" QSOs with many stations in different locations may require innovations in Amateur networking, but progress in these areas is currently being made at lower frequencies as digital Amateur radio progresses. The store-and-forward bulletin boards and network nodes currently being used in Amateur packet radio are steps toward a more complete information handling structure. Increased traffic over these channels requires higher data rates and more bandwidth becomes necessary. As our channels support higher information rates, digitized voice and other linear modes can be used. The available spectrum and point-to-point nature of microwave DX links make them ideal for such operation. As Amateur packet radio becomes increasingly popular, implementation of high-speed microwave "backbones" for long-distance cross-country communication becomes an attractive possibility.

design approach

This article describes one approach to designing and implementing a station for the Amateur microwave bands (as well as the two highest UHF bands, 1296 and 2304 MHz). The intent is to promote interest and show how to build a station that can provide all-mode contacts over significant distances. CW, SSB, and NBFM are obvious initial choices, but digital and video modes can be used as well, since the entire system supports linear signal frequency translation.

This open-ended approach minimizes complexity by taking advantage of readily available components and equipment. The intent is simply to demonstrate a practical way for more Amateurs to get on the microwave bands.

The bulk of this series focuses on generation of stable and precise local oscillator signals, since this is a primary hurdle that must be crossed for any narrowband operation at microwave frequencies. The local oscillator, a mixer, and an antenna are the minimum requirements for converting low-frequency Amateur transmitters and receivers for microwave operation.

Real communications over significant distances are indeed possible with this minimal station. The basic station doesn't have to be expensive. Receive preamplifiers and transmit amplifiers may be added at very little cost, and the recent availability of low-noise amplifiers in the 4-GHz range makes a good low-noise, moderate-power (by microwave standards) station actually affordable. QSOs over hundreds of miles are possible with such equipment and just a small, reasonably priced dish antenna. For less than the cost of a 2-meter fm transceiver, one may assemble a station capable of providing microwave DX and the excitement of participating in the exploration of still-uncharted regions of the radio spectrum. A block diagram for generating microwave SSB is shown in **fig. 1**.

A secondary goal of this article is to demonstrate a way to put a high-performance station on the microwave bands with as little test equipment as possible. Wherever possible, rf technology replaces complex microwave hardware. I've tried to use commonly available parts wherever possible, and kept the number of pc boards to a minimum by the use of common circuits. Only two microwave circuits, the mixer and PLL downconverter, have to be built, since the oscillator — a Gunnplexer™, for example — can be obtained commercially or as surplus (look for motion detectors and automatic door openers). Except for these, the entire station is assembled from standard VHF or lower frequency components and circuits.

I'd encourage anyone interested in getting on the microwave bands to seek out other interested Amateurs. Small users groups can be a great help in getting started and maintaining enthusiasm. They're also a good way to share not only measurement equipment, but expertise.

local oscillator

As shown in **fig. 1**, the basic blocks for a microwave station are the signal mixer, a bandpass filter to

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Vol. MCMLXXXVIII | Fact Sheet | February 1988

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Table 1. Combinations of a few basic reference signals allow great flexibility in generating phaselocked microwave signals. Combinations using a 1010 MHz reference allow access to every amateur microwave band, using either a 280-290 MHz or a 420-440 MHz SSB i-f.

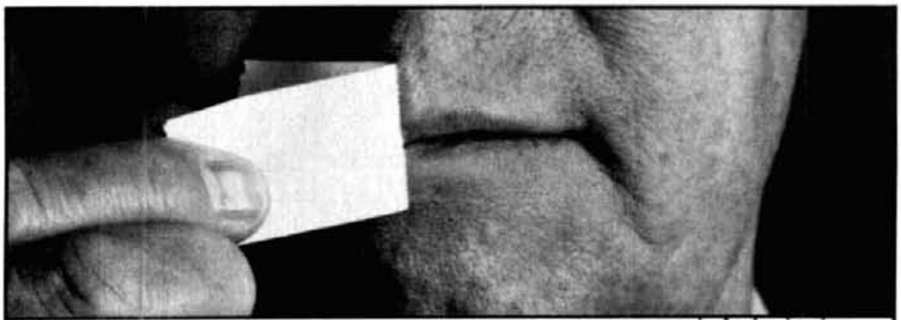
Reference signals

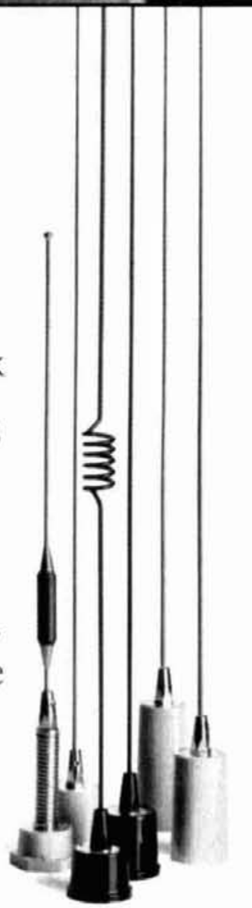
ref osc frq = (harm x ref) + PLL i-f

10_{std}	quartz standard in proportional oven
100_{ref}	100_{ref} = 10 x 10 std + 0 (decade divider instead of i-f)
150_{ref}	3 x (100_{ref} / 2)
20_{ref}	100_{ref} / 5
30_{ref}	3 x 10_{std}
10_{ref}	20_{ref} / 2
40_{ref}	20_{ref} x 2
8_{ref}	40_{ref} / 5
330_{ref}	(3 x 100_{ref}) + 30_{ref}
990_{ref}	(10 x 100_{ref}) - 10_{ref} or 3 x 330_{ref}
992_{ref}	(10 x 100_{ref}) - 8_{ref}
1008_{ref}	(10 x 100_{ref}) + 8_{ref}
1010_{ref}	(10 x 100_{ref}) + 10_{ref}

Microwave Local Oscillators

Band	1st LO = (harm x ref) + PLL i f	SSB i-f	hf SSB dial (260 MHz 2nd LO)
1296	1008 = 1008_{ref} directly	288 MHz	28 MHz
	1010 = 1010_{ref} directly	286 MHz	26 MHz
2304 H*	2016 = 2 x 1008_{ref}	288 MHz	28 MHz
2304 H	2020 = 2 x 1010_{ref}	284 MHz	24 MHz
L	2020 = 20 x 100_{ref} + 20_{ref}	284 MHz	24 MHz
3456 H	3030 = 3 x 1010_{ref}	426 MHz	
L	3020 = 30 x 100_{ref} + 20_{ref}	436 MHz	436 MHz SSB
L	3020 3 x 1010_{ref} - 10_{ref}	436 MHz	436 MHz SSB
5760 H	6048 = 6 x 1008_{ref}	288 MHz	28 MHz
5760 H	6060 = 6 x 1010_{ref}	300 MHz	
L	6020 = 60 x 100_{ref} + 20_{ref}	280 MHz	20 MHz
L	6020 = 6 x 1010_{ref} - 40_{ref}	280 MHz	20 MHz
10368 H	10080 = 10 x 1008_{ref}	288 MHz	28 MHz
10368 L	10080 = 10 x 1010_{ref}[6] - 20_{ref}	288 MHz	28 MHz
24192 H	23760 = 24 x 990_{ref}	432 MHz	432 MHz SSB
24192 L	23760 = 24 x 992_{ref} - 48	432 MHz	432 MHz SSB
24192 L	23910 = 24 x 990_{ref} - 150_{ref}	282 MHz	22 MHz
L	23910 = 24 x 1010_{ref} - 330_{ref}	282 MHz	22 MHz





remove unwanted mixing products, and the microwave local oscillator. Transmit and receive amplifiers are optional. Of these blocks, the local oscillator is probably the most complex. The problem of generating a stable local oscillator in the microwave bands may be solved in more than one way. Certainly an obvious method is to adopt the same techniques used at lower frequencies and traditional in Amateur VHF and UHF construction. That is, one may simply start with a high quality quartz oscillator — perhaps in the 100-MHz vicinity, since crystal oscillators there can have good spectral purity — and multiply up to the final microwave frequency selected. One problem with this is the necessity of including an active device and filter in each harmonic stage. If too high a harmonic number is used, it may be difficult to separate the preferred harmonic from the undesired ones around it. Even if suitably high *Q* filters are available, the extra complexity and amplifier power necessary to generate sufficient energy to operate a mixer is undesirable. This approach also isn't very versatile, since modifying oscillator fundamental frequencies, not to mention a whole string of harmonic filters, makes significant QSY extremely difficult. In addition, signal levels and multiplier stage gains need to be controlled to avoid multiplying broadband noise and degrading the ultimate microwave signals.

One alternative to frequency multiplication is to directly phase lock a microwave oscillator with a phase-locked loop (PLL). This is particularly interesting because when pushing any active device to its upper frequency limit, an oscillator may be built even when frequency multipliers and amplifiers aren't possible. A phase-locked oscillator may be used to produce a signal at frequencies where few active devices are available. Since suitable oscillators are readily available both commercially and on surplus markets, phase lock is an attractive possibility for Amateur operation that requires a minimum of microwave equipment construction.

Normally, to phase lock an oscillator, a signal that's a sample of the oscillator is compared to a high-stability reference signal. A loop amplifier is then used to steer, or "lock" the oscillator in step with the reference signal. A block diagram for phase locking an oscillator is shown in **fig. 2A**. The "phase sample" may be the oscillator signal itself, a frequency-divided or a frequency-converted version of it, *as long as phase information is retained*. Using the oscillator signal itself would require that the reference signal already be the desired stable signal. Using a divided signal would require dividers that operate at the oscillator frequency, but microwave frequency dividers aren't yet Amateur junk box items!

In its simplest form, frequency conversion presents the same disadvantage as using the oscillator signal directly — a precise microwave signal must already be available. However, it's possible to use a harmonic mixer to downconvert the oscillator signal (see **fig. 2B**). Such a mixer mixes the oscillator signal with a harmonic of a much lower frequency reference. This lower frequency reference signal may itself be produced by PLL techniques.

If phase lock is achieved by harmonic downconversion, the phase comparison may be accomplished at low i-f frequencies where gain is easy to come by and measurements are far easier to perform. Additionally, the harmonic converter need not have particularly low conversion loss, since only a reasonable signal-to-noise ratio is sufficient, the absolute signal level can easily be modified with i-f amplifiers. This is in contrast to harmonic multiplication, which requires enough drive at each stage to drive the active device or mixer into its nonlinear region. Step-recovery diode multipliers often need several hundred milliwatts of drive to function properly. Harmonic downconverters may need only 10 to 50 milliwatts of lower frequency drive, a much easier proposition. Additionally, phase-locked oscillators don't usually require filtering to remove unwanted signals as do multiplied oscillators, as long as the reference frequency components are suitably reduced by the loop filter.

Any available microwave oscillator may be used as long as a means of electronic tuning *that will allow correcting the frequency over a range which is larger than the drift caused by mechanical and environmental factors* can be found. Many Gunn diode oscillators may be tuned by slight adjustments to their power supply voltage. Other oscillators, such as the Gunnplexers made by M/A-Com, have an electronic tuning input. The PLL bandwidth must also be sufficiently high to clean up the oscillators' instabilities acceptably. It must be possible to modulate the microwave oscillator at frequencies a few times higher than this loop bandwidth for good loop stability. The bandwidth is easily adjusted by changing the PLL loop parameters, and usually something in the tens of kHz area is adequate. Because any local oscillator must be stable and accurate, the PLL is automaticlly a suitable choice in this respect. If all oscillators in a system can be referenced to one precision (preferably ovenized) reference oscillator, the best in frequency stability and accuracy may be obtained.

common oscillator circuits

After many years of VHF and UHF construction, there are two kinds of circuits I've gotten tired of building: one is power supplies (especially those with tubes) and the other is local oscillator/multiplier strings. Although using a PLL can simplify local oscillator construction, there are several microwave bands, and it's best to avoid duplication of circuits as much as possible. To simplify multiband stations, many microwave enthusiasts are adopting a scheme using a 288-MHz SSB/CW i-f and local oscillators derived from a 1008-MHz precision source. All of the Amateur microwave bands except 24 GHz may be reached by mixing either 288 MHz or 432 MHz with a harmonic of 1008 MHz and selecting the appropriate sideband.

I've chosen a modification of this approach that uses 1010 MHz and a 280- to 290-MHz VHF intermediate frequency. Allowing slight deviation of the signal i-f from 288 MHz can also allow the MHz digit on the hf signal source to indicate correctly. For example, for operation on 2304 MHz, the hf transceiver is tuned to 24 MHz, giving an i-f of 284 MHz. This signal mixed with two times 1010 MHz gives 2304 MHz, and the MHz digit of the hf rig displays the MHz of the output frequency correctly. Since many of the newer transceivers will operate in transverter mode over 20 to 30 MHz, this approach seems to be suitable.

Table 1 shows some alternatives for generating signals on the microwave calling frequencies using this phase-locking approach. The reference signals are all derivatives of the 10-MHz standard oscillator. Phase lock is obtained by mixing a harmonic of a reference signal with the local oscillator to produce a PLL i-f frequency. This PLL i-f is then locked to a reference signal, producing a phase-locked local oscillator signal which can then be used to convert the 280- to 290-MHz VHF i-f to and from the preferred microwave band. The VHF i-f is generated using a 260-MHz phase-locked LO and the 20- to 30-MHz range of the SSB transceiver.

Also shown are some more traditional multiplying approaches for achieving a microwave LO.

Notice that the 1010-MHz phase-locked approaches all provide correct readout of the MHz digit of the SSB source. This may be of little concern for operation on only one microwave band, but since one LO system can effectively put you on all the microwave bands, frequency confusion in the heat of a contest may be something to reckon with!

The additional flexibility of the PLL i-f also provides access to the 24-GHz band. If this band seems esoteric, I'd like to point out that a 24-GHz oscillator with a built-in mixer diode is available for approximately $50. It may be possible to phase lock and mix for the receive i-f all in one diode, eliminating any other microwave hardware! Similar oscillator/mixer modules, designed for radar "gun" use, are available for 10 GHz.

preparing the 10-GHz station

The approach just described, which can be used either directly or with some modifications to get on all of the Amateur bands — affords a great deal of commonality in phase-lock circuitry. It's possible to lay out one circuit board that can be configured for almost every variation required for any of the locked loops. (This is important to avoid the power supply syndrome mentioned above!) Standard ECL integrated circuits and commonly available operational amplifiers are all that's required except for the harmonic downconverters. Adding another microwave band generally requires only an oscillator, a harmonic downconverter, and an additional phase-lock pc board. Also, significant frequency change within a band is often accomplished by simply selecting a different reference signal and changing the coarse frequency adjustment of the microwave oscillator.

In all cases, from 10 MHz to microwave, the best "raw" oscillator should be used. At 10 MHz, the long-term stability of the standard is the desired characteristic. The loop bandwidth of this PLL is kept low so that the poorer phase noise contributions of the 10-MHz standard won't contaminate the less accurate but cleaner 100-MHz quartz oscillator. At 1010 MHz, the operating Q of the resonator and the resulting oscillator spectral purity is kept as high as possible. This oscillator is really rather stable when unlocked. Measurements over a period of several days and at a relatively constant temperature showed only a few

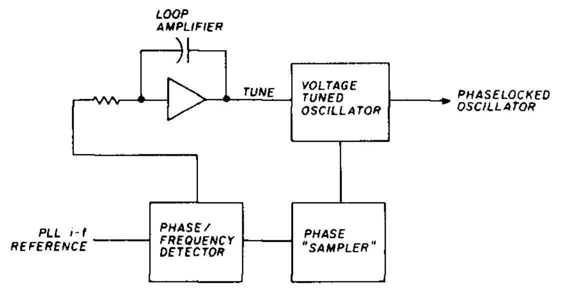


fig. 2A. A simple phaselocked loop (PLL) holds a voltage tuned oscillator (VTO) in constant phase relationship to a reference signal. A signal which is a representative "sample" of the VTO's phase must be provided.

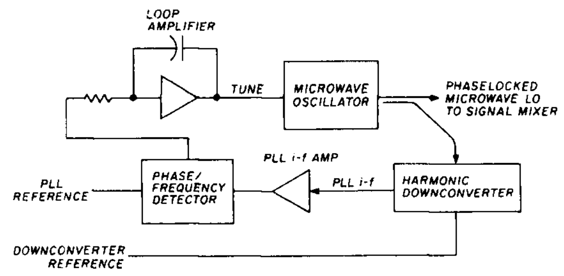
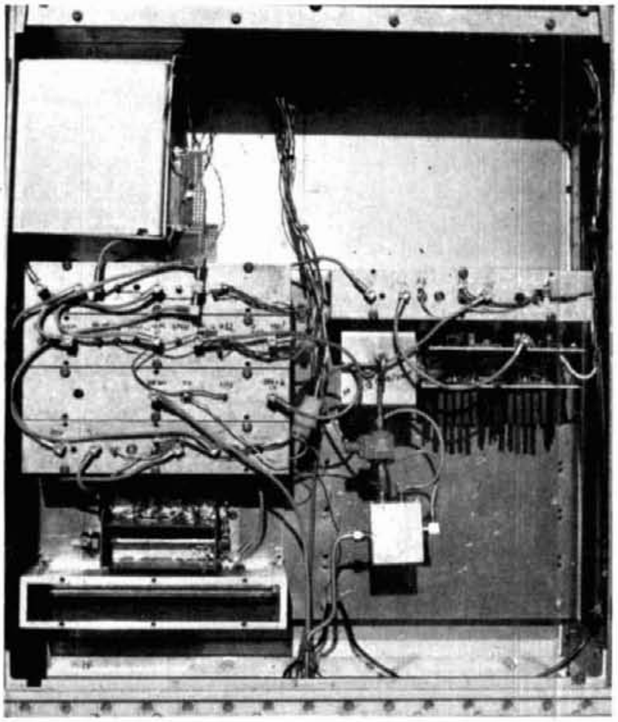


fig. 2B. Phaselocked microwave source block diagram.



Phaselocked 10080 MHz local oscillator and 288 MHz SSB transverter. 10 MHz standard in upper left.

tens of kHz drift. The spectrum of this oscillator when locked is essentially a replica of the tenth harmonic of the 100-MHz crystal oscillator. This feature, combined with simplicity of construction, makes this an effective solution to the problem of generating a clean and stable GHz signal.

A multiplier/filter string could be used to generate this 1010-MHz signal instead of the direct phase-locked approach. This would probably require either an ovenized 101-MHz oscillator or dividing by 101 and phase locking to a 1-MHz reference derived from a master station oscillator. In that case, sufficient filtering at 1010 MHz must be provided to reduce spurious signals to -70 dBc or less. The direct phase-lock approach provides a clean spectrum that's directly attributable to the high-*Q* resonator, as long as loop bandwidth is small enough to keep reference frequency sidebands sufficiently low. The only other spurious signals present are multiples of the 100-MHz signal from the harmonic downconverter. These signals, predominantly the odd harmonics at 900 MHz and 1100 MHz, are attenuated by the isolation between the 1010-MHz buffer amplifiers.

The microwave oscillator may be whatever you can get or build. However, using Gunn diode oscillators (such as the M/A-Com Gunnplexers) results in a 10.080-GHz signal that has as clean a spectrum as many hf transceivers on 20 meters. The resulting 10,368-MHz SSB signal is also of excellent quality.

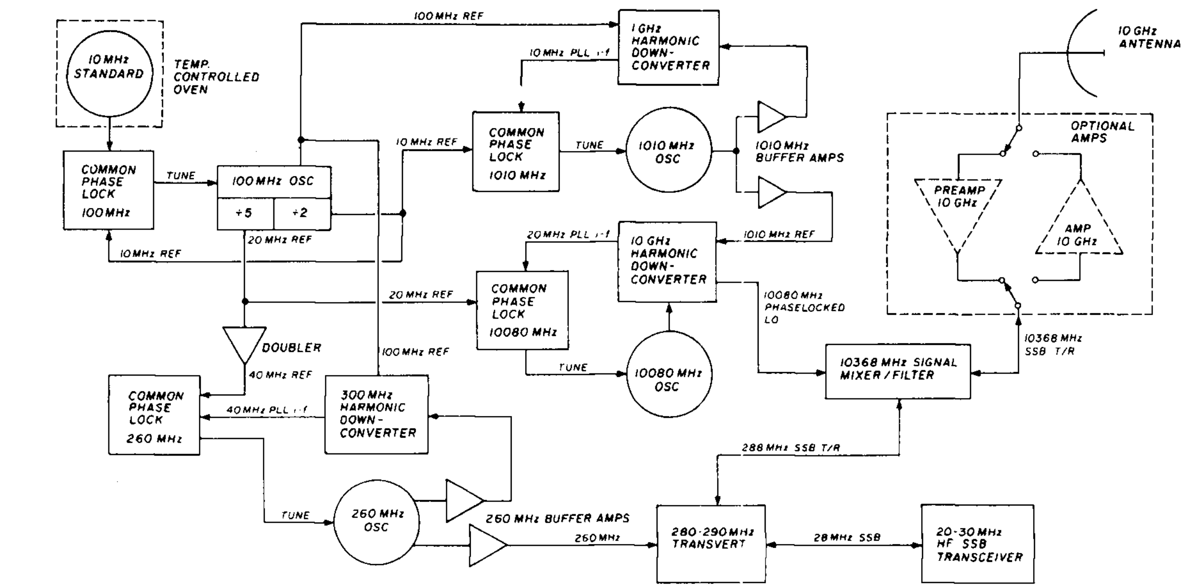


fig. 3. A precise 10,368 MHz SSB station can be made from only 5 oscillators and 4 nearly identical phaselock circuits. The PLL downconverters are similar in topology and easy to construct.

