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FM Filter	9MI5A	9011.5	15 (-3dB)
SSB Narrow (Hygrade Crystal)	FL-44A	455	2.4
OPTIONAL FILTERS			
CW Narrow	FL-52A	455	0.500
CW Narrow	FL-53A	455	0.250
SSB Wide	FL-70	9011.5	2.8
CW Narrow	FL-32	9010.6	0.500
CW Narrow	FL-63	9010.6	0.250
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DECEMBER 1984

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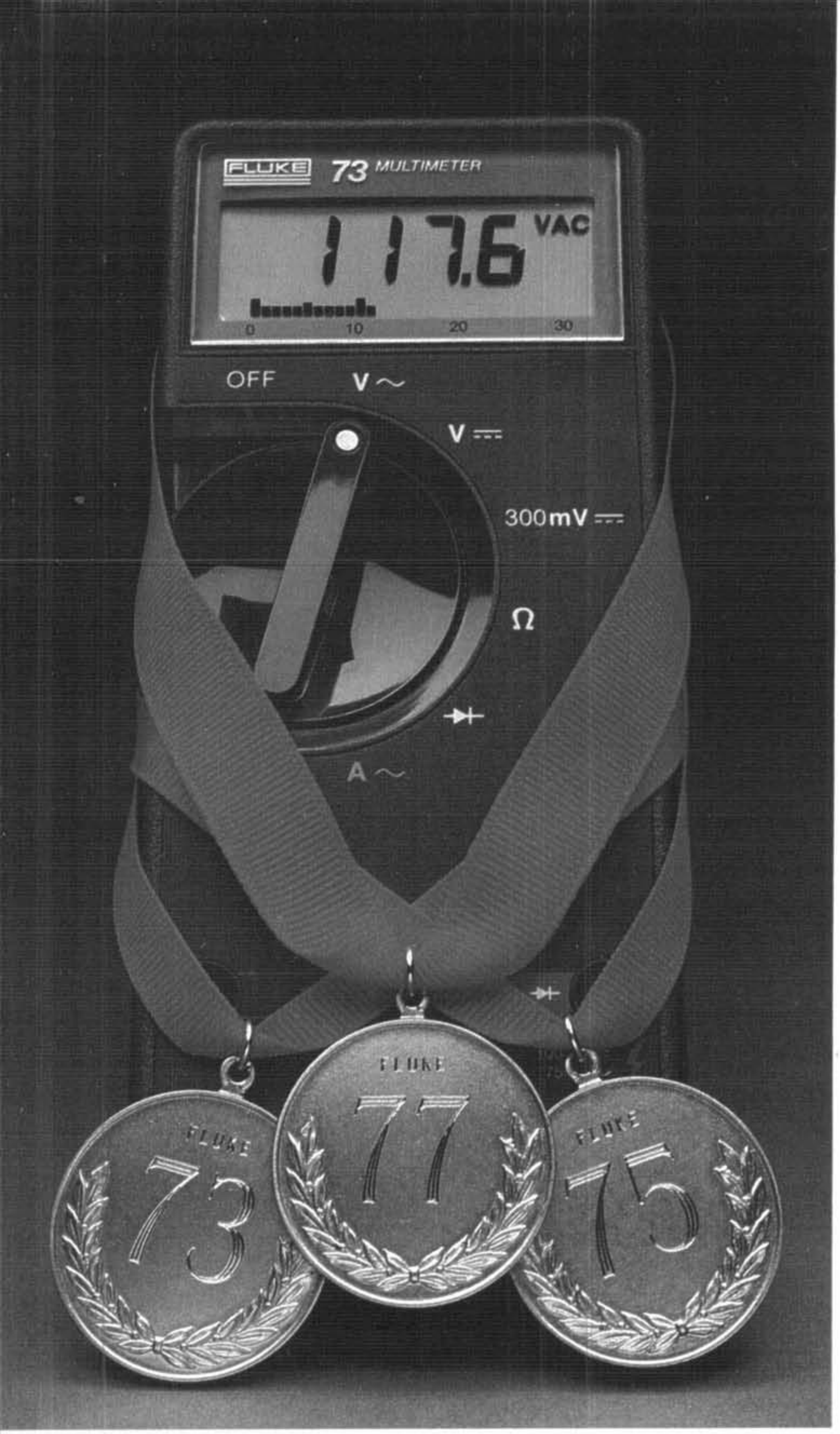

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REFLECTIONS
the good, the bad, and the ugly

There was plenty of good news in Amateur Radio in 1984. For example, we saw the beginning of the VEC program as the FCC, per its plan, extricated itself from testing. Fears were expressed that we wouldn't be able to effectively provide our own examining program. Now, however, there are 21 VEC's doing quite a good job — particularly the Volunteer Examiners, who are essential to the success of this program. And there were those who said it couldn't be done! (I feel it's appropriate to both thank and list the VEC's currently in operation. See **table 1.**)