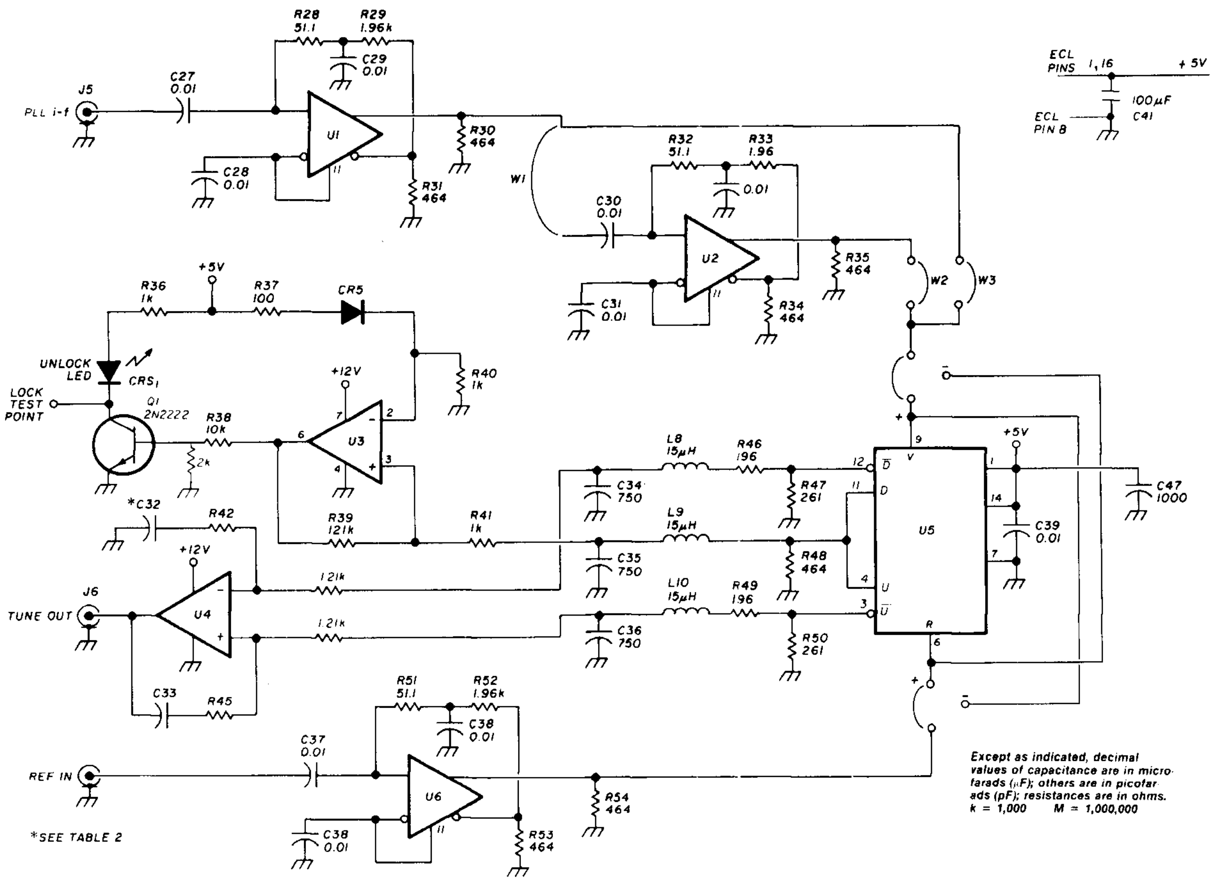


fig. 4. The common phaselock circuit uses inexpensive ECL and op-amp integrated circuits. Only 4 components and 2 jumpers need be changed for a great variety of different applications.

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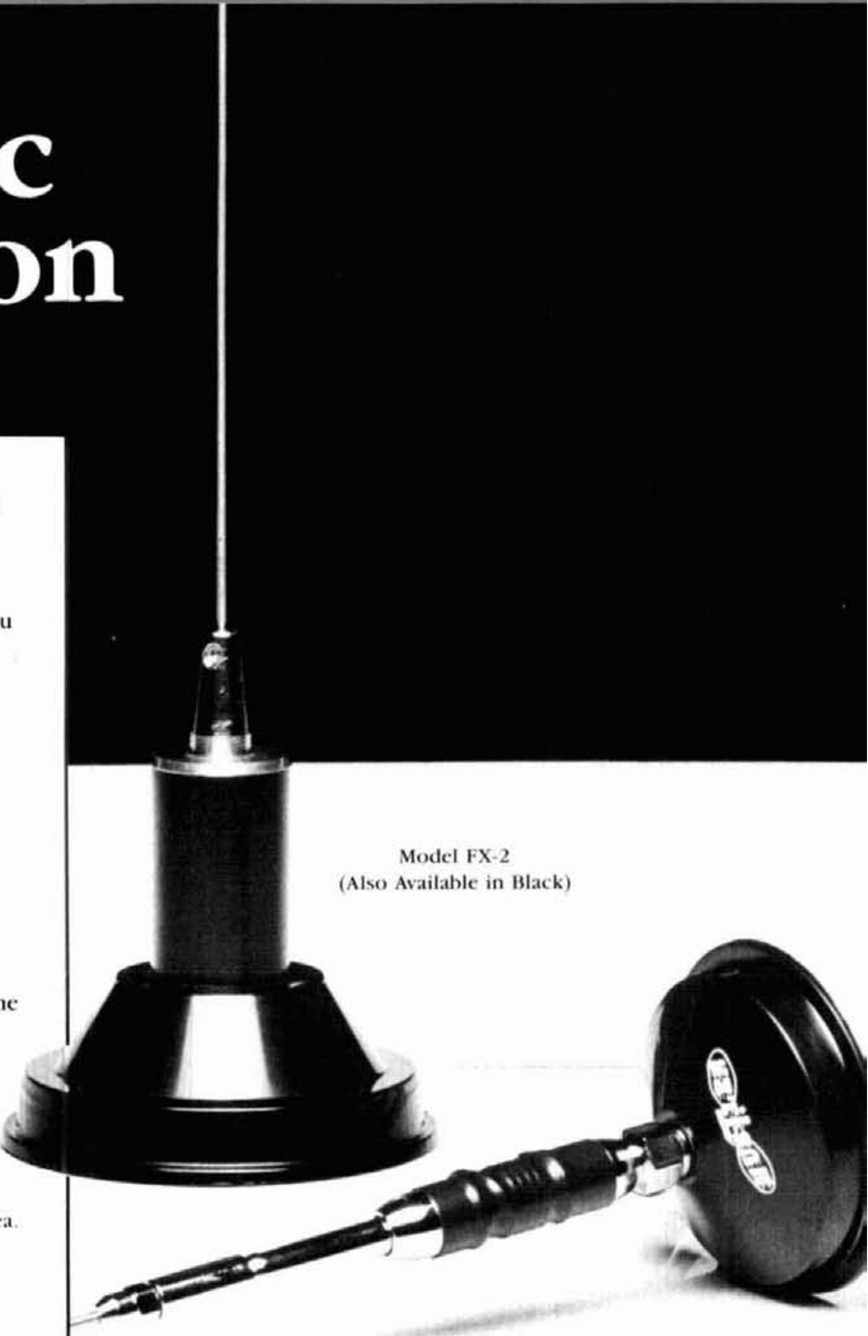

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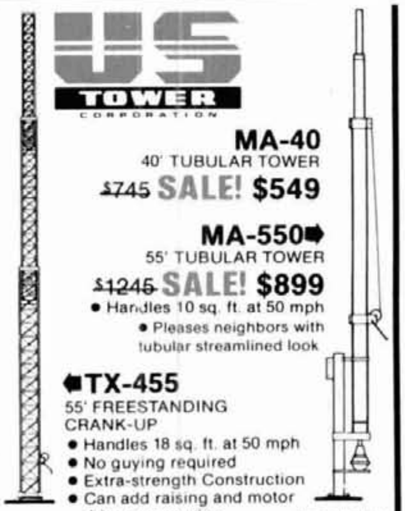

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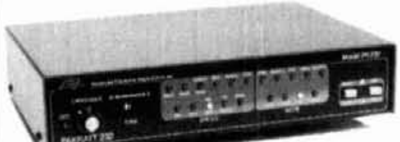

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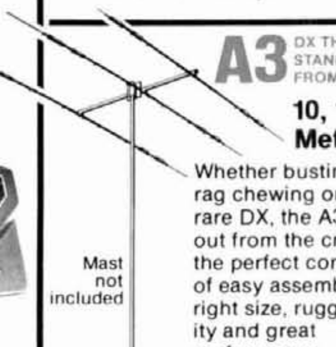

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Gunn diode oscillators used as sources for automatic door openers have been tried with similar results.

Using this LO approach on 10 GHz is quite amazing. Even when out on a hilltop, once the oven for the 10-MHz frequency standard is operating at temperature, the hf transceiver can be set exactly to a scheduled frequency. When the another station calls, the SSB signal from a similarly phase-locked distant station is tuned correctly. A detailed system block diagram is shown in **fig. 3**.

order of construction

For those planning to build an entire 10-GHz station, construction of the 100-MHz oscillator and ECL dividers is a good place to start. If a 10-MHz standard is available, the phase-lock circuit may be built and the 100-MHz oscillator phase locked. If possible, a common pc board should be made and several boards loaded and tested. Then, building the 1010-MHz oscillator, buffer amplifers, and downconverter, and changing four components on one of the pretested phase-lock boards, will result in a precision 1010-MHz reference signal. Next, building the 260-MHz oscillator and locking it with another of the phase-lock boards and adding mixers and amplifiers produces the 280- to 290-MHz i-f transverter. At this point, both 1296 and 2304 are within easy reach by adding just a mixer (and amplifiers, if you wish). For 2304, an anti-parallel diode mixer just like the harmonic downconverters will suffice to get a signal on the band without even building a doubler or additional locked 2020-MHz oscillator. Inexpensive MMIC amplifier blocks can be used for transmit and receive amplifiers to produce a respectable 1296- or 2304-MHz station almost immediately. Finally, the 10,080-MHz LO is obtained by locking a Gunn oscillator with one more common pc board and the 10-GHz harmonic downconverter. Here again, one can immediately get on the band at the few hundred microwatt level by just adding the signal mixer. This mixer can even be built on the same Teflon™ pc board and at the same time as the harmonic downconverter. A simple bandpass filter for 10,368 MHz has already been discussed, and a two-stage amplifier will be described at the end of this article.

For those who want to get on the 10-GHz band as rapidly as possible, the flexibility of this phase-locked approach offers many choices. It is possible, for example, to get on 10,368 MHz SSB using only a 148-MHz SSB signal as the i-f. This can be done by generating a 1020-MHz signal instead of 1010 MHz for the microwave downconverter reference, simply by using a 20-MHz reference in that PLL and coarse tuning the resonator 10 MHz higher. Also, using a 20-MHz reference on the 10-GHz Gunn oscillator PLL gives 10 x 1020 MHz + 20 MHz = 10,220 MHz. 10,220 MHz LO + 148 MHz SSB = 10,368 MHz SSB. The 10,220-MHz LO could also be used with a 30-MHz wideband i-f to work other Amateurs in the area not yet on narrowband modes using 10,220/10,250 full duplex. SSB operation could be utilized by just changing to the 148-MHz SSB i-f. Notice also that reversing the reference and VCO inputs to the phase comparator allows locking on "the other side." In this last example, 10,180 MHz would result from such a reversal. Many combinations of reference frequencies and tuning directions are possible, and this versatility is an attractive aspect of this whole approach.

common phase-lock circuit

By taking a uniform approach to local oscillator generation, that of phase locking to a relatively low frequency reference, a great deal of commonality in circuit design is achieved. In fact, all four phase-lock loops used to generate 10,368-MHz SSB have identical phase detector, loop amplifier, and lock indicator schematics. Only the loop filter component values need to be varied to accommodate the differing oscillator characteristics. This is a great boon to construction, since one circuit board may be laid out which serves all four loops. The phase-lock circuit board uses ECL line receivers to amplify the reference and VCO inputs, phase detection is in a 12040-ECL phase comparator, and standard operational amplifiers are used in the loop amplifier/filter and phase-lock detect circuits. A schematic of the common phase-lock circuit is shown in **fig. 4**.

I made a common pc board that contained the phase-lock circuits, the 100-MHz oscillator and dividers, and the 1-GHz harmonic downconverter/PLL i-f amplifier. All this fits on a double-sided 3- x 6-inch board, and only the parts required for a particular function need be added. Normal VHF construction practices are followed, including good bypassing on the ECL logic, which is important since operation is from +5 instead of −5 volt supplies. The only circuit that operates higher than 100 MHz is the anti-parallel diode mixer, and the two diodes can be mounted right next to a coax connector to keep lead length to near zero.

I chose to use a large number of small coaxial connectors between circuits. This is more expensive, but it adds a great deal of versatility in changing reference frequencies, measuring signal levels, and so forth. For VFO control, any of the PLL reference frequencies may be substituted with a variable frequency reference as long as ECL logic levels are provided. This may be useful in testing or experimenting. Substituting the 10-MHz reference on the 1010-MHz PLL oscillator with a 10- to 11-MHz variable oscillator, for example, could give a variable 10,080- to 10,090-MHz microwave signal with the same stability as the tenth harmonic of the variable oscillator. This might be useful as a signal source for microwave testing.

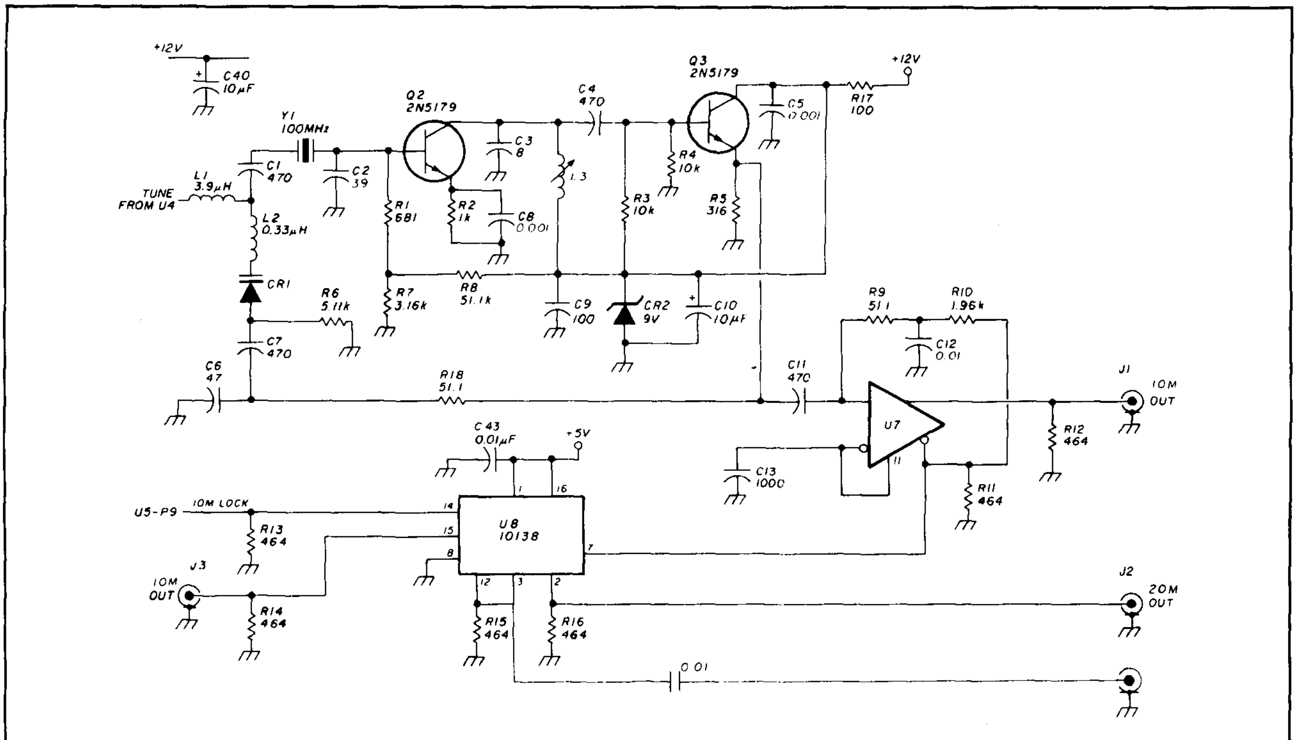

fig. 5. The 100 MHz crystal oscillator and divider circuits provide reference signals for generating all higher frequency signals. The oscillator itself can be locked to a precision 10 MHz frequency standard.

The PLL characteristics are modified by changing R42/R45 and C32/C33. This is necessary to provide different loop bandwidths and accommodate different oscillator tuning sensitivities. **Table 2** gives some values for some selected bandwidths and oscillators. For situations not listed, approximate values may be calculated. Referring to **fig. 6**:

$$A(s) = (K_d\ K_a\ K_v)/(N)$$

Where $A(s)$ = loop gain at frequency s

K_d = phase detector sensitivity in volts/radian

K_a = loop integrator gain at s

K_v = oscillator sensitivity in radians/(volt sec)

N = frequency division of oscillator before phase detector

Component values can be calculated by first selecting a loop bandwidth and setting K_a to give unity gain at that frequency:

$$K_a = (N\ \omega_n)/(K_d\ K_v)$$

Where ω_n = loop bandwidth in radians/sec

(radians/sec = $2\pi f$ with f in Hz)

R_2 can be selected by R_2:
$= K_a\ R_1$

Next C can be selected by choosing $1/(R_2\ C)$ to be four or five times lower in frequency than the loop bandwidth to provide adequate loop stability.

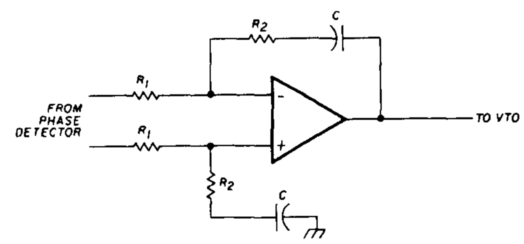

fig. 6. The loop amplifier (integrator) controls the characteristics of the PLL. Loop parameters are set by properly selecting component values. In this balanced configuration, each component appears twice.

$$C = 5/(\omega_n R_2)$$

As an example, let's assume we want to lock an oscillator with a tuning sensitivity:

$$K_v = 0.5\ \text{MHz/volt} = 3.1 \times 10^6\ \text{rad/(volt sec)}.$$

The 12040 phase detector has a

$$K_d = 0.13\ \text{volt/radian}.$$

Since we have no frequency division (a harmonic mixer is a *mixer*, not a divider), N = 1.

Suppose we want a 50-kHz loop bandwidth to "clean up" the oscillator. Then:

$$K_a = (1 \times 2 \times \pi \times 50{,}000)/(0.13 \times 3.1 \times 10^6) = 0.77$$

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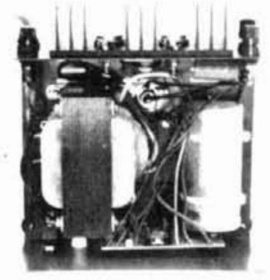

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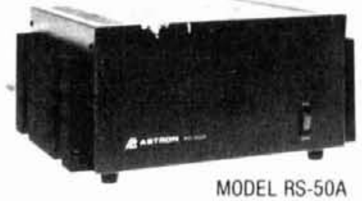
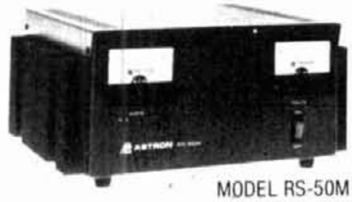


MODEL RS-50A



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RS-7A	5	7	3¾ × 6½ × 9	9
RS-7B	5	7	4 × 7½ × 10¾	10
RS-10A	7.5	10	4 × 7½ × 10¾	11
RS-12A	9	12	4½ × 8 × 9	13
RS-12B	9	12	4 × 7½ × 10¾	13
RS-20A	16	20	5 × 9 × 10½	18
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VS-20M	16	9	4	20	5 × 9 × 10½	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13¾ × 11	46
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RS-10S	7.5	10	4 × 7½ × 10¾	12
RS-12S	9	12	4½ × 8 × 9	13
RS-20S	16	20	5 × 9 × 10½	18

*ICS—Intermittent Communication Service (50% Duty Cycle 5min. on 5 min. off)

Since R_1 is fixed at 1210 + 196 = 1400 ohms.

R_2 = 0.77 x 1400 ohms or 1000 ohms

and

$C = 5/(2 \times \pi \times 50{,}000 \times 1000) = 0.015\ \mu F$

It is important for loop stability that both the oscillator and the op amp operate well below their roll-off frequency. For the oscillator, its tuning sensitivity should be constant to at least five times the loop bandwidth selected. In the case of the op amp, the product $K_a\omega_n$ should be no more than 10 to 20 percent of the specified gain-bandwidth. This specification will depend on the particular op amp used, but if ω_n is always selected to keep $K_a\omega_n$ below 1M radians/second, performance should be satisfactory. When in doubt about loop stability, design for a lower loop bandwidth; noise sidebands on the oscillator may be higher, but at least the loop will be stable!

100-MHz oscillator and divider circuits

A clean 100-MHz quartz oscillator is locked to a stable, though less spectrally pure, 10-MHz standard oscillator. In the event that a stable 10-MHz standard is not available, this oscillator may be operated unlocked, but the frequency accuracy and drift may not be as good as you would like for SSB and CW weak-signal operation, particularly on the higher bands. In any case, this 100-MHz reference may be built first and used for locking the higher frequency oscillators even without a 10-MHz standard.

Table 2. Typical loop amplifier component values for phaselocking the 10 GHz station oscillators. Other oscillator sensitivities can be accommodated by calculating values as demonstrated in the text.

Oscillator	100 MHz VCXO	1010 MHz	10,080 MHz Gunn Osc	260 MHz 2nd LO
Loop bandwidth				
f (Hz)	50	50k	50k	50k
Wn (rad/sec)	310	310k	310k	310k
VTO sensitivity, Kv				
MHz/volt	0.0002	0.5	3	1
rad/(volt sec)	1300	3.1M	19M	6.3M
Loop Gain, Ka	18	0.77	0.13	0.38
divide # N	10	1	1	1
R42/45 (ohms)	25	1.1	180	530
C32/33 (μF)	0.6	0.015	0.09	0.03

These values assume R43/44 = 1210 ohms and R46/49 = 196 ohms and
phase detector sensitivity, Kd = 0.13 volts/radian

A two-stage Pierce harmonic circuit is used with a UHF TV tuner varactor diode to allow frequency control. A variable capacitor may be substituted when no 10-MHz standard is used. An ECL line receiver is used to level shift and buffer this oscillator and drive the bi-quinary ECL divider. The divider output drives one input on the common phase-lock circuit (if used), the other input being provided by the 10-MHz standard oscillator. The loop parameters are set to give 10 to 100 Hz of loop bandwidth. The divide-by-five output of the 10138 divider provides a 20-MHz reference signal for use in the other loops. The 10-MHz ECL signal is also brought out. The schematic of this circuit is shown in **fig. 5**.

Beyond following the usual sound VHF practices, no special precautions need be taken in constructing this circuit. Whether or not 10-MHz standard is being used, it's a good idea to position components so that the entire oscillator circuit can be shielded and, if possible, thermally insulated to avoid frequency drift from ambient temperature change.

testing

If an oscilloscope, spectrum analyzer, or 100-MHz frequency counter isn't available, both the oscillator and divider circuits should be built at the same time. Then a 10-MHz WWV receiver or any receiver that tunes 10 MHz may be used to ascertain oscillation and division. Before trying to "close the loop" and lock the 100-MHz crystal oscillator, it's a good idea to check that the tuning range is from approximately 1 kHz below 100 MHz to 1 kHz above. This can be done by applying both 5-volt and 12-volt supplies and listening to the divider output at 10 MHz to see that the signal swings 100 Hz on either side of 10 MHz when a 2- to 10-volt tuning voltage is applied. Next, hook up the loop amplifier and the 10-MHz reference signal (from any source that is within this tuning range) and monitor the tuning voltage. If the loop locks correctly, the unlocked LED should extinquish and the tuning voltage should move between ground and the positive 12 volts as L3 is adjusted. At each end of that range, the unlocked light should light, indicating unlock due to an out-of-range 100-MHz oscillator. If a problem exists, make sure that the phase comparator inputs are correct, the connections marked "+" are being used, and recheck for wiring errors.

The second installment will cover construction of the 1010-MHz oscillator, harmonic downconverters for both 1 and 10 GHz and some circuits for biasing and driving the 10-GHz Gunn oscillators.

reference

1. Glenn Elmore, N6GN, "A Simple and Effective Filter for the 10-GHz Band," *QEX*, Volume 65, July 1987, page 3.

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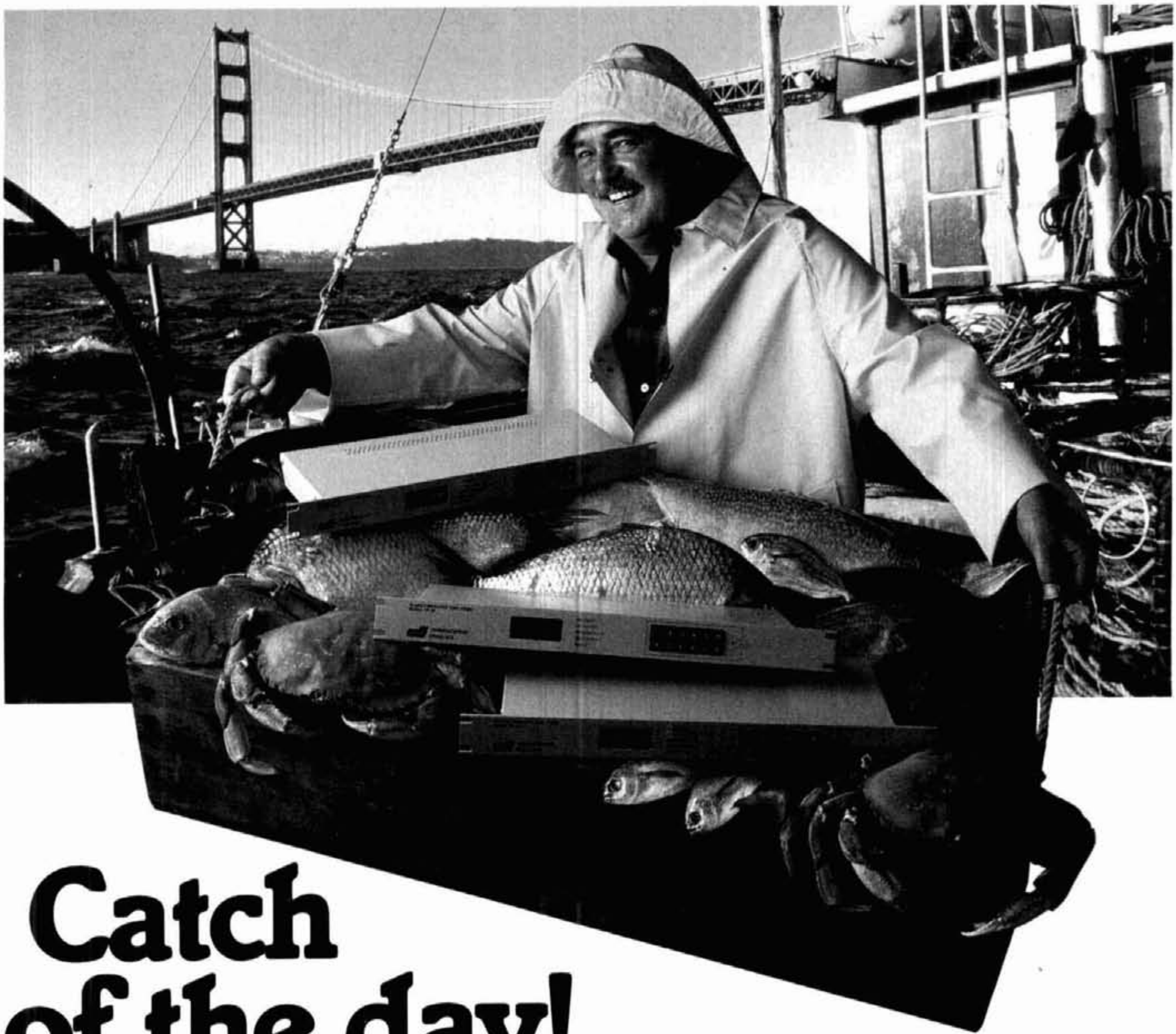

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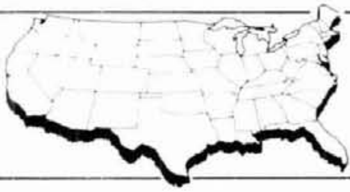

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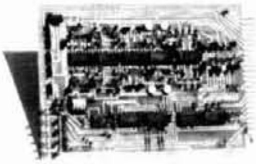

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simple program solves for dimensions

Computer-Aided Design (CAD) techniques have been used to design these stubs for Amateurs.

Equation 1 is an algebraic expression yielding a linear approximation of the CAD segmented microstrip model, solved for the outer radius, R_L for either 60- or 90-degree stub angles.[1]

$$\log R_L = A \log (\sqrt{E_R} \cdot f) + B \log H + C \log R_S + D \quad \textbf{(1)}$$

Note: R_L = *RL* (in listing)
E_R = *ER* (in listing)
f = F (in listing) R_s = RS (in listing where R_L is the long radius of the stub in meters; A, B, C, and D are constants, which are functions of the stub angle; E_R is the dielectric constant; *f* is the frequency in GHz; H is the dielectric thickness in meters; and R_S is the short radius in meters.

The BASIC program, RAD-STUB, (**fig. 2**) gives the value of R_L, the outer radius. The required parameters are the frequency, dielectric constant of the substrate and its thickness, and an estimate of the inner radius, R_S. The result approximates the resonant frequency within 1 to 2 percent for 60-degree stubs, which are the most frequently used.

The BASIC program is straightforward. While it was written for the IBM PC, it should run with other BASIC languages if modifications for screen, input, and exponents are made. **Line 9** clears the screen and removes the function keys from the PC screen. **Lines 30-60** may have to be split into print and input state-

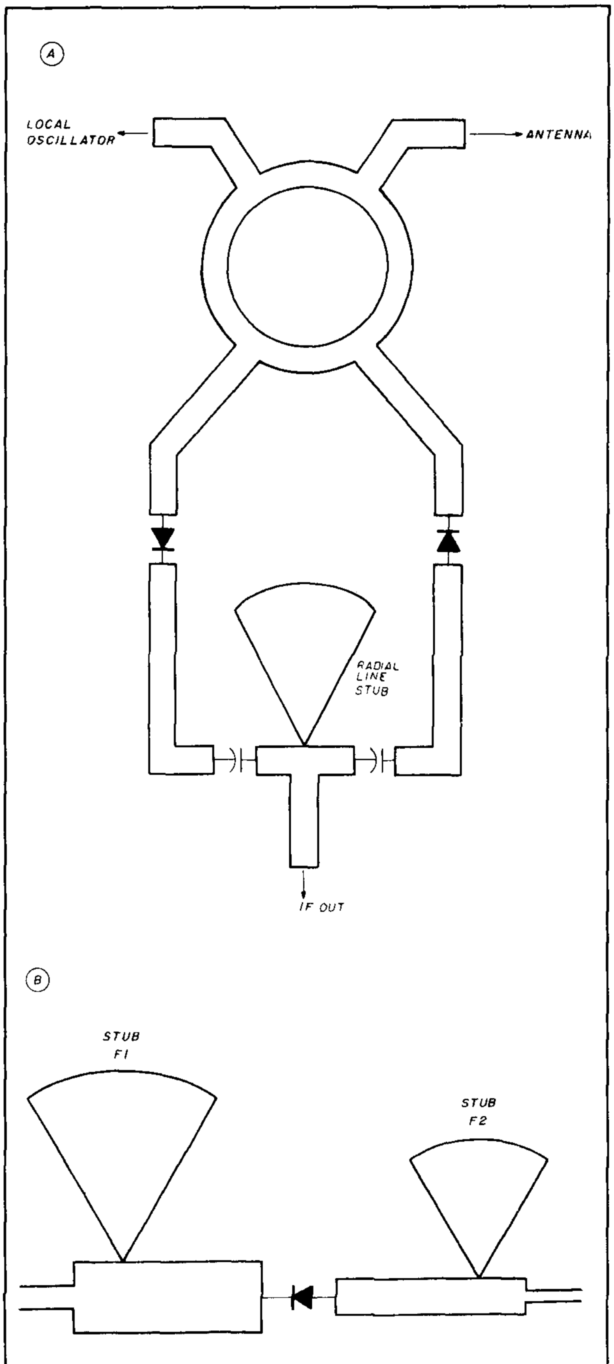

fig. 1. Examples of microstrip design using radial line stubs to provide rf potential grounds: (A) mixer; (B) doubler.

By George W. Allen, N1BEP, 731 Coral Drive, Cape Coral, Florida 33904

```
9 CLS: KEY OFF
10 PRINT "          *************** RADIAL STUB DESIGN ******************"
20 REM                          by George W. Allen, N1BEP
25 PRINT: PRINT
30 INPUT "Frequency, gHz (0.3 - 30)";F: '              *** F=Frequency, gHz
40 INPUT "Dielectric constant";ER: '                   *** ER=Dielectric constant
50 INPUT "Substrate thickness, inches";H: '            *** H=Substrate thichness
60 INPUT "Inner radius, inches";RS: '                  *** RS=Inner radius
70 H=H*.0254: RS=RS*.0254
80 PRINT "Stub angle 1-60, or 2-90 degrees";
85 INPUT K: '                                          *** K = STUB ANGLE
90 ON K GOTO 100,120
100 A=-.8232: B=.0572: C=.1169: D=-.8082
110 GOTO 130
120 A=-.851: B=.0614: C=.0877: D=-.8695
130 L=LOG(10)
140 E=A*LOG((ER^.5)*F)/L + B*LOG(H)/L + C*LOG(RS)/L + D
150 '                                                  *** E=LOG(RL)
160 RL=10^E: RL=RL*39.370079#: '                       *** Meters to inches
170 PRINT: PRINT "Long radius is";
180 PRINT USING "###.####";RL;
190 PRINT "  inches"
200 END
```

fig. 2. Basic program for radial line stub design.

(A) Frequency, GHz (0.3-30)? 10
Dielectric constant? 9.8
Substrate thickness, inches? 0.25
Inner radius, inches? .002
Stub angle 1-60, or 2-90 degrees? 1
Long radius is 0.0743 inches

(B) Frequency, GHz (0.3-30)? 1.296
Dielectric constant? 2.3
Substrate thickness, inches? .06
Inner radius, inches? .002
Stub angle 1-60, or 2-90 degrees? 1
Long radius is 0.7625 inches

fig. 3. Program results for (A) 10 GHz and (B) 1.296 GHz.

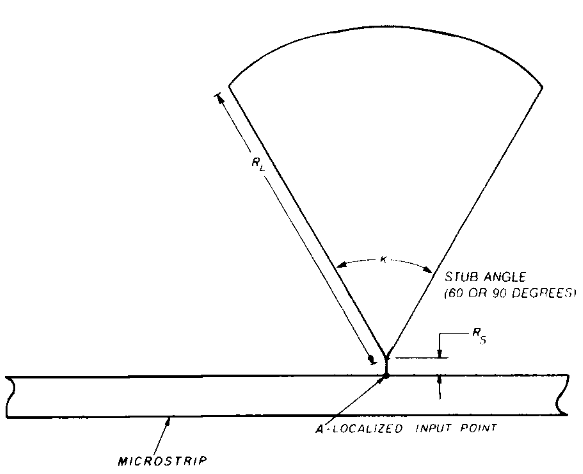

fig. 4. Mixer stub geometry, top view.

ments. **Lines 140** and **160** use the caret for exponentials. **Line 180**, which truncates the printout to limit useless digits, may be just a PRINT command. The constants for 60 and 90 degrees, **lines 100** and **120** are based on metric units, although common usage is in inches for printed circuit material; **lines 70** and **160** accommodate the conversion. The equation is in logarithms to the base 10, so **lines 130, 140** and **160** convert to natural logarithms and back to solve the equation.

examples

Two examples of the program's output are shown in **fig. 3**. The limits of the approximation are from 0.3 to 30 GHz, but the result becomes quite broad at the high end. The dielectric constant values should lie between 2 and 15 for the values of the constants used in the approximation. The results are consistent with the values of dielectric constant in practical materials at given frequencies, and with the ability to measure and etch printed circuits at these small dimensions. **Figure 4** illustrates a top view of the stub geometry.

summary

A BASIC program solves the approximate equation for the dimensions of the geometry of a radial line stub for the frequencies between 0.3 and 30 GHz for values of dielectric constant from 2 to 15.

reference

1. H.A. Atwater, "The Design of the Radial Line Stub: A Useful Microstrip Circuit Element," *Microwave Journal*, Volume 28, No. 11, November 1985.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

"the greatest little radio magazine in the world"

VHF? Yes. Single sideband? Yes. Beam Antennas? Yes. Microwaves? Yes. Propagation? Yes.

The year was 1930. America was in the grips of the Great Depression. The unemployment rate, during that harsh winter, was over 30 percent. Wall Street was in a panic. The future looked bleak, indeed.

From the perspective of today, the world we knew in 1930 is hard to imagine. No one had heard of Ronald Reagan or Olivia Newton John. No homes contained TV sets, VCRs, air conditioners, or electric blankets. No one had heard of atomic meltdown, jet aircraft, coffee machines, water beds, hand calculators, word processors, or microwave ovens.

Young men, with little money but plenty of time on their hands, slowly grew interested in shortwave radio. For a dollar or so, an old battery radio could be bought, stripped down, rebuilt into a two-tube shortwave receiver. A second-hand car battery would provide filament power, and B-batteries were cheap.

Once the set was completed, the builder had the world at his fingertips. But it didn't come easy. Where was the information he needed to build a two-dollar shortwave receiver? And when it was built, what could you hear? What was going on in the world of shortwave radio?

meet Hugo Gernsback

Into the growing hobby stepped a giant with an imposing track record of accomplishing feats that others

fig. 1. Early issues of "Short Wave Craft" magazine were full of exciting, new ideas. Feature articles on short wave medical applications (diathermy), "moonbounce", and inexpensive phone transmitters guaranteed high interest among would-be hams for each new issue. Articles by Dr. Alfred N. Goldsmith and Dr. Willis Whitney satisfied the more technical readers.

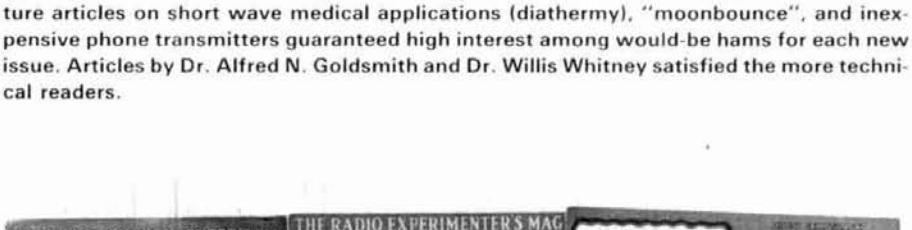

fig. 2. As "Short Wave Craft" matured, it appealed more to the amateur market. Covers by "Paul" (above), who did the covers for "Amazing Stories" were a feature of some issues. Interest in shortwave listening seemed to be dropping off, so the magazine catered to radio hams until World War II.

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fig. 3. The "Short Wave Craft" magazine changed it name to "Radio and Television" in 1940, after a brief existence as "Short Wave and Television." But by 1941 the magazine was fragmented, vainly trying to appeal to various radio hobbyists. After the war, the magazine combined with other Gernsback publications and eventually became "Radio Electronics."

thought impossible. He had established the world's first radio store and sold the world's first radio magazine. His name, Hugo Gernsback, was well known as that of the publisher of science fiction magazines. Now he was going to turn his hand to publishing the greatest magazine in the world for shortwave enthusiasts!

Short Wave Craft magazine

The first issue of *Short Wave Craft* magazine hit the newstands in 1930 at the exorbitant price of fifty cents. Though the pricing was a blunder, the magazine was an immediate success; the cover price soon dropped to a quarter. Bursting with new ideas and exciting circuits that a high-school lad could understand, the magazine quickly became a focus of interest among the younger generation. It opened the door not only to shortwave radio, but to Amateur Radio as well (**figs. 1, 2, 3**).

While a certain proportion of the articles sailed over the heads of most readers — for example, those about sideband transmitters (see **fig. 4**), inverted speech for transatlantic telephony, and curtain array antennas — each issue had enough down-to-earth contruction material to satisfy beginning hobbyists.

"afternoon radio DX"

One of the greatest attractions of *Short Wave Craft* was the number of "do-it-yourself" articles. Ham radio was going through a population explosion, with ranks increasing from slightly over 20,000 to over 34,000 in only two years! Most of the new Amateurs were high-school boys who had plenty of enthusiasm — all they needed was information. As the magazine pointed out, a second-hand battery-operated broadcast receiver bought for 25 cents or less could provide everything needed to get on the air with a receiver and a phone transmitter!

The receiver was a two-tube job (**fig. 5**). Either type 30 or type 201A tubes could be used. All of the parts came from the defunct battery set. A passable pair of headphones cost less than a dollar, and doorbell batteries could be had for next to nothing.

The transmitter circuit (**fig. 6**), however, was another matter. The big stumbling block was the microphone. Everything else came from the old broadcast receiver. If you knew the right guys, you could get a "mike" from a telephone for fifty cents. (They were probably "liberated" from pay phones, but it was wise not to ask too many questions.)

Once all the parts were assembled, a day's work would produce a breadboard transmitter that would work in the 160-meter band. It ran about 20 watts input, which, with a 120-foot antenna, provided a range of about 50 miles.

This was a lot of fun! The band came alive about 3:30 in the afternoon, when everybody started coming home from high school. From that time, until just before dinner, the band was yours! These wobbly, little phone

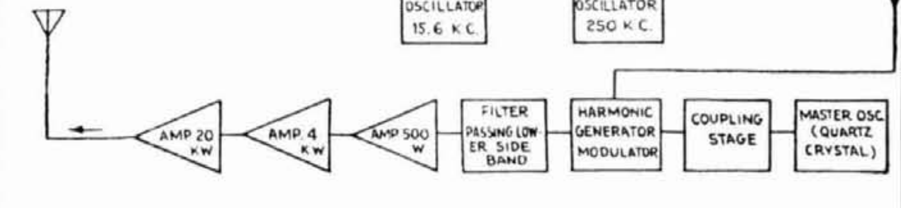
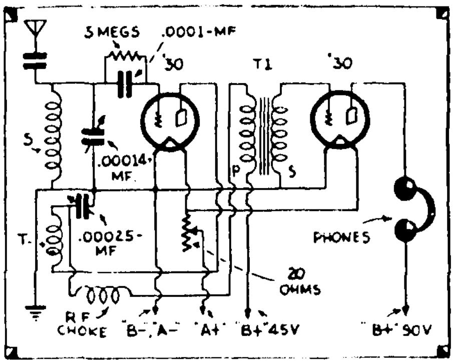


fig. 4. A look at single sideband in 1930, when it was hardly a household word! Because of the problem of building a high-Q sideband filter, the early transmitters started out with a balanced modulator in the 20 kHz to 50 kHz range. This circuit shows sideband generation in the 16-19 kHz range, with additional mixing to 265 kHz.



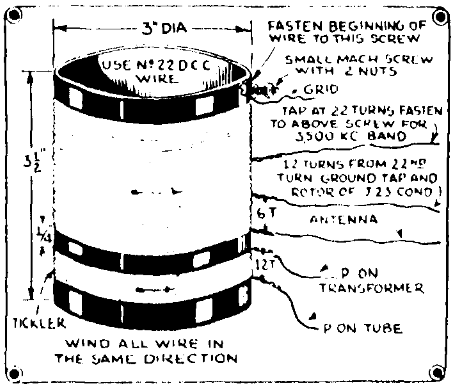


fig. 5. A real shortwave receiver for pennies that actually works. Thousands of would-be hams built this simple set from the pages of "Short Wave Craft" during the "thirties" and were thrilled with world-wide reception.

transmitters allowed groups of enthusiasts to talk all over town and — once in a while — even out of state!

Then, shortly after 5:00, the "big boys" started coming on the air, after getting home from work. How could a little 20-watt modulated oscillator compete with a high-power, 50-watt crystal-controlled rig? It couldn't, and the high-school gang regretfully went QRT until the next day.

As the months went by, the high-school afternoon gang slowly disappeared. Some went to college, others got jobs. Most of them upgraded their equipment and the little oscillators were forgotten.

So was *Short Wave Craft* magazine. The breathless, wide-eyed approach to shortwave radio was out of fashion. Build-it-yourself projects still appeared in the magazine, but they were aimed

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at more technically minded readers. An increasing portion of the magazine was devoted to television. More and more young Amateurs turned to *QST*, which suited their moods better. Even so, the "greatest little radio magazine in the world" lasted until after World War II, when it was reborn as *Radio Electronics*, which endures today as one of the many Gernsback Publications.

TVI revisited

Now that 10 meters is coming back to life, the problem of TVI is becoming more prevalent. Since the last big period of 10-meter activity (around 1978), television has undergone changes. More and more receivers are either on cable systems or, if they're not, employ 75-ohm coax lead-in instead of the popular 300-ohm ribbon line.

TVI (high-pass) filters are available for 75-ohm line, but they don't seem to do the job. I've given some filters to my neighbors, but they report very little improvement when I'm on 10 meters and they're watching channel 2. I crosshatch the screen with lines regardless of whether the filter is installed in the line or not.

The problem seems to be that the 75-ohm line acts as an antenna for my signals, and the TV antenna itself has little to do with it. If the coax line can be broken at the set, from a TVI point of view, the interference should disappear.

My friend "Bip", W6BIP, who's an expert when it comes to TVI problems and solutions, tipped me off on how to cure this worrisome interference. His solution, which works fine for me, is to use two 4:1 TV balun transformers back-to-back with a 300-ohm TVI filter connected between them (**fig. 7**). The balanced filter breaks both line conductors and the baluns prevent line unbalance caused by inserting the filter.

Bip cautioned me that the little TV baluns (which are nothing more than a few fine windings on a ferrite bead) have some loss, and if this trick is tried in an area of low TV signal strengths, it might put too much "snow" on the picture. The solution, then, is to install a broadband amplifier after the filter and before the TV set.

Bip also pointed out that it's good insurance to wrap the power cord to the TV around a high-permeability ferrite core just to prevent the possibility of your signal sneaking in through "the back door" of the TV set.

I'd be interested in hearing from readers who have solved tough interference problems. Your solution may help others who are in trouble. Write to me at Box 7508, Menlo Park, California 94025, with your ideas.

ham radio

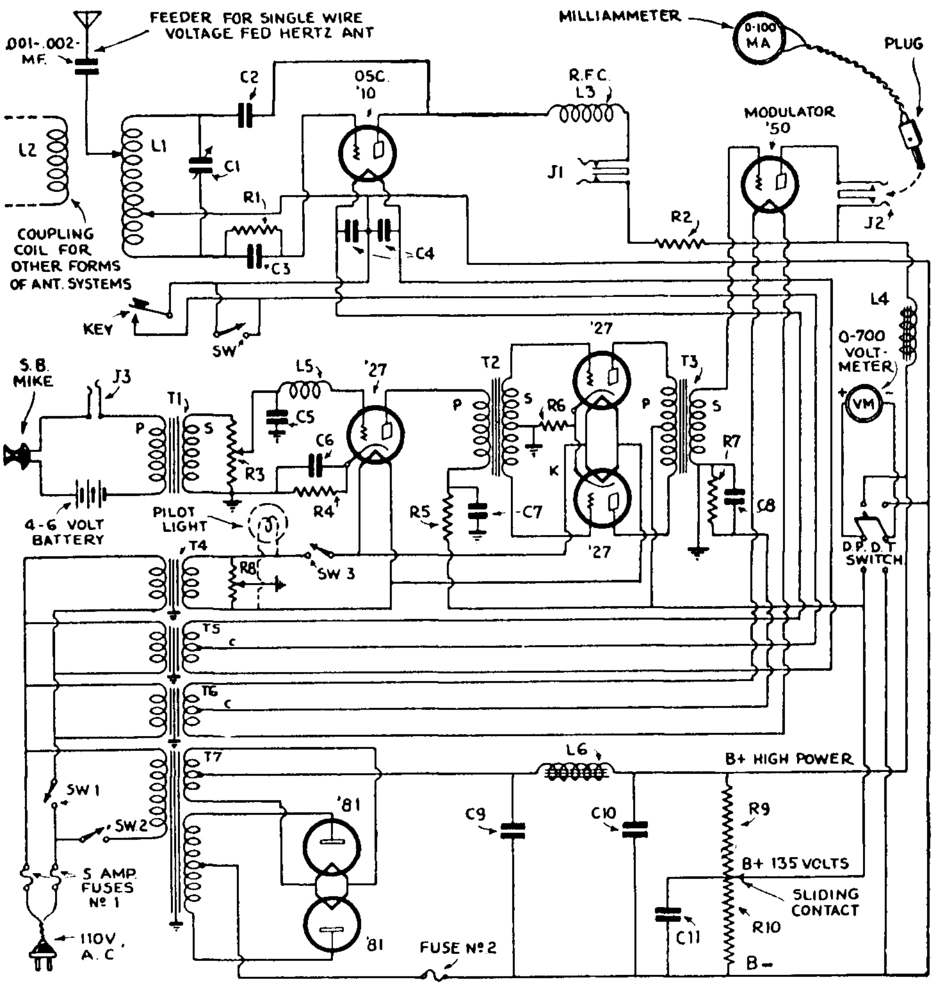


fig. 6. This modulated oscillator provided a 15 watt carrier. If properly built, the stability was passable and audio quality was quite good, considering a telephone microphone was used. The whole rig was breadboard construction on three shelves. Hams who could not afford a 10 transmitting tube, dropped the voltage and used a 245 or a 226, depending upon the state of their pocketbook.

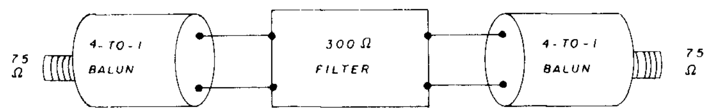
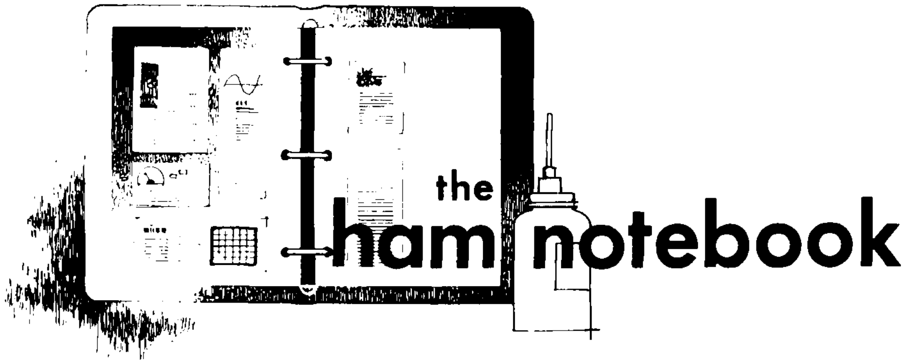


fig. 7. Balanced filter breaks both conductors of 75 ohm coax for improved TVI reduction.



rebuild your C-64 keyboard with C-16 parts

Some heavily used Commodore 64 computers develop the dreaded "keyboard stutter" and certain letters fail to print as their keys are depressed. Disassembling the keyboard unit and cleaning the printed circuit board provides only a temporary solution as the cause of the problem is a loss of elasticity in a small plastic part in the keys.

Radio Shack stores, Jameco, and All Electronics Corporation sell a surplus keyboard unit for the Commodore 16 computer. While not electrically compatible with the C-64 keyboard, the moving mechanical parts are identical and can be used to rebuild it. The job takes a couple of hours and requires more patience than technical expertise. Use the procedure below to minimize parts fall-out.

Start by removing all the keytops from the C-16 board. They can be snapped off their pins by prying under one side with a flat screwdriver while pressing against the opposite side with your thumb. There is a small coil spring under each keytop. Set the keyboard aside.

Open the C-64 by removing the three screws underneath and gently prying the unit apart at the front. Unplug the top unit from the main board and set the bottom unit aside. Remove the keyboard from the top part of the computer by taking out the large Phillips head screws. Set the top part of the case aside. Take the keytops from the C-64 keyboard and install them on the C-16 board. Make sure that a spring is in place under each keytop and each key is snapped in place on its pin.

Unsolder the two wires attached to the small gray plastic square on both keyboards. Detach the printed circuit board from the C-16 keyboard by removing the 30 or so tiny Phillips screws and gently prying it out. Discard this board and the attached wires. Take the printed board from the C-64 keyboard and install on the C-16 board, making sure the two wires are soldered in place. Discard the remains of the old C-64 keyboard.

Install the rebuilt keyboard in the C-64 and reassemble the computer. Enjoy!

Don Norman, AF8B

curing FT-101ZD key clicks

The FT-101ZD transceivers are indeed excellent. But when we used a couple of them in our multi-multi cw contest station, we received several reports of heavy key clicks from frequency neighbors.

After the contest, I spent a number of hours trying various modifications to the solid-state keying unit located in Rectifier B, with two burned-out transistors and only marginal improvement. Finally I decided that the time constants in the 12BY7A driver and the 6146s final grid bias circuits had to be changed.

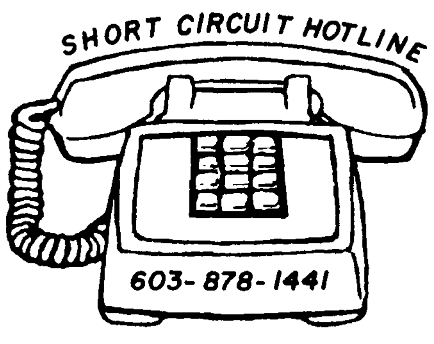
This was a quick and easy solution. Installing two capacitors — a 0.47-to-1.0 μF capacitor from the 12BY7A bias terminal to ground and a 1.0-to-3.0 μF capacitor from the 6146 bias terminal to ground — eliminated all clicks. Both of these capacitors should be rated at 200 volts dc or more, and the *positive* ends of the capacitors should go to ground. (The lower values are for the 80 wpm types. At 30 wpm, the higher values are fine, although either set will eliminate the problem.)

The modification is simple. Remove the top and bottom covers of the FT-101ZD and place the set on its right side. The Rectifier B Board is located near the back. Terminals are clearly identified: the ground terminal is towards the rear of the set on the second row of terminals; the 12BY7A bias terminal is the third terminal, and the 6146 bias terminal is the fourth terminal in that row.

I later decided that a slightly better solution would be to connect the negative end of the capacitor for the 12BY7A to the junction of R12 and R17, and the negative end of the capacitor for the 6146 to the junction of R13 and the center lead of VR01. I tried this with clip leads and it worked, but a permanent installation would have been more difficult. The keying is clickless now, so I left it alone.

(Others have tried this modification on the FT-901, FT-902, and FT-102 and found it to be effective.)

R.H. Mitchell, N5RM



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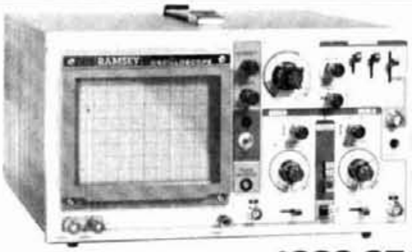

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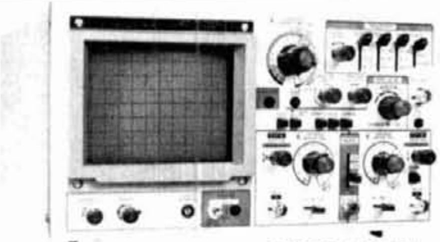

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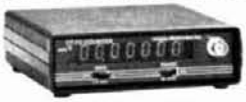

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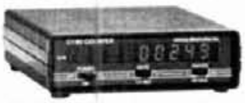

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PRACTICALLY SPEAKING ... Joe Carr K4IPV

noise, signals, and amplifiers

Gain, bandwidth, and passband shape are important amplifier characteristics on any band, but circuit noise is the first consideration with VHF, UHF, and microwave amplifiers. In the spectrum below VHF, manmade and natural atmospheric noise sources are so dominant that most receiver noises are masked. But at VHF and above, receiver and amplifier noise determines system performance. In this month's column, we look at the various sources of noise and why low-noise amplifiers (LNA) are used only in the first stage or two of a receiver or cascade chain.

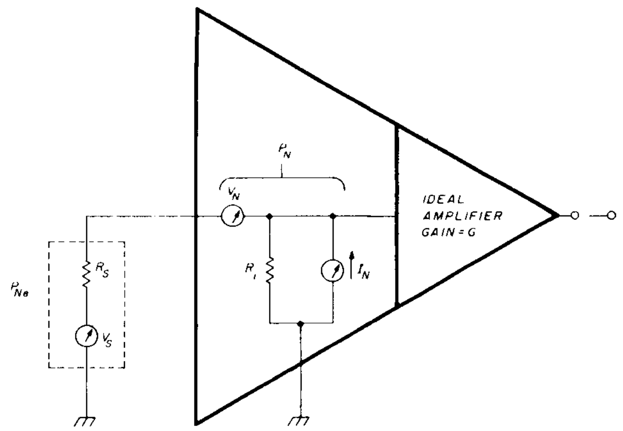

fig. 1. Amplifier equivalent circuit showing noise sources.

At any temperature above Absolute Zero (0°K or −273 °C), electrons in any material are in constant random motion and there is no detectable current in any single direction. Electron drift in any single direction is canceled over short time periods by equal drift in the opposite direction. There is, however, a continuous series of random current pulses generated in the material. Those pulses are seen by the outside world as a noise signal. This signal is called: thermal agitation noise, thermal noise, or Johnson noise.

It's important to understand what is meant by *noise* in this context. In a communications system the designer may regard all unwanted signals as noise, including manmade electrical spark signals, adjacent channel communications signals, and Johnson noise. In other cases, the harmonic content generated in a linear signal by a nonlinear network could be regarded as noise. But in the context of VHF/UHF amplifiers and receivers, noise refers to thermal agitation (Johnson) noise.

Amplifiers and other linear networks are frequently evaluated using the same methods, even though the two classes appear radically different. In the generic sense, a passive network is merely an amplifier with negative gain or a complex transfer function. We will consider only amplifiers here, but keep in mind that linear (passive) networks follow the same rules.

Amplifiers and receiver front ends are evaluated on the basis of signal-to-noise ratio (S/N or SNR). Designers try to enhance the SNR as much as possible. Ultimately, the minimum signal detectable at the output of an amplifier or receiver appears above the noise level. Therefore, the lower the system noise, the smaller the minimum detectable signal (MDS).

Noise resulting from thermal agitation of electrons is measured in terms of noise power (P_n), and carries the units of power (watts or its subunits). Noise power is found from:

$$P_n = KTB \quad \textbf{(1)}$$

where:

P_n is the noise power in watts
K is Boltzmann's constant (1.38×10^{-23} J/°K)
B is the bandwidth in Hertz

Notice in **eqn. 1** that there is no center frequency term, only a bandwidth. True thermal noise is gaussian (or near-gaussian) in nature, so frequency content, phase, and amplitudes are equally distributed across the entire spectrum. In bandwidth limited systems, such as a practical amplifier or network, the total noise power is related only to temperature and bandwidth. A 20-kHz bandwidth centered on 440 MHz should, in theory, produce the same thermal noise level as a 20-kHz bandwidth centered on 144 MHz or some other frequency.

Noise sources can be categorized as either internal or external. Thermal currents in the semiconductor material resistances produce internal noise. Consider the noise component contributed by the amplifier. If noise, or S/N ratio, is measured at both input and output of an amplifier, the output noise is greater. The internally generated noise is the difference between output and input noise levels.

External noise is produced by the signal source(s), so it is often called source noise. This noise signal is caused by thermal agitation currents in the signal source, and even a simple zero-signal input termination resistance has some amount of thermal agitation noise.

Both types of noise generator are shown schematically in **fig. 1**. Here we model an amplifier as an ideal "noiseless" amplifier with a gain of G, and a noise generator at the input. This generator produces a noise power signal at the input of the ideal amplifier. Although noise is generated throughout the amplifying device, it is common practice to consider all noise generators as a single input-referred source. This is shown as voltage V_N and current I_N.

noise factor, figure, and temperature

System or network noise can be defined in three ways: noise factor (F), noise figure (NF) and equivalent noise temperature (T_e). These properties are definable as a ratio, decibel, or temperature, respectively.

Noise Factor (F). The noise factor is the ratio of output noise power (P_{no}) to input noise power (P_{ni}):

$$F = \left.\frac{P_{no}}{P_{ni}}\right|_{T = 290\ ^\circ K} \qquad (2)$$

To make comparisons easier, the noise factor is always measured at the standard temperature (T_0) 290 °K (approximately room temperature).

Input noise power P_{ni} is the product of the source noise at standard temperature (T_0) and the amplifier gain:

$$P_{ni} = GKBT_0 \qquad (3)$$

Noise factor F can also be defined in terms of output and input S/N ratio:

$$F = \frac{SNR_{in}}{SNR_{out}} \qquad (4)$$

which is also:

$$F = \frac{P_{no}}{KT_0BG} \qquad (5)$$

where:
SNR_{in} is the input signal to noise ratio
SNR_{out} is the output signal to noise ratio
P_{no} is the output noise power
K is *Boltzmann's constant* (1.38×10^{-23} J/°K)
T_0 is 290 degrees Kelvin
B is the network bandwidth in hertz
G is the amplifier gain

The noise factor can be evaluated in a model that considers the amplifier ideal, and therefore only amplifies through gain G the noise produced by the *input* noise source:

$$F = \frac{KT_0BG + \Delta N}{KT_0BG} \qquad (6A)$$

or,

$$F = \frac{\Delta N}{KT_0BG} \qquad (6B)$$

where:
ΔN is the noise added by the network or amplifier
All other terms are as defined above.

Noise Figure (NF). The noise figure is a frequently used measure of an amplifier's *goodness*, or its departure from *idealness*. Thus it is a figure of merit. The noise figure is the noise factor converted to decibel notation:

$$NF = 10 \text{ LOG } F \qquad (7)$$

where:
NF is the noise figure in decibels (dB)
F is the noise factor
LOG refers to the system of base 10 logarithms

Example 1

Calculate the noise figure in dB of an amplifier that has a noise factor of 5.27.

NF = 10 LOG F dB $\doteq$ 10 • LOG 5.27 dB = 7.2 dB

Noise Temperature (T_e). The noise *temperature* specifies noise in terms of an equivalent temperature. **Equation 1** shows that noise power is directly proportional to temperature in degrees Kelvin, and that noise power reduces to zero at Absolute Zero (0 °K).

The equivalent noise temperature T_e is not the physical temperature of the amplifier, but a theoretical construct that is an equivalent temperature producing that amount of noise power.

The noise temperature is related to the noise factor by:

$$T_e = (F - 1)\, T_0 \qquad (8)$$

and to noise figure by:

$$T_e = \left[ANTILOG \left(\frac{NF}{10} - 1 \right) \right] \cdot T_0 \qquad (9)$$

Using noise temperature T_e, we can define noise factor and noise figure in terms of noise temperatures:

$$F = \frac{T_e}{T_0} + 1 \qquad (10)$$

and,

$$NF = 10 \text{ LOG} \left(\frac{T_e}{T_0} + 1 \right) \qquad (11)$$

Curves of noise figure and noise temperature are plotted in **fig. 2**.

Total noise in any amplifier or network is the sum of internally and externally generated noise. In terms of noise temperature:

$$P_n(\text{total}) = GKB(T_0 + T_e) \qquad (12)$$

where:
P_n (total) is the total noise power
All other terms are as previously defined.

Although the equations tend to show absolute equivalence and convertibility between F, NF, and T_e, there can be confusion about proper practices for optimizing an amplifier as regards to matching the input and source resistances. There is an optimum source resistance for minimizing input noise power and for maximum power transfer to the amplifier (source resistance equals amplifier input resistance). Unfortunately, the two optimum resistances are rarely the same. While impedance matching is useful, some common tactics are not.

Some designers modify the source resistance by adding a series or shunt resistance to the circuit. This brings the total source resistance seen by the amplifier to the optimum value for noise figure reduction. In the case cited, the noise contributed by the added resistor (KTBR) increases input noise to a point that dominates and masks amplifier internal noise. Unfortunately, this tactic while appearing to

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improve F, actually does not affect it, but deteriorates output signal to noise ratio (SNR_{out}).

noise in cascaded amplifiers

Since noise is considered a valid signal, the last stage in a cascaded amplifier chain receives an amplified version of the original input signal and prior stage noise contributions (**fig. 3**). Each stage in the chain amplifies signals and noise from previous stages, and contributes some noise of its own. The overall noise factor for a cascaded amplifier chain can be calculated using Friis' noise equation:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots + \frac{F_n - 1}{G_1 G_2 \cdots G_{n-1}} \quad \textbf{(13)}$$

or, in terms of noise temperature

$$T_e = T_1 + \frac{T_2 - 1}{G_1} + \frac{T_3 - 1}{G_1 G_2} + \cdots + \frac{T_n - 1}{G_1 G_2 \cdots G_{n-1}} \quad \textbf{(14)}$$

where:
F is the overall noise factor of N stages in cascade
T_e is the overall noise temperature of N stages in cascade
F_1 is the noise factor of stage 1
F_2 is the noise factor of stage 2
F_n is the noise factor of the nth stage
T_1 is the noise temperature of stage 1
T_2 is the noise temperature of stage 2
T_{n-1} is the noise temperature of the (n – 1)st stage
G_1 is the gain of stage 1
G_2 is the gain of stage 2
G_{n-1} is the gain of stage (n – 1).

As you can see from **eqns. 13** and **14**, the noise factor or noise temperature of the entire cascade chain is dominated by the noise contribution of the first stage or two. Typically, high sensitivity microwave amplifiers use a low-noise amplifier (LNA) stage for only the first stage or two in the cascade chain. For example, a microwave satellite ground station receiver will have an LNA at the antenna feedpoint, and then non-LNA stages following the transmission line.

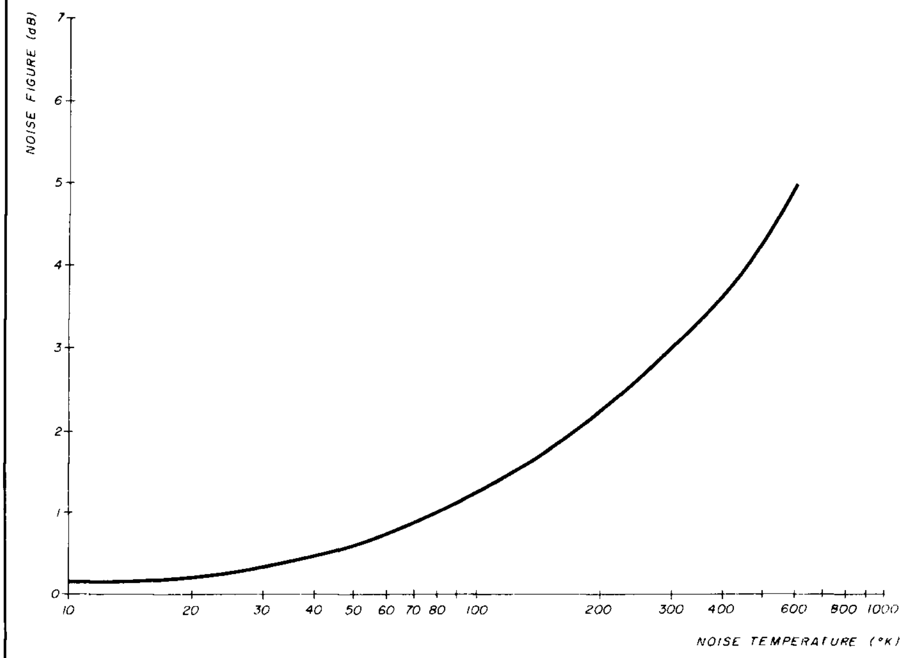

fig. 2. Noise figure vs. noise temperature.

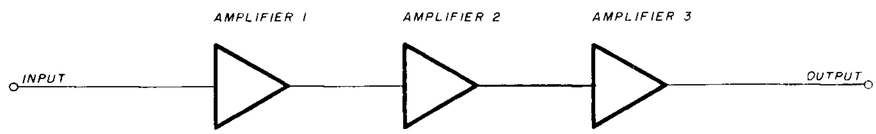

fig. 3. Cascade amplifier chain with gain = $G_1 \cdot G_2 \cdot G_3$.

example 2

A three-stage amplifier (**fig. 3**) has the following gains: $G_1 = 10$, $G_2 = 10$, and $G_3 = 25$. The stages also have these noise factors: $F_1 = 1.4$, $F_2 = 2$ and $F_3 = 3.6$. Calculate a) the gain of the cascade chain in decibels, b) the noise factor, and c) the noise figure.
Solution:

a) $G = G_1G_2G_3 = 10 \cdot 10 \cdot 25 = 2500$
G (dB) = 10 LOG 2500 = 34 dB

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2}$$

$$F = (1.4) + \frac{(2-1)}{10} + \frac{(3.6-1)}{10 \times 10} \quad \textbf{(15)}$$

$$F = 1.4 + \frac{1}{10} + \frac{2.6}{100}$$

$$F = 1.4 + 0.1 + 0.026 = 1.53$$

c) NF = 10 LOG F = 10 log 1.53 = 1.9 dB

Note in the above example, that the overall noise factor (1.53) is only slightly worse than the noise factor of the input amplifier (1.4), and is better than the noise factors of the following stages (2 and 3.6, respectively). Clearly, the overall noise factor is set by the input stage. This is why we use low-noise amplifiers in the frontends of VHF, UHF, and microwave receivers, while amplifiers later in the cascade chain are lesser quality circuits with respect to noise.

Joe Carr can be reached at POB 1099, Falls Church, Virginia 22041.

ham radio

VHF/UHF WORLD

Joe Reisert
W1JR

loose ends

I must apologize for missing the January *ham radio* issue. No, I wasn't in outer Mongolia as might have been reported!

Every so often it's good to sit back and take stock of where you are and where you're heading. This column is no exception. It's hard to believe that this is the start of the fifth year for "VHF/UHF World". Your suggestions for topics are always appreciated.

This month's column will tie up some loose ends and answer some of your letters. I appreciate all of them. I'll also update the North American VHF/UHF records.

upcoming events

Many readers send information on upcoming events. Please remember to write three or four months before an event so that I can list it.

There has been interest expressed in the meteor shower peaks. The lists came from information given in my talk at the Central States VHF Conference in Sioux Falls, South Dakota in July 1981. Later, Russ Wicker, W4WD, generated a computer program that automatically calculates the peaks. My son, Jim, AD1C, modified this program for personal computers. See reference 1.

This method of prediction relies on data and information accumulated from earlier showers. The information isn't always accurate for some of the meteor showers. If the shower radiant (the point in the sky from which the meteors seem to be emanating) is not optimum for the desired transmission direction, the peak time may be poor or nonexistent.

Reference 2 shows how to tie the shower peak and the radiant together. This information, available for personal computers, takes much of the guesswork out of meteor scatter communications. It highlights the optimum dates, a help with the shorter duration showers especially in leap years.

references

I'm told that few Amateur Radio writers cite as many references as I do. I try to make each month's column stand on its own. Sufficient material is included in each column and references are provided for those who want to dig further. While some of the references are either unavailable or difficult to obtain, I do have all of them on file.

I try to arrange the material in this column in building-block fashion so I can cite my columns in back issues. They can be purchased from *ham radio* Bookstore ($5 each postpaid, 3 for $13.95) and I often see back issues at Amateur flea markets.

state of the art

There has been considerable progress in the VHF/UHF field in the last few years. In January 1984, SSB on 23 cm (1296 MHz) was uncommon in many parts of the United States. Now commercial 23-cm CW/SSB transverters and stand-alone multimode transceivers can be purchased for all Amateur bands below 13 cm (2300 MHz). Only transverters are available for the new 33-cm (902-928 MHz) band. Commercial Amateur CW/SSB transverters now run as high as 3 cm (10 GHz).

Amateurs now use SSB as high as 48 GHz.[3] EME operation above 13 cm was an untouched field through 1986; this has changed with Amateur EME contacts now reported through 6 cm (5760 MHz) and possibly 3 cm before this column goes to print.

The VHF/UHF/Microwave field is expanding with new technology such as MMICs (monolithic microwave ICs),[4] GaAs (Gallium Arsenide) FETs, and HEMTs (High Electron Mobility Transistors) in common use.[5] I will try to keep you informed about developments.

new VHF/UHF/Microwave records

Until 1985, VHF/UHF/Microwave DX records were usually shown on the basis of worldwide accomplishments with little regard to the propagation mode. This tended to discourage record challenges since many of them took place in areas where special

table 1. North American VHF and above claimed DX records. Revised October 29, 1987.

N. American VHF and Above Claimed DX Records (notes 1 & 2) Revised 29 Oct. 1987

Frequency	Record Holders	Date	Mode DX	Miles	(km)
50 MHz	*Note 3*				
EME	K6MYC (CM97EB)-K8MMM (EN91BK)	84-07-24	CW	2127	(3422)
144 MHz					
Aurora	KA1ZE (FN31TU)-WB0DRL (EM18CT)	86-02-08	CW	1347	(2167)
Ducting	KH6GRU (BL01XH)-WA6JRA (DM13BT)	73-07-29	CW	2586	(4161)
EME	VE1UT (FN63XV)-VK5MC (QF02EJ)	84-04-07	CW	10,985	(17676)
Spor. E	KD4WF (EN92LK)-NW70/7 (DM25GV)	87-06-14	SSB	1980	(3186)
FAI	W5HUQ/4 (EM90GC)-W5UN (DM82WA)	83-07-25	CW	1228	(1976)
MS	K5UR (EM35WA)-KP4EKG (FK68VG)	85-12-13	SSB	1960	(3153)
TE	KP4EOR (FK78AJ)-LU5DJZ (GF11LU)	78-02-12	SSB	3933	(6328)
Tropo	K1RJH (FN31XH)-K5WXZ (EM12QW)	68-10-08	CW	1468	(2362)
220 MHz					
Aurora	W3IY/4 (FM19HA)-WB5LUA (EM13QC)	87-07-14	CW	1145	(1842)
Ducting	KH6UK (BL11AQ)-W6NLZ (DM03TS)	59-06-22	CW	2539	(4086)
Spor. E	K5UGM (EM12MS)-W5HUQ/4 (EM90GC)	87-06-14	CW/SSB	932	(1499)
EME	K1WHS (FM43MK)-KH6BFZ (BL11CJ)	83-11-17	CW	5058	(8139)
MS	K1WHS (FM43MK)-K0ALL (EN16NW)	85-08-12	SSB	1279	(2057)
TE	KP4EOR (FK78AJ)-LU7DJZ (GF05RJ)	83-03-09	CW/SSB	3670	(5906)
Tropo	VE3EMS (EN86QJ)-WB5LUA (EM13QC)	82-09-28	SSB	1181	(1901)
432 MHz					
Aurora	W3IP (FM19PD)-WB5LUA (EM13QC)	86-02-08	CW	1182	(1901)
Ducting	KD6R (DM13NI)-KH61AA/P (BK29GO)	80-07-28	CW	2550	(4103)
EME	K2UYH (FN20QG)-VK6ZT (QF78VB)	83-01-29	CW	11,567	(18612)
MS	W2AZL (FN20VI)-W0LER (EN35IA)	72-08-12	CW	1019	(1640)
Tropo	WB3CZG (FN21AX)-WA5VJB (EM12LQ)	86-11-29	SSB	1318	(2121)
903 MHz					
Tropo	W2PGC (FN02OR)-K3SIW/9 (EN52WA)	86-12-24	SSB	478	(769)
1296 MHz					
Ducting	KH6HME (BK29GO)-WB6NMT (DM12KU)	86-08-13	SSB	2528	(4068)
EME	K2UYH (FN20QG)-VK5MC (QF02EJ)	81-12-06	CW	10,562	(16995)
Tropo	WB3CZG (FN21AX)-KD5RO (EM13PA)	86-11-29	CW	1287	(2070)
2304 MHz					
EME	PA0SSB (JO11WI)-W6YFK (CM87WI)	81-04-05	CW	5492	(8837)
Tropo	KD5RO (EM13PA)-W8YIO (EN82BE)	86-11-29	CW	940	(1513)
3456 MHz					
Tropo	WA5TNY/5 (EM11AU)-WB5LUA/5 (EM24UQ)	86-10-19	CW	288	(464)
EME	W7CNK/5 (EM15FI)-K0KE/0 (DM79NO)	87-04-12	CW	498	(802)
5760 MHz					
Tropo	K5PJR (EM26OP)-W5UGO/0 (EN00PH)	87-07-04	CW	332	(535)
EME	WA5TNY (EM12KV)-W7CNK/5 (EM15FI)	87-04-24	CW	174	(279)
10.368 GHz					
Tropo	N6GN/6 (CM89PX)-W6SFH/6 (DM04MS)	87-07-19	CW	414	(666)
24.192 GHz					
LOS	WA3RMX/7 (CN93IQ)-WB7UNU/7 (CN95DH)	86-08-23	SSB	116	(186)
47.040 GHz					
LOS	WA3RMX/K7RUN (CN85PL)-WB7UNU/W7TYR/W7ADV (CN85NH)	87-03-07	SSB	5.42	(8.72)
76-149 GHz	None reported				
474 THz					
LOS	K6MEP (DM04IO)-WA6EJO (DM04KT)	79-06-09	LASER	15	(24)

Note 1. The records are listed alphabetically by mode. Ducting is suspected where the path is mostly over water. No efforts are made to separate out ducting on overland paths so they're grouped under tropo.
Note 2. The information within the brackets ( ) is the grid square locator.
Note 3. 6-meters records, excepting EME, were left off since the primary mode is often hard to distinguish. Also long-path QSOs have been reported during solar cycles 19 and 21 which exceed 12433 miles (20004 km).

propagation phenomenon occurs. In my July 1985 column,[6] I started listing VHF/UHF/Microwave records for North America by band and propagation mode when known. Judging from the response, this type of record listing has become very popular.

Since the last listing of North American DX records was published[3], many have been broken and some new modes were added. I have updated the DX records based on the latest information (**table 1**).

New record challenges must be carefully documented and compared with existing records. The distance is very important since records on these frequencies are often extended by as little as one mile. Therefore, precise locations (latitude and longitude) are essential. Send me an SASE if you want record forms.

table 2. This table shows some of the most famous mountain top locations that have been popularized by VHF/UHF/Microwave and Millimeter-wave'rs to set DX and contest records.

Mountain	Location (approximate)	Grid Square
Eastern USA		
Cadillac Mountain	Bar Harbor, Maine	FN54VI
Mount Washington	Glen, New Hampshire	FN44IG
Pack Monadnock	Peterborough, New Hampshire	FN42BU
Mount Equinox	Manchester, Vermont	FN33JC
Mount Mansfield	Burlington, Vermont	FN34ON
Mount Greylock	North Adams, Massachusetts	FN32JP
High Point	Port Jervis, New Jersey	FN21EH
Watchusett Mountain	Princeton, Massachusetts	FN42BL
Mount Mitchell	Asheville, North Carolina	EM85US
Mount Toxaway	Oakland, North Carolina	EM85MD
Spruce Knob	Simoda, West Virginia	FM08FQ
Western USA		
Pikes Peak	Colorado Springs, Colorado	DM78BA
Mount Rose	Reno, Nevada	DM09AQ
Mount Potasi	Las Vegas, Nevada	DM25GV
Mount Diablo	Walnut Creek, California	CM97AU
Mount Hamilton	San Jose, California	CM97EI
Mount Frazier	Frazier Park, California	DM04MS
Mount Pinos	Frazier Park, California	DM04KT
Mount Palomar	Julian, California	DM13OJ
Mount Ashland	Ashland, Oregon	CN82PB

worldwide locator system

The Maidenhead system, in its six-digit form, is accurate to within a few miles. Some contests now require grid squares or locators on QSLs; reference 3 shows how to determine yours. It's important that all VHFers know their latitude and longitude to at least 3 seconds accuracy.

Computer programs can now determine the grid square.[7] If you'd like, send me your latitude and longitude in degrees, minutes, and seconds with an SASE and I will figure your six-digit grid square.

I have upgraded the list of the most well-known mountain tops in the United States[8] to include the six-digit locators; a revised list is shown in **table 2**. If you have information, please send it for inclusion in a later column.

There is some controversy about the equity of the new ARRL VUCC (VHF/UHF Century Club) awards. Before the grid square awards, the coveted prize for VHFers was the Worked All States award. Depending on the frequency, this tended to favor either the Northeastern or Midwestern states.

The grid squares were considered a partial equalizer. Some say this, too, is unfair because grid squares are wider in the southern United States than near the Canadian border. A distance calculation program shows this to be true; the width of a grid square is about 120 miles at 30 degrees and just below 100 miles at 45 degrees latitude.[7]

It seems difficult to develop a truly fair system because of all the factors. Tropo propagation is much more prevalent in the lower latitudes, especially from Florida to Texas, and where the land is flat. Mountains, on the other hand, can either impede or aid in line-of-sight communications. Almost half of the grid squares in coastal regions are under water.

I'll be willing to bet that there is more activity, interest, and equity generated by grid squares than working all states. VUCC awards are possible on any band but how does one obtain a Worked All States on 2 meters or above without EME?

new leadership

Another new trend, the formation of clubs specializing in VHF/UHF and microwaves, is rapidly developing. Some sponsor conferences; others have monthly meetings and publish newsletters. Clubs can unite Amateurs for group projects and purchases.

My list of clubs involved in VHF and above is probably incomplete. Please help by supplying information so we can publish a list and create interest in the higher frequency bands.

antennas

I get lots of questions on antenna selection. There is no firm answer on which is the perfect antenna. Some have high gain on a long boom while others are aimed at the cleanest possible pattern on a shorter boom. Some Amateurs prefer to use single long Yagis while others like to stack shorter or longer boom antennas. Each Amateur has specific needs for his or her setup.

The cost of antennas and/or antenna improvements are usually well below the cost of improving your receiver or transmitter power (if you're not already at the legal limit). VHF/UHF

antennas are also more manageable than hf antennas.

Last May I mentioned the availability of the MININEC 3 antenna modeling program. The program is now only available through NTIS as described in **Short Circuits** in July 1987.[3] In the May column[9] I also described optimized 6 and 2-meter Yagis that are easy to duplicate, have clean patterns and high gain per unit boomlength. The materials used, and the way I matched the driven element of the 2-meter Yagi, made the driven element longer than the reflector. I was surprised that so many people duplicated it exactly as shown, since most Amateurs have their own ideas about materials. The length of the driven element in **fig. 2** is correct as shown.

Then there are the insulators and keepers used on the insulated boom 2-meter Yagi[9] and the 70-cm (432 MHz) Yagi.[10] The ones recommended use either Lexan™ or polystyrene with black carbon impregnated (for UV protection) but never nylon. The outer portion of the insulator is approximately 0.425 inches in diameter with a thickness of 0.165 inches, and the inner diameter is 0.305 inches with a thickness of 0.150 inches. The hole in the center is 0.190 inches to accomodate a 3/16 inch diameter aluminum rod. Suitable insulators are now available from Tom Rutland, K3IPW*. Tom will also supply kits or parts for the 31-element 70-cm Yagi.

stacking antennas

The basic principles and tables detailed in references 11 and 12 still apply to stacking antennas. However, there is confusion about an article on this subject by Steve Powlishen, K1FO.[13] I used wavelength for the spacings while Steve used inches. To convert wavelengths to inches, multiply the wavelength shown in **table 1**[11] to 82, 53.7, 27.3, and 9.1 inches on 2 meters, 135 cm, 70 cm, and 23 cm, respectively.

There are small differences between the stacking distances shown on Steve's and my charts. As stated in reference 11, the stacking distance is not critical and errors of 5 to 10 percent will not matter, especially if the antennas are not *over* stacked. **Figure 6** in reference 11 clearly shows this relationship. *It is always better to stack on the short rather than the long side.* This means less feedline loss, lower sidelobe levels, and a smaller mechanical structure.

I'm often asked about stacking different frequency antennas on the same boom (reference 11). Generally, the separation from another antenna on a different frequency should be at least one half the stacking distance indicated for each. The theory of effective apertures and aperture overlap is explained in the reference. For instance, if a 2-meter and 70-cm antenna are stacked on the same mast, and the recommended stacking distances are 12 and 6 feet, respectively, the 2-meter antenna should be spaced at least 6 feet from the 70-cm antenna and the 70-cm antenna at least 3 feet from the 2-meter antenna.

This may be too great a distance but there are ways to enhance spacing. One antenna could be on the main mast and the other mounted on a stacking frame to the side of the affected antennas. Different frequency antennas could be placed at other locations on the mast.

I prefer to place my 2-meter and 135-cm antennas at the top of the mast and the higher frequency antennas at the bottom or center. Thus feedlines are shorter on the antennas that would normally have greater feedline losses. If a 6-meter antenna is used, place it at the bottom, and keep it away from the 2-meter antennas. Because of large frequency differences, very little interaction should be noted between the higher frequency antennas.

If you are pressed for space, use one-half wavelength at the lowest frequency, which for a 2-meter antenna equals 40 inches, the minimum recommended distance. Before stacking antennas, draw the typical physical apertures on graph paper to visualize potential interactions.

There are simple tests to check performance after stacking antennas of dissimilar frequency. First, see that the VSWR of each antenna does not increase noticeably, then inspect the radiation pattern to be sure it is similar to what you'd expect when the antenna is mounted alone.

Another letter questions the length of phasing line (reference 12). For best power distribution, keep the phasing line length at odd multiples of quarter wavelengths. Figure the measurement of the phasing line and lengthen or shorten slightly to fit. Don't forget to consider the dielectric constant of the phasing lines to determine its length.

receivers

Readers have asked about the noise figures of commercial rigs. The majority of transceivers I've measured were over two or three years old and in the 6-to-8 dB noise figure region; they often need an external preamplifier for serious DXing.

Part of the reason for the higher noise figures is input switching, which often uses lossier solid-state switching instead of mechanical relays. Internal convenience switching and small diameter coaxial cable often add additional losses. Dynamic range is a consideration since low noise figures require moderate-to-high gain ahead of the first stage of selectivity. Then there are those terrible UHF input connectors!

A problem with many commercial transceivers is that, for economic reasons, they often employ a low first i-f such as 10.7 MHz. Additional input filtering is required to effectively suppress images and this often adds loss. The newer transceivers sport lower noise figures, typically 2 to 3 dB.

The subject of IMD is seldom given enough attention. Receivers and the later stages must be protected from large signals even out of band. This is particularly troublesome where large RF emitters such as local FM or TV transmitters are present. Gary Field, WA1GRC, recently revised his RF CAD program.[7] It now includes a quick IMD analysis and printout with up to 50 input emitters.

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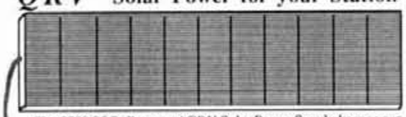

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MRF492	Q	90W	16.00	35.00
SRF2072	Q	65W	12.75	28.50
SRF3662	Q	110W	24.00	53.00
SRF3775	Q	75W	13.00	29.00
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MRF644	25W	407-512	21.00	46.00
MRF646	40W	407-512	25.00	54.00
MRF648	60W	407-512	31.00	66.00
2N6080	4W	136-174	6.25	—
2N6081	15W	136-174	8.00	—
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One reader rightly chided me for not adequately treating the transmission line losses ahead of the preamplifier. When very low noise figure preamplifiers are used in a quiet environment such as EME or above 220 MHz, and the preamplifier is not mounted right at the antenna with little or no transmission line losses ahead of it, some serious noise temperature increases can occur. This is a separate subject in itself and it will be discussed in a later column.

transmitters

There still seems to be much interest in high-power amplifiers. There aren't many to choose from and they are expensive. 8877s are now well over $500 and prices are rising. Older, more available tubes such as the 7213 and 4CX1000 are of interest but straightforward designs are still scarce.

The FAA surplus AM6154 and AM6155s for 144 through 432 MHz[14] have just about disappeared except when resold by another Amateur. One problem with this rig is the unavailability of 8930 tube replacements. Don Cook, K1DPP, and Dave Hackford, N3CX, have taken a different tack. They have machined reducer rings so that a 4CX250B tube can be substituted for the 8930. The only penalty is a reduction in output power of about 25 percent.

Some of you are interested in the use of microwave oven magnatrons. These ovens nominally operate as an unlocked pulsed oscillator in the 2400-2450 MHz frequency range, the upper segment of the United States 13-cm Amateur band. Amateurs would like to somehow phase or frequency lock the magnatrons.

I've discussed this subject with Hank Cross, W1OOP, who experimented with these ovens for industrial purposes. He notes that the magnatrons used are very difficult to move in frequency, even minimally, because of the waveguide and coupling probe design.

Furthermore, microwave ovens are designed for pulse operation and the dc voltage is generally provided by a nonfiltered half-wave rectifier. Conduction usually occurs above 1000 volts. Consequently, they would probably only be good for 75-150 watts of actual CW RF output in the 2400-2450 MHz range. Until someone can prove otherwise, your time might be wasted pursuing this type of amplifier.

propagation

Whenever Amateurs gather, radio propagation comes up. Amateurs have a unique ability to exploit radio wave propagation due to the large numbers of distributed stations.

Many thanks to Dennis L. Harrsager, N7DH, who pointed out that the constants I used in the equations for path loss on line-of-sight communications were in error.[15] I checked my file and sure enough, Dennis was right. The correct constants are 36.6 (**eqn. 1** for miles) and 32.45 (**eqn. 2** for kilometers).

Thanks also to Len Sheer, W7WRQ, for reminding me that the first radar used for EME communications[16] was not commercial but U.S. Army radar.

Finally, with the increase in solar activity as solar cycle 22 begins, there are more opportunities for long haul DX on 6 meters and transequatorial propagation on the other VHF bands. Six-meter operators will be glad to know that Norwegian stations, LA, now have 6-meter privileges (with low effective radiated power). I've learned that the Madeira Islands, CT3, and Greece, SV0, are now authorized to operate on 6 meters. Many other countries, especially in Europe, are seeking operating privileges on 6 meters now that many TV stations have abandoned channels in this frequency region.

QSLs

Recently someone suggested that I advocate a QSL card without the *QSO confirmed* or *confirming QSO*. I guess he was referring to the sample QSL shown in last June's column[2]. Until his note, I hadn't noticed that the printer didn't include all the information on my portable QSL.

I send filled-out QSLs for valid two-way contact only if I can determine from my log that I was active at that time and frequency. In the case of SWLs, I make it very clear on the QSL that it is *not* confirming a two-way contact. As a reminder,[3] always use UTC dates and time.

construction

I sometimes get requests for sources of printed circuit boards. Unfortunately, I do not have the resources to produce them. Please let me know if you want to volunteer to make up printed circuit boards for any of the circuits published in this column.

Until such time, I will continue to build most of my circuits in self-contained shielded enclosures using double-sided printed circuit material for the ground plane. This method is easy and effective even up through the UHF bands.

product reviews

This column is primarily aimed at the experimenter who builds some of his own gear, so I will probably continue to recommend homebrew gear with a smattering of recommendations where commerical gear is appropriate. If there are specific items that you'd like reviewed, please let me know.

computers in the ham shack

Without a doubt, one of the most important recent innovations for the Radio Amateur is the personal computer. At first it was primarily used for simple calculations and repetitive operations. Nowadays it's being used for logging, word processing sending/receiving Morse/RTTY/Packet, propagation and satellite predictions, schematic layout, and printed circuit board generation. Looks as if we'll all have to adapt to the personal computer!

summary

This month's column was sort of a *potpourri*. Hope I've answered some of your questions and cleared up any misunderstandings. I'm always receptive to new ideas and suggestions on topics or how I can improve the column. Please let me know.

acknowledgments

Although many shall remain anonymous, thanks to all those who wrote to me either with questions, suggestions, or constuctive criticism. Without your support, this column would not be possible.

important VHF/UHF events:

February 17	*EME perigee*
February 17	*New moon*
March 16	*EME perigee*
March 18	*New moon (eclipse of the sun)*
March 21	*±2 weeks. Optimum time for TE propagation.*

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ham radio

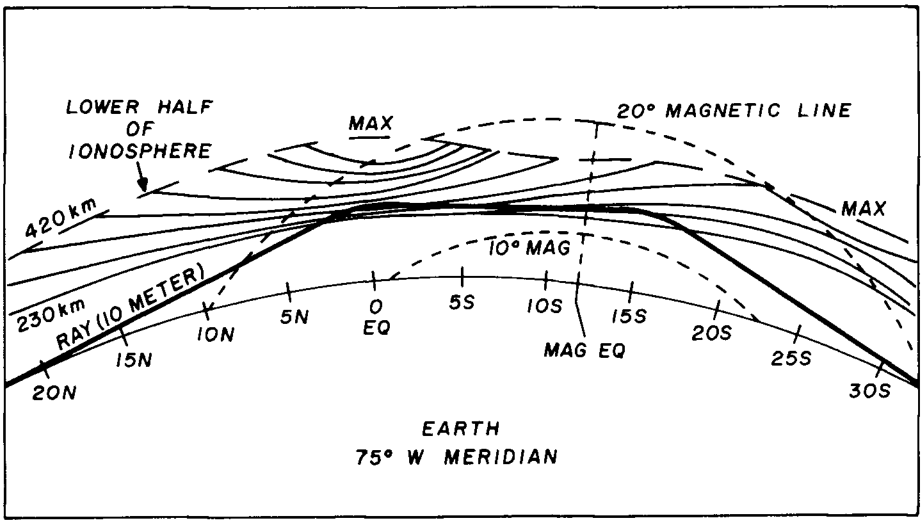
DX FORECASTER

Garth Stonehocker, K0RYW

transequatorial DX update

Transequatorial (TE) propagation is the mode used by signals that cross the equator by other than the normal one-to-three hop ionospheric F, or sporadic E paths. Intermediate ground reflections over a 5000 to 7000 mile path don't occur as shown in the figure's heavy ray traced line representing a 10-meter operating frequency. The indicated contour lines are equidensity electron profiles for the lower half of the ionosphere along the 75 degree west meridian (East Coast of United States and down through western South America). These contours show two maximum regions: one above the geographic equator and the other about 25 degrees south. These two high density areas create the ionospheric tilt that makes TE possible.

Ionospheric scientists have studied the phenomena with vertical and oblique ionosondes to determine the conditions necessary for this to occur and how it can be used in communications. The diurnal conditions are normal during the morning as the sun builds the lower ionosphere at the subsolar point. By local noontime the single maximum of ionization that has grown in the F region begins to divide and shift locations, one moving further south and the other to the north. Throughout the afternoon and evening (to about 2000 local time) the electron drift continues to build at ± 15 degrees on either side of the geomagnetic equator, represented by the dotted line and circle segments in the figure. These building conditions are effective only in the wintertime, September through March, the *same* months for both hemispheres (an unusual condition not yet understood). The magnitude – density or maximum usable frequency – of these maxima increases with sunspot number or solar flux. This is further enhanced during geomagnetic disturbances, when electrons from polar solar wind particles also drift up the field lines from the auroral zone associated trough to the geomagnetic equator's maximum areas.



These studies have brought about an understanding of the conditions needed for TE-DX. Transequatorially propagated signals are only 6 to 30 dB weaker than a significantly shorter (up to one eighth the distance) one-hop signal and 8 to 10 dB stronger than a 3-hop signal. The TE signals are not as coherent, fading independently of the usual F region propagation. Hence TE signals are like scatter propagation except for signal strength. To access TE, use a low take-off angle antenna. It should skip the intervening ground reflection and experience horizon focusing that will strengthen the signal. The best TE direction should be a bearing in a direction perpendicular to the geomagnetic equator, although the aspect angle width is about 50 degrees wide. The geomagnetic equator at 75 degrees west is 12 degrees south of the geographic equator (GE). It then curves north to cross the GE at 25 degrees west, becoming north 10 degrees at 0 degrees longitude. That latitude is continued to the Phillipines where it curves down to meet South America, again crossing the GE at about 70 degrees west. I hope you will have opportunities to try TE-DX by springtime before it begins to wane for the summer.

last-minute forecast

The first two weeks of the month are expected to favor the lower frequency bands, 30 through 160 meters, with nighttime DX and some one-hop daytime short skip for intracontinental contacts. The noise levels should be low on these bands. However, expect weak and fading signals around February 8th, 17 to 19th, and 27th. The higher frequency bands should be best

FEBRUARY

WESTERN USA

	2300	2200	2100	2000	1900	1800	1700	1600	1500	1400	1300	1200	1100	1000	0900	0800	0700	0600	0500	0400	0300	0200	0100	0000	GMT
	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	PST
ASIA FAR EAST	20	30	30	40	40	40*	40	40*	30	30	30	30	30	30	30	20	20	20	20	20	20	20	20	20	N
EUROPE	20	20	20	20	20	20	20	20	30	30	30	40	40	40	40	30	30	30	30	30	30	30	20	20	NE
S. AFRICA	15	12	12	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	E
S. AMERICA	12	10	10	10	10	12	12	12	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	SE
ANTARCTICA	15	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	15	15	15	15	15	15	15	S
NEW ZEALAND	10	10	12	12	15	15	15	15	20	20	20	20	20	20	20	20	20	15	15	15	12	12	10	10	SW
OCEANIA AUSTRALIA	10	12	12	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	12	12	10	10	10	W
JAPAN	20	20	20	20	30*	30	30	40*	40*	30	30	30	30	30	30	30	20	20	20	20	20	20	20	20	NW

MID USA

	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	MST
ASIA FAR EAST	30	30	30	30	20	20	20	20	20	20	20	20	30	30	30	40	40	40	40	40	30	30	30	30	N
EUROPE	20	20	20	20	20	20	20	20	20	30	30	30	40	40	40	40	40	40	30	30	30	30	20	20	NE
S. AFRICA	12	10	10	10	12	12	12	12	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	E
S. AMERICA	12	12	10	10	10	10	12	12	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	15	15	S
NEW ZEALAND	10	10	10	12	12	12	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	12	12	10	SW
OCEANIA AUSTRALIA	10	10	12	12	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	12	12	10	10	W
JAPAN	20	20	20	20	30	30	30	30	30	30	40*	40*	30	30	30	30	30	20	20	20	20	20	20	20	NW
	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	CST

EASTERN USA

	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	EST
ASIA FAR EAST	30	30	30	20	20	20	20	20	20	20	20	20	30	30	30	30	40	40*	40	40	30	30	30	30	N
EUROPE	20	20	20	20	20	20	20	20	20	30	30	40	40	40	40	40	40	30	30	30	30	30	20	20	NE
S. AFRICA	15	12	10	10	10	12	12	12	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	E
CARIBBEAN S. AMERICA	15	12	10	10	10	10	12	12	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	S
NEW ZEALAND	10	10	10	10	12	12	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	12	10	10	SW
OCEANIA AUSTRALIA	10	10	10	10	12	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	12	12	10	10	W
JAPAN	20	20	20	30	30	30	30	30	30	30	30	40*	40*	30	30	30	30	30	20	20	20	20	20	20	NW

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.
*Look at next **higher** band for possible openings.

the latter two weeks, particularly the 20 to 25th when the MUFs should be 10 to 15 percent above the median (18 MHz noontime midlatitude estimate). Expect TE openings these weeks with enhancements during disturbed periods.

No significant meteor showers are scheduled to appear in February. A full moon will occur on the 2nd, with its perigee on the 17th.

band-by-band summary

Ten and twelve meters, the highest day-only DX bands, are nearest the MUF for southern hemisphere paths. They will be open most days when the solar flux is above 85 during the 3-to-5 hour period centered on local noon. These bands open on paths toward the east and close toward the west. The paths are up to 2400 miles (4000 km) in single-hop length, and on occasion triple that during evening transequatorial openings.

Fifteen meters, a day-only DX band open most days, will be best when the MUF is slightly above this band in a transition period that occurs right after sunrise and just before sunset. Transequatorial openings will occur, with distances similar to 10 and 12 meters.

Twenty, thirty, and forty meters are both daytime and nighttime DX bands. Twenty is the maximum usable band for DX in the northern directions during the day. In combination with 30 meters, it provides nighttime paths for the day-only bands. Forty meters becomes the main over-the-pole DX nighttime band, with some hours covered by 30 meters.

Eighty and one-sixty meters, the night-only DX bands, exhibit short-skip propagation during daylight hours, then lengthen at dusk. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the north-south near midnight, and ending up in the Pacific areas for a few hours before dawn. On some nights, 80 meters, with its higher signal strengths, will be the best band to use. One-sixty is also expected to provide similarly good conditions.

ham radio

NEW products

two new hf transceivers

Kenwood has announced two new high performance HF transceivers: the TS-140S and the TS-680S. The TS-140S is an all band, all mode, 100 watt HF transceiver with general coverage receiver. The all band, all mode, 100 watt HF TS-680 transceiver, includes a ten watt, six meter section.

The new programmable band marker is useful for staying within the limits of your ham license and prevents out-of-band operation. For contesters, there is a program in the suggested frequencies to prevent QRM to non-participants.

A Morse Code beeper status indicator has been included. The indicator verifies the operating mode with Morse Code characters, signals empty or full memory banks, and lets you know when frequency lock is on.

Other features are: dual digital VFOs, 31 memory channels (ten of which can store receive and transmit frequencies separately for repeater or cross band operation), programmable scanning, and automatic selection of USB or LSB. Kenwood interference reducing circuits: IF shift, dual noise blankers, RIT, RF attenuator, selectable AGC, and FM squelch are also included. Suggested retail price is $899.95 for the TS-140S and $999.95 for the TS-680S.

A complete line of accessories is available. For details contact Kenwood Communications and Test Equipment Group, 2201 E. Dominguez Street, Long Beach, California 90810.

resonators and a new mobile antenna

Hustler, Inc. has introduced two new resonators for use with the H.F. Mobile System. Specifically designed for WARC bands it is completely compatible with your present H.F. Mobile System, using the MO-1, MO-2, or MO-3 mast.

The RM-12, 12 meter resonator has a bandwidth 90-120 kHz under 2:1 or better, a 400 watt rating, for use with the Hustler Mobile H.F. System, for $13.95.

The RM-17, 17 meter resonator with a bandwidth 150 kHz under 2:1 or better and 400 watt rating also for use with the Hustler Mobile H.F. System priced at $19.95.

Hustler's RMX 10 meter Super Mobile antenna has a 1000 watt rating, with a bandwidth 350 kHz under 2:1 or better. Including spring, it is 48 inches tall and can be mounted using the Hustler HLM or TLM. The coil is compatible with the Hustler full size Mobile H.F. System. Available in black, white, or red, the antenna is priced at $31.95.

For further information contact Hustler, Inc., One Newtronics Place, Mineral Wells, Texas 76067.

Circle #301 on Reader Service Card.

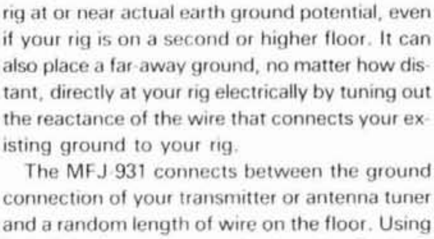
improving rf ground

Don't we all sometimes have problems with not having a good rf ground — problems like rf "hot spots" that "bite" our lips or fingers when we transmit; rf feedback that causes our rigs to quit working on certain bands; excessive rf coupling to ac lines that causes everything to quit working; neighbors screaming about TVI or RFI; computers spewing out gibberish; being unable to talk across town because of extreme ground losses or radiation pattern distortion?

The new MFJ-931 creates an artificial rf ground with just a random length of wire thrown along the floor. It's very effective at placing your rig at or near actual earth ground potential, even if your rig is on a second or higher floor. It can also place a far-away ground, no matter how distant, directly at your rig electrically by tuning out the reactance of the wire that connects your existing ground to your rig.

The MFJ-931 connects between the ground connection of your transmitter or antenna tuner and a random length of wire on the floor. Using its built-in rf ammeter, two knobs adjust for maximum rf ground current; this resonates the random wire, converts it into a tuned counterpoise, and presents an effective low impedance near ground potential to your rig, thus creating an artificial rf ground.

To place a distant ground directly at your radio equipment electrically, simply connect the MFJ-931 between your rig and connecting ground wire. Adjust its two knobs for maximum rf current; this tunes out the reactance of the connecting wire, reduces the electrical ground lead length to virtually zero, and electrically



places your distant ground directly at your rig.

The MFJ-931 covers 1.8 to 30 MHz. Ruggedly built, it's housed in an all-aluminum cabinet with a brushed aluminum front panel, measures 7-1/2 x 3-1/2 x 7 inches, and retails for $79.95.

For additional information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762.

Circle #302 on Reader Service Card.

Kanterm-PC™

Kanterm-PC™ is a new terminal program for IBM PCs and compatibles designed specifically for use with its KPC-4 and KAM dual-port units.

Kanterm-PC features five different selectable screen configurations created to enhance dual-port operation. Screen options include the horizontal port split, the vertical port split, hf (port 1) only, VHF (port 2) only, and the standard combined port 1/port 2 full-screen display. All screen displays include a separate transmit display window.

In addition to screen options, Kanterm-PC also features 37 user-loaded buffers of 254 characters each, and a pop-up menu that displays additional program options. Kanterm-PC includes a real-time clock and date display, word wrapping of received text, and many other valuable features.

For further information, contact Kantronics, 1202 East 23 Street, Lawrence, Kansas 66046.

Circle #303 on Reader Service Card.

tv/fm interference highpass filters

Ameco has introduced two new tv/fm interference highpass filters. The new models available are HP-75T, for 75 ohm applications, which comes with a length of coaxial cable for easy installation, and HP-300T, for twin-lead applications, which has a twin-lead termination for simple installation. Each filter contains nine shielded sections, with a total of 25 elements in five individually shielded compartments. They provide 70 dB attenuation below 50 MHz. The list price for each is $12.95.

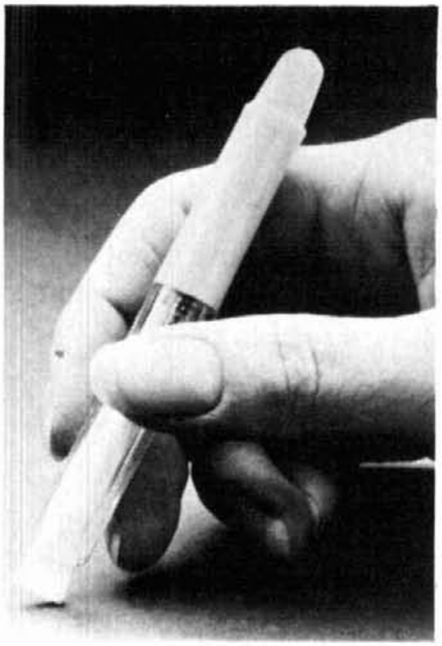
For further information, contact Ameco Equipment Company, 220 East Jericho Turnpike, Mineola, New York 11501.

Circle #304 on Reader Service Card.

117

118

NEW products



cleaning, polishing tool

The Eraser Company has announced the availability of its new Superbrush Cleaner and Burnisher Kit, which contains a Superbrush holder with a coarse-grade FybRglass refill installed and two extra refills made of stainless steel wire and fine-grade FybRglass. Each gives a different cleaning action, ranging from a fine polish to a coarse, sandpaper finish.

Priced at $6.98, the Superbrush may be used to clean, burnish or polish a variety of surfaces. Some of its many uses include removing rust spots, cleaning electrical contacts, roughing surfaces before using epoxy or glue, cleaning lead in stained glass, deburring small metal parts, scrubbing stains off pots and pans, and a number of other household and office uses. A twist-tip control allows the entire length of the refill to be used; replacement refills are available in packs of two.

For details, contact The Eraser Company, Inc., P.O. Box 4961, Olivia Drive, Syracuse, New York 13221.

Circle #305 on Reader Service Card.

tiny TNC-2 announced

Pac-Comm announces a totally redesigned TNC-2 clone, which uses unmodified TAPR TNC-2 software. This new Tiny TNC-2 uses an integrated circuit modem and simplified circuitry to insure reliability, size reduction and a much lower cost. This unit includes as standard features 32k RAM and 32k EPROM, RS 232 and TTL compatibility, watchdog timer, modem disconnect header, switch selectable terminal baud rates and 12V DC operation. The Tiny TNC-2 is approximately 50 percent the size of the original TNC-2. This is not a kit, it's a fully assembled and tested 1200 baud unit, ready for use on 10 meters and VHF. For additional information contact Pac-Comm Packet Radio Systems, Inc., 3652 W. Cypress Street, Tampa, Florida 33607.

Circle #306 on Reader Service Card.

self-teaching Novice package

The *New Novice Voice Class* instructional package from Master Publishing contains everything would-be hams need to learn about, prepare for, and pass the new Novice examination and obtain a license. Two audio cassette tapes and a 112-page book provide all the material covered by the code and theory elements of the Novice Class examination.

The two-sided, 44-minute cassette tapes teach the student how to receive and send 5-WPM CW; the book contains the 302-question pool — with explanations of the answers — from which examiners select 30 questions for the theory examination. Charts, diagrams, and equations illustrate the subject matter. A copy of Form 610 is included, as are directions for the examiners.

The *New Novice Voice Class* package is available in Radio Shack stores for $19.95. For more information, contact Master Publishing, P.O. Box 834158, Richardson, Texas 75083-4158.

Circle #309 on Reader Service Card.

network design, analysis program

RF Notes No. 4 (Network Analysis) from Etron is the sixth in a series of design assist programs for problems frequently encountered in rf and analog design. R, L, and C elements (up to 14 sections) may be entered in schematic diagram form (you actually draw the schematic as you go) or entered directly. Output data is in tabular and graphical form. Graphics plot parameters are user-selectable. Final designs can be saved and retrieved (export/import) in schematic form. Designs from the other Etron *RF Notes* programs (proper version) can be imported for Edit and Analysis.

Priced at $179.00, plus shipping and handling, the fully menu-driven program includes a tutorial section, is color/monochrome selectable, and runs on the IBM PC, PCXT, PCAT, PCjr. (Enhanced), PC/MS DOS 2.1, 256K. A color graphics card (CGA) is required.

For information, contact Etron RF Enterprises, P.O. Box 4042, Diamond Bar, California 91765

Circle #310 on Reader Service Card.

new MS-DOS logbook

QSO PRO from Morlan Software is a high-quality MS-DOS logbook. QSO-PRO was developed for radio amateurs who are committed to keeping track of QSO details. The problem is compounded when keeping track of countries worked and verifying which of those states (or countries) still haven't sent that all-important QSL card.

QSO PRO is designed to be user friendly and stores information on the most often used statistics involved in the average QSO. Additionally, it offers the convenience of REAL TIME LOGGING; PRESET for presetting routine responses; DUPE checks all logged contacts and brings important facts to the current contact. The BROWSE command allows scanning for a specific response using one letter to the entire response.

QSO PRO is self documenting (using help screens). MEMO allows storing personal notes of up to 55 characters about each contact. CURSOR CONTROL gives the ability to easily change any of the information of any contact. TALLY QSO PRO counts QSOs, QSLs sent and QSLs received by state and country. TALLY also ignores any QSOs that are marked as a duplicate.

QSO PRO List Price is $39.95 and is available through the *ham radio* Bookstore (add $3.50 for shipping and handling) or can be ordered directly from Morlan Software, P.O. Box 2400, East Liverpool, Ohio 43920-0400.

Circle #307 on Reader Service Card.

components catalog

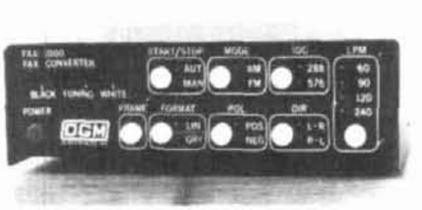
A new catalog from Mouser Electronics offers 176 pages featuring 16,000 items including capacitors, potentiometers, resistors, transformers, lamps, switches, batteries, holders, jacks, plugs, speakers, knobs, fuses, semiconductors, hardware, tools, test equipment, relays, cabinets, meters, alligator clips, delay lines, crystals, soldering equipment, integrated circuits, heat sinks, fans, diodes, coils and chokes, connectors, microprocessors, and more.

For a free copy, contact Mouser Electronics, P.O. Box 719, Mansfield, Texas 76063.

Circle #308 on Reader Service Card.

FAX-1000 fax converter

DGM Electronics, Inc. has just introduced the FAX-1000 Facsimile Converter to its line of quality communications products. The FAX-1000 simply connects between your communications receiver and any Epson graphics compatible printer.



The FAX-1000 allows you to print weather charts, satellite pictures, and press photos. It will copy a-m facsimile signals sent by satellite or fm facsimile signals, which are normally sent on the hf frequencies. It will copy all standard speeds and indices of cooperation. Pictures can be inverted or printed in either direction. A ten-segment bar graph allows accurate tuning of the station being copied.

The FAX-1000 can be operated in the automatic or manual mode. In the automatic mode, it will wait for the appropriate signals from the sending station to start, frame, and stop printing. In the manual mode, the operator can start the printing and manually frame the picture. Front-panel LED indicators and pushbuttons make the FAX-1000 easy to operate.

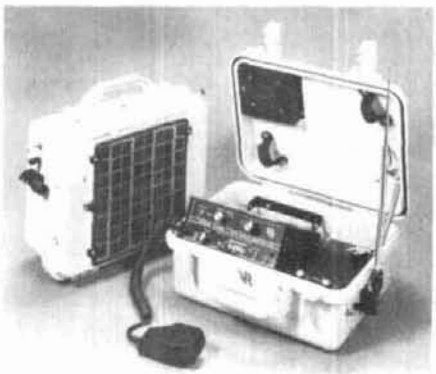
Housed in a compact, RFI-proof aluminum enclosure, the FAX 1000 measures only 7 x 2 x 6 inches deep. The unit is powered by a 110 VAC wall transformer (included), and priced at $299. For more information, contact DGM Electronics, Inc., 901 Elmwood Avenue, Beloit, Wisconsin 53511.

Circle #311 on Reader Service Card.

marine radio

The new Vector Radio VR-50 transceiver, priced at $1,295, offers worldwide hf, SSB communications in the Amateur, marine, aircraft, and emergency rescue bands. Powered by a rechargeable battery which is maintained at full charge by a photovoltaic solar panel on top of its waterproof floating case, this self-contained, portable unit measures only 14 x 11 x 6 and weighs 16 pounds.

Its 8-foot telescoping whip antenna stores in-



side the case when not in use. While an internal antenna tuner is built in, the VR-50 may be used with a wide range of auxiliary antennas.

Crystal-controlled for "on the money" tuning on a total of 24 channels in the 1.8- to 17-MHz bands, the unit puts out 50 watts PEP on voice and 25 watts on CW (code). Type acceptance by the FCC on marine and aircraft frequencies is expected by November 1987.

For information, contact the Vector Radio Company, 3207 Roymar Road, Oceanside, California 92054.

Circle #312 on Reader Service Card.

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ELMER'S NOTEBOOK

Tom McMullen, W1SL

Standing Wave Ratio — what does it mean?

Every radio amateur has at least heard the term, many have measured it and some know what it means.

SWR or VSWR?

SWR or Standing Wave Ratio is a term often used by Amateurs and others in the rf world. But this neat, easy-to-say phrase, doesn't really tell us what kind of waves it describes.

The proper term is VSWR or Voltage Standing Wave Ratio. Some people even try to pronounce it — something in the order of "vizwahr" comes out. I'm not sure that pronouncing it improves anything, but the electronic world is full of buzz words. However, back to the questions: what is VSWR, how do we get it, and how is it measured?

Figure 1A shows a cross-section of coaxial transmission line terminated with a resistor (or load that appears as one) that matches the impedance of the line. Stick a voltage-sensing probe in the transmission line, move along the line for a considerable fraction of a wavelength at the frequency of your transmitter, and you will get an equal voltage reading at *all* points. This is called a *flat* line. There are no peaks or depressions in the voltage readings from one end of the line to the other. The VSWR is 1:1.

How does it become *unflat?* Let's lay the groundwork for understanding that by looking at some basic dc and ac theory. When you apply voltage to a resistor, current flows through it, and the energy (power) is dissipated as heat. That's the only thing a resistor can do with energy — turn it into heat and get rid of it by letting the air (water, oil, metal, or some other medium) soak it up. It accepts all the energy you give it.

Applying dc voltage to a coil (inductance) doesn't produce the same result. When current flows in a wire it creates a magnetic field. If that field cuts across any nearby wires, it creates a counter emf. The counter emf tries to make current flow in the opposite direction of the one that created the field and the two currents oppose (buck) each other. Consequently, maximum current flow is delayed while voltage is not and they get out of phase. After the initial surge of current flows through the coil, the field becomes steady and the maximum dc current flow is determined by the resistance of the wire and the applied voltage.

Things get more complex if you apply an ac voltage to an inductor. The magnetic fields build and decrease with each ac cycle. When the field builds in the first quarter of a cycle, it opposes current flow. As the ac voltage decreases during the next 1/4-cycle, the field decreases. A decreasing field creates a current flow that reinforces the original current. During the next two 1/4-cycles the process is repeated and the voltage and direction of current flow reverse, causing the magnetic fields to increase, decrease, and reverse as well. The result is a complex interaction of current flow that slows down and speeds up as determined by the frequency of the ac applied voltage and the reactance of the coil. If the reactance is high enough at the frequency of the ac, almost no current will get through because the magnetic fields are strong enough to oppose it.

To understand the difference between dc and ac power flow remember: with dc maximum voltage and current occur at the same time, and with ac this happens only when the load is a pure resistance (or by the use of electronic trickery appears as one). At all other times, the magnetic fields cause the voltage and current flow peaks to be out of phase. Now, let's see what this has to do with antennas and VSWR.

the antenna as a resistor

An antenna that is exactly the right length for the frequency applied is *resonant.* This antenna accepts all the energy it is given and radiates it into space — much like the resistor that radiates all its energy as heat. The more you give it, the more it radiates. Such an antenna can be called *resistive* or *matched.*

When the antenna length is not right for the frequency, some of the applied energy is radiated into space but some is used in opposing the current flow by means of magnetic fields, just as with the inductance. These *reactive* anten-

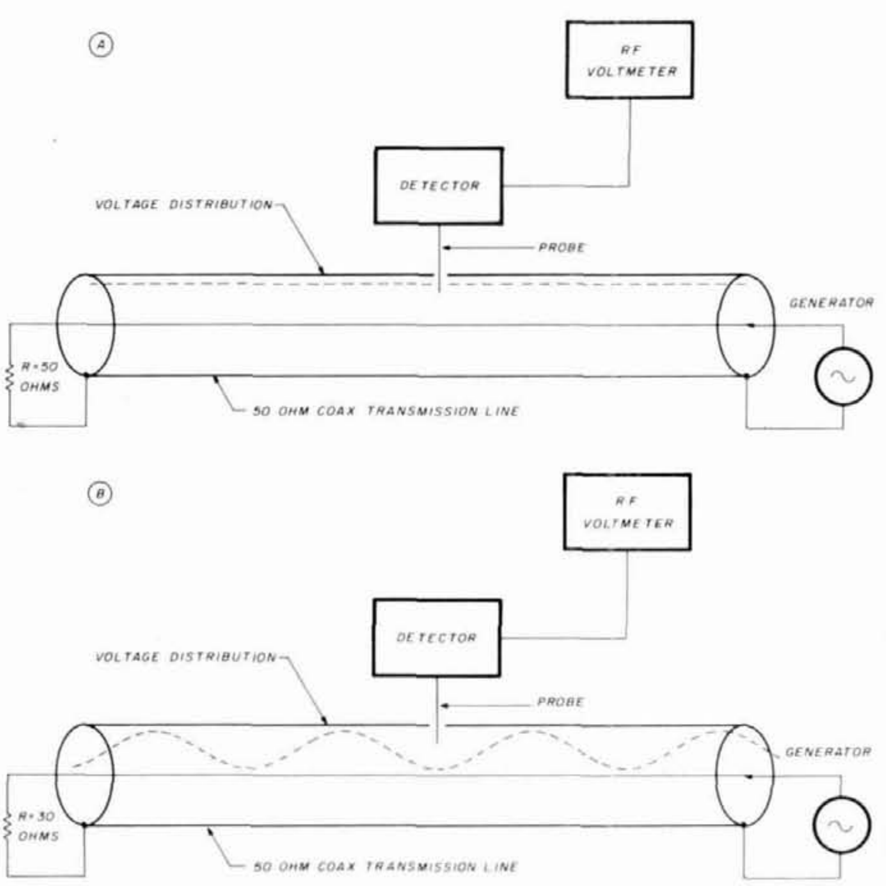

fig. 1. When the load at the end of a transmission line is the same as the impedance of the line, the rf voltage along the line is constant, so the line is called "flat" as at A. If the load (resistance) is not the same as the line, then reflected, out-of-phase voltage interacts with rf energy from the transmitter, which causes peaks and valleys in voltage readings along the line, as at B.

nas cause some energy to be reflected back down the transmission line toward the transmitter and the phase of the reflected energy is not the same as the energy coming from the transmitter. This reflected, out-of-phase wave interacts with the next cycle of energy from the transmitter producing peaks and valleys of voltage along the transmission line. By measuring the voltage with a probe, you get readings with curves like those in **fig. 1B**. To determine the VSWR, use the formula VSWR = Vmax/Vmin. Note that the ratio is always greater than (or equal to) 1, and expressed as 1:1 (for a flat line), 2.3:1 (for moderate SWR), or 11:1 (for a badly mismatched system).

A standard laboratory method of determining VSWR involves moving a probe along a calibrated section of transmission line (either coaxial line or a waveguide), and using the readings for calculations. Commercial SWR meters that use loops to sample rf current in a transmission line are calibrated against this voltage-probe type of instrument (called a *slotted line* because of its construction).

amateur instruments

With this background on VSWR, you can understand the performance of the many commercial and homebrew SWR meters available. It is impractical to keep a slotted line in your backyard to check an 80-meter antenna; it would have to be more than 65 feet long and even on the UHF bands is a slow, unwieldy tool. Also, they are made to work with only a few milliwatts of power. The current-pickup-loop type is by far the most common instrument available though some meters use a bridge circuit. Both types sample energy in a very small portion of a wavelength, translating that energy into a meter reading using diodes and resistors to create dc from the rf energy that is picked up. **Figure 2** shows the basic circuitry of two types of current-loop devices and one bridge type. None of them approach the accuracy of a slotted line, but are more than adequate for our purposes.

There are three steps to take when checking out your antenna/feedline/transmitter system. First tune your transmitter with the power/SWR meter connected between it and a good 50-ohm dummy load (assuming you are using 50-ohm coaxial cable*). Second connect the dummy load at the far end of your transmission line and check the power and SWR again. If your cable is good, the reading should be the same as when the dummy load was near the transmitter. If not, your cable is not as good as the salesman said it was. Finally, if the cable checks out, connect the antenna. Any reflected power (SWR) that shows up now indicates that the antenna is not matched. Don't get excited if there is some — few antennas are totally resistive and equal to the transmission line characteristic impedance. A ratio of 1.5:1, 2:1, or more can be tolerated. Check the instruction book for your transmitter's limitations. The SWR you can live with is usually determined by what the final amplifier stage can handle and many solid-state units shut down before it gets high enough to do any damage.

If you want to be a purist, check the transmission line with the power/SWR meter after testing it with the dummy load. By comparing the power readings at both ends, you will see how much is being lost in the cable. Some power is lost because of the wire's resistance, but more is lost if the cable has poor quality dielectric or has been contaminated with moisture or pollution. After you have this bit of good (or bad) news, keep the meter at the far end of the transmision line and check the antenna SWR right on the spot. Be careful here; rf burns can be nasty. You can do all this at relatively low power if your meter is sensitive enough. To keep from annoying fellow occupants on the band, use the lowest power possible; preferably when the band is not open or fairly inactive.

How can an SWR meter fool you into thinking all is well? The voltage and current at the current pick-up point is of such amplitude and phase that it makes the SWR look right and you assume the line is matched. But you notice that the transmitter doesn't load just right, is putting out lower power, or the tuning is super critical. To troubleshoot the problem, add or remove a section of transmission line and see if the SWR reading changes.

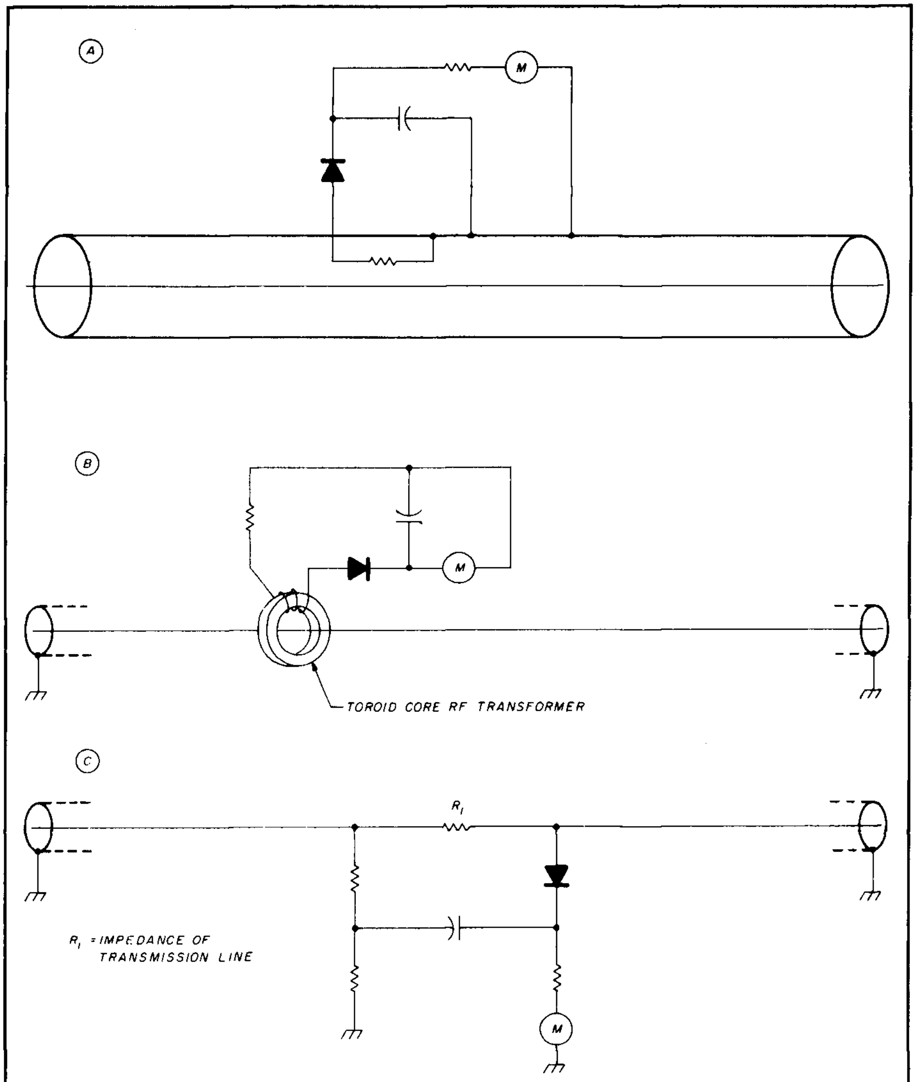

fig. 2. Most common SWR meters used by Amateurs have a current-pick-up loop to sample the rf current in the transmission line. Two types of current-sensing circuits, in simplified form, are shown at A and B. Shown at C is a bridge type of circuit used in some instruments. By adding more components and a switch, any of these circuits can be used to read forward power or reflected power.

*We call it 50-ohm cable, but the impedance is usually closer to 52.5 ohms for most coaxial line.

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ANTENNAS ANTENNAS Austin VHF/UHF Tri-Banders, Yagis HF antenna experimenters kit, books. SASE brings lists. Ed Noll, W3FQJ, POB 75, Chalfont, PA 18914.

HAM LAB PROJECT. Want several pieces HP G-382A variable attenuator. Will consider any repairable condition. K6GOX, PO Box 10, O'Neals, CA 93645 (209) 868-3548 Collect.

30 YEARS PROVIDING QSL's. Full color Old Glory, Cartoon. Also Parchment, Golden Eagle and others. Free samples. SASE appreciated. Rusprint, Rt 1, Box 363-HR, Spring Hill, KS 66083.

YAESU FT-727-R COMPUTER INTERFACE. For info write Gerald Hogsett Consulting, 1581 Woodland, Palo Alto, CA 94303.

RUBBER STAMPS: 3 lines $5.00 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave ... all types available. Large stock. Next day delivery, most cases. DAILY ELECTRONICS, PO Box 5029, Compton, CA 90224. (213) 774-1255.

CUSTOM MADE EMBROIDERED PATCHES. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. HEIN SPECIALTIES, Inc., Dept 301, 4202 N. Drake, Chicago, IL 60618.

ANNOUNCING a new heavy duty C-64 Commodore replacement power supply especially for the Packet Radio Amateur. The new higher amperage output will now allow for 24 hour continuous "Packet" operation without voltage change or failure which the existing unit can succomb to. This P.S. is an exact physical replacement and will not run hot. $27.95 + pp...VISA/MC. Kasara Inc, 36 Murray Hill Drive, Spring Valley, NY 10977. 1-800-248-2983 (Nationwide) or 914-356-3131.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC. ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

February 5-14. NORAC, the North Okanagan Radio Amateur Club, will operate a special event station daily during the Vernon Winter Carnival, Canada's largest winter carnival. Listen for VE7NOR. For a special commemorative certificate send QSL and 2 IRC's or $1.00 to NORAC, Box 1706, Vernon, BC V1T 8C3.

MICHIGAN: February 7. The 18th annual Livonia Amateur Radio Club's Swap 'n Shop, Dearborn Civic Center, Dearborn. 8 AM to 4 PM. ARRL/VEC exams given by the Motor City Radio Club. Plenty of tables, refreshments and free parking. Talk in on LARC Repeater 144.75/5.35 and 146.52 simplex. For further information SASE to Neil Coffin, WA8GWL, Livonia ARC, POB 2111, Livonia, MI 48151.

MASSACHUSETTS: February 14. Electronics Flea Market sponsored by the Algonquin ARC, Marlboro Middle School Cafeteria, Union Street off Rt 85, Marlboro. 10 AM to 2 PM. Sellers 8 AM. Admission $2.00. Tables $8/advance; $10/door. WHEELCHAIR ACCESSIBLE. For more information Dan, KB1WW, (617) 481-1587 or write AARC, Box 258, Marlboro, MA 01752.

MICHIGAN: February 13. The Cherryland ARC will hold its 15th annual Swap N Shop. Immaculate Conception Middle School gymnasium, 218 Vine Street, Traverse City. 8 AM to 1:30 PM. Admission $3.00. Tables $5.00. Talk in on 146.85 repeater. For info contact Mick Glasser, N8DBK, 4102 Peninsular Shrs Dr, Grawn, MI 49637. (616) 276-9203.

OHIO: February 14. The Mansfield Mid-Winter Hamfest/Computer Show, Richland County Fairgrounds, Mansfield. Doors open to public 7 AM. Tickets $3.00/advance; $4.00/door. Tables

(continued to page 103)

Unless the SWR is really what the meter indicates, the reading will get better or worse. You can use this method to temporarily fool the transmitter into providing power to a mismatched transmission line; just vary the line length until it loads properly. This is called *tuning the line,* and it does not change the SWR, but presents the transmitter with a load that it can handle. It is not a good idea to leave it this way, however, as things will be frequency sensitive, and the loading will change with antenna movement and weather.

Does a high SWR cause TVI? This is a trick question. High SWR by itself does not cause TVI but, its effects on other parts of the environment can. For instance if the SWR causes rf to appear on telephone wires, electrical wiring, metal roof gutters, downspouts or tv-antenna lead-in wire, you run a good chance of hearing from your neighbors. Also, the harmonic filter at the transmitter output is designed to work best at one particular impedance and when it sees a mismatch, can let harmonic energy pass through. If the final amplifier in the transmitter is overly sensitive to mismatch, it could start oscillating at some unpredictable frequency. All-in-all, a low SWR is more than just a nice meter reading — it is good operating practice!

power meter update

Not more than a week after I wrapped up last month's column which mentioned rf power meters, I found an ad for some available from a new supplier. These *REVEX* meters cover the frequency range 1.6 to 1300 MHz, power ratings of 1 watt to 5 kW, and include both average and PEP power monitoring. There's also a REVEX Wave Monitor Scope MS1 advertised that allows you to check what's being transmitted from 1.8 to 54 MHz and look at audio frequencies. Check the Amateur-Wholesale Electronics ad on page 76 of the November 1987 *ham radio.*

ham radio

flea market

$5.00/advance, $6.00/door. Talk in W8WE on 146.34/94. For tickets/tables SASE by February 4 to Dean Wrasse, KB8MG, 1094 Beal Road, Mansfield, OH 44905 or phone (419) 589-2415 after 4 PM EST.

NEW YORK: February 14. LIMARC Hamfest, Electricians Hall, 41 Pinelawn Road, Melville, Long Island. Doors open 9 to 3. Admission $4.00. 4x6 tables $12. Bring your own $1.50/ft. Advance registration only. Talk in on 146.85. For information Hank Wener, WB2ALW, 53 Sherrard St, East Hills, NY 11577 (516) 484-4322. Or Mark Nadel, NK2T (516) 796-2366.

GEORGIA: February 27. The Dalton Amateur Radio Club will hold its annual Hamfest, North GA Fairgrounds, Dalton. 9 AM to 3 PM. VE exams will be offered. Suggest reservations for exams. Contact DARCI, POB 143, Dalton, GA 30722-0143.

MINNESOTA: February 27. The 7th annual Midwinter Madness Hobby Electronics Show sponsored by the Robbinsdale ARC. Medina Ballroom, Hwy 55, Hamel. 8 AM to 2 PM. Admission $3/advance, $4/door. Flea market tables $8. 1/2 table $4. FCC exams, large indoor flea market, satellite TV and more. To register SASE with fees to Robbinsdale ARC, POB 22613, Robbinsdale, MN 55422 or call Bob (612) 533-7354.

VIRGINIA: February 28. The Vienna Wireless Society will sponsor its annual Winterfest at the Vienna Community Center, Vienna, VA. Admission $4.00. Talk in on 146.685 or 146.91. For more info contact Dave French, N4KET, 1911 Dalmation Drive, McLean, VA 22101. Tel (703) 356-0996.

INDIANA: February 28. The LaPorte ARC's Winter Hamfest, LaPorte Civic Auditorium, 50 miles SE of Chicago. Donation $3.00. Tables $3.00. Advance reservations accepted. SASE to LPARC, POB 30, LaPorte, IN 46350.

KENTUCKY: March 12. The annual Glasgow Swapfest sponsored by the Mammoth Cave ARC, Cave City Convention Center, Cave City. Starts 8 AM. Admission $3.00. Tables $3.00 each. Forums and excellent flea market. Talk in on 146.34/94 and 147.63/.03. For more information N4HCO, 1379 Whites Chapel Road, Glasgow, KY 42141.

NEW YORK: March 13. The 1988 ARRL Hudson Division Convention in conjunction with the WEÇAFEST 88 Hamfest, Westchester Community College, Valhalla. Sponsored HARC, WECA and WARY FM, the college's radio station. Activities include forums, ARRL workshops, giant flea market FCC exams and more. Admission $4.00 at the door. Talk in on 147.06, 146.91, 224.40 MHz repeaters. Exhibitor info Bob or Sarah Wilson, 2 Soundview Avenue, White Plains, NY 10606. (914) 997-8491. General info Rich Moseson, NW2L, Program Chairman, (201) 680-1585 or write Great '88, c/o NW2L, 19 Linden Avenue, Bloomfield, NJ 07003.

OPERATING EVENTS

"Things to do . . ."

February 6-8: 1988 New Hampshire QSO Party sponsored by the NH Amateur Radio Association. For information contact Pete Cantara, KI1M, 19 Haverhill St, Hudson, NH 03051.

HAM EXAMS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday February 17, 7 PM, MIT Room 1-150, 77 Mass Ave, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann at (617) 646-1641. Exam fee $4.25. Bring a copy of your current license (if any), two forms of picture ID, and a completed form 610 available from the FCC in Quincy, MA (617) 770-4023.

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Introducing the only mobiles that double as answering machines.

Now you can stay in touch—even when you're away from your radio.

With Yaesu's 2-meter FT-212RH and 70-cm FT-712RH, an optional, internal digital voice recorder serves as a convenient answering machine for you and your friends. And that's just the beginning!

High performance mobiles. The FT-212RH features wideband receive coverage of 138-174 MHz (144-148 MHz Tx), while the FT-712RH covers 430-450 MHz. An oversize amber display includes an innovative photo-sensor which increases the display brightness during the day. The function buttons are arranged in a chromatic musical scale—ideal for visually-impared operators. You get 45 watts output on 2 meters, 35 watts on 70 cm.

An autodialer DTMF microphone with 10 memories, each ready to store telephone numbers up to 22 digits long.

And, like our FT-211RH Series mobiles, you'll enjoy surprisingly simple controls, yet highly sophisticated microprocessor-based flexibility. Including 18 memories that store frequency, offset, PL tone, and PL mode (CTCSS unit optional). Band or memory scanning. Offset tuning from any memory channel. Memory channel lockout for scanning. High-low power switch.

All in an amazingly small package, shown actual size below.

Digital voice recorder option. Only Yaesu brings you the advanced technology found in our digital voice recorder option.

You can store messages or your call sign—in your own voice, not a synthesized replica—or give your friends a private code for leaving messages on your radio. All they need is a DTMF microphone! Then you can play back your messages either in-person, or remotely by using another radio with a DTMF microphone. And you've always got security because you can command your radio to respond only to in-person playback requests.

Visit your Yaesu dealer today. And test drive Yaesu's FT-212RH and FT-712RH mobiles. The only radios with the power to keep you in touch. Always.

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Yaesu USA 17210 Edwards Road, Cerritos, CA 90701 (213) 404-2700. **Repair Service:** (213) 404-4884. **Parts:** (213) 404-4847.
Yaesu Cincinnati Service Center 9070 Gold Park Drive, Hamilton, OH 45011 (513) 874-3100.

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ALL NEW!

Double Vision

ACTUAL SIZE FRONT PANEL

TM-721A

Deluxe FM dual bander

The Kenwood TM-721A re-defines the original Kenwood "Dual Bander" concept. The wide range of innovative features includes a dual channel watch function, selectable full duplex operation, 30 memory channels, extended frequency coverage, large multi-color dual digital LCD displays, programmable scanning, and more with 45 watts of output on VHF and 35 watts on UHF. TM-721A—Truly the finest full-featured FM Dual Band mobile transceiver!

- Extended receiver range (138.000-173.995 MHz) on 2 meters; 70 cm coverage is 438.000-449.995 MHz. (Specifications guaranteed on Amateur bands only. Two meter transmit range is 144-148 MHz. Modifiable for MARS/CAP. Permits required.)
- 30 multi-function memory channels. 14 memory channels and one call channel for each band store frequency, repeater offset, CTCSS, and reverse. Channels "A" and "b" establish upper and lower limits for programmable band scan. Channels "C" and "d" store transmit and receive frequencies independently for "odd splits."
- Separate frequency display for "main" and "sub-band."
- 45 Watts on 2 meters, 35 watts on 70 cm. Approx. 5 watts low power.
- Call channel function. A special memory channel for each band stores frequency, offset, and sub-tone of your favorite channel. Simply press the CALL key, and your favorite channel is selected!
- Automatic Band Change (A.B.C.) Automatically changes between main and sub-band when a signal is present.
- Dual watch function allows VHF and UHF receive simultaneously.
- CTCSS encode/decode selectable from front panel or UP/DWN keys on microphone. (Encode built-in, optional TSU-6 needed for decode.)
- Balance control and separate squelch controls for each band.
- Dual antenna ports.
- Full duplex operation.
- Programmable memory and band scanning, with memory channel lock-out and priority watch function.
- Each function key has a unique tone for positive feedback.
- Illuminated front panel controls and keys.
- Dimmer control.
- 16 key DTMF mic. included.
- Handset/remote control option (RC-10).
- Frequency (dial) lock.
- Supplied accessories: 16-key DTMF hand mic., mounting bracket, DC cable.

Complete service manuals are available for all Kenwood transceivers and most accessories. Specifications, features and prices are subject to change without notice or obligation.

TM-721A shown with optional RC-10

Optional Accessories:
- **RC-10** Multi-function handset/remote controller
- **PS-430** Power supply
- **TSU-6** CTCSS decode unit
- **SW-100B** Compact SWR/power/volt meter
- **SW-200B** Deluxe SWR/power meter
- **SWT-1** 2m antenna tuner
- **SWT-2** 70 cm antenna tuner
- **SP-40** Compact mobile speaker
- **SP-50B** Deluxe mobile speaker
- **PG-2N** DC cable
- **PG-3B** DC line noise filter
- **MC-60A, MC-80, MC-85** Base station mics.
- **MA-4000** Dual band mobile antenna (mount not supplied)
- **MB-11** Mobile bracket
- **MC-43S** UP/DWN hand mic.
- **MC-48B** 16-key DTMF hand mic.

KENWOOD

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