Band expansion: it's incredible, but it's actually occurred. There was a time when I thought the HF bands would never see the increased U.S. phone allocations, but as of September 1, 1984, the 80, 15 and 10-meter bands did indeed change — for the most part, for the better . . . although it all depends on your point of view. (25, 50, and 200 kHz respectively were added to the lower portion of the phone segments of the 80, 15, and 10-meter bands, respectively.) Some Canadian Amateurs expressed disappointment in the U.S. phone band expansion on 80 meters from 3.775 down to 3.750 MHz and indicated that it might be a good idea to completely modify *their* band plan, eliminating the CW/PHONE mode subband segments. This, of course, would affect U.S. Novice operations and could cause yet another round of band changes in the U.S. As of this writing, it hasn't happened yet, though the Canadian D.O.C. is studying the proposal. (For those who operate on 80 meters — please keep the internationally regarded DX window (3.790-3.800) clear for DX contacts only.)

Some of the news was bad. Two petitions threatening a portion of the 220 MHz Amateur band were presented to the FCC by organizations (LMCC and STI) representing the land mobile industry. This situation requires close scrutiny and action by the Amateur community. (For details, see "The Endangered Spectrum: 220 MHz Under Fire," *ham radio*, October, 1984, page 6.)

One of the decisions made at WARC '79 concerned the allocation of the upper portion of the 160-meter Amateur band or more precisely, 1.9 to 2.0 MHz. In implementing the WARC agreements, the FCC has proposed to divide the 1.9 to 2.0 MHz spectrum into two parts: reserving the lower 50 kHz exclusively for non-Amateur service and assigning the upper 50 kHz to be shared by Amateurs and commercial services. Unfortunately the frequencies of 1907-1912 are used exclusively by JA hams, and this change may eliminate any chance of working JA's on 160 meters. Nevertheless, in terms of net frequency allocations gained as a result of WARC '79, the Amateur service is *still* a winner.

There was good news for those working DX. 1984 was a bumper crop year, with a number of rare countries — for example, BV, BY, and CEOX — showing up regularly on the HF bands.

One item — an ARRL "Request for Issuance of Declaratory Ruling" submitted to the FCC — has potentially vast, beneficial implications for Amateur Radio . . . but *only if hams respond* — and respond in adequate numbers. In its request, the League asked the FCC to exercise preemptive authority over local ordinances such as restrictive zoning and emissions limitations that would inhibit Amateur operations. A weak response from the Amateur community could actually cause this request to backfire. (Remember, it's very difficult to communicate on any of the bands without an antenna *or* transmitter.) Formal comments should be filed with the FCC before December 14, 1984.

In other news, Los Angeles-area jammers tried unsuccessfully to disrupt the 2-meter radio link supporting the Olympic Torch Run. Event organizers foiled the jammers by switching to private repeaters, and the run continued without interruption.

The FCC took decisive action against malicious interference, taking one jammer all the way to Federal Court, where he was convicted and placed on three years' probation. Several other investigations resulted in license suspensions.

Legislation that would make malicious interference a criminal violation (rather than an administrative violation) is pending in both the Senate and the House. Senator Barry Goldwater, K7UGA, introduced S-2975, which would make interference with any radio service a violation of federal law; Representative Jim Bates of California introduced a similar bill in the House.

It was quite a year for those who made Amateur Radio news and for those who reported it. Special thanks are due the latter: W9JUV, W5YI, Westlink, and the many other sources — including the many club newsletters that arrive regularly — that help keep us all informed.

Rich Rosen, K2RR
Editor-in-Chief

table 1. VEC's in operation as of August, 1984.

call region	organization and location
2	Metroplex Amateur Radio Association P.O. Box 237 Leonia, New Jersey 07605
	Schenectady Amateur Radio Association, Inc. P.O. Box 6 Alplaus, New York 12008
3	Laurel Amateur Radio Club, Inc. P.O. Box 3039 Laurel, Maryland 20708
4	Mid-South VEC 2020 St. Elmo Memphis, Tennessee 28213
	Central Alabama VEC 606 Tremont Street Selma, Alabama 36701
	Western Carolina Amateur Radio Society P.O. Box 1618 Ashville, North Carolina 28816
	Charlotte VEC 227 Bennett Lane Charlotte, North Carolina 28213
	Triad Emergency Amateur Radio Club 3504 Stonehurst Place High Point, North Carolina 27260
5	Dallas Amateur Radio Club P.O. Box 173 Dallas, Texas 75211
6	Greater LA Amateur Radio Group c/o Steve L. Shafit, NE6L 21921 Lanark Street #201 Canoga Park, California 91304
	SANDARC-VEC P.O. Box 5023 La Mesa, California 92041
7	Boeing Employees Amateur Radio Society P.O. Box 3707 Seattle, Washington 98124
8	Dayton Amateur Radio Association P.O. Box 44 Dayton, Ohio 45401 (Phasing out as of January 7, 1985)
9	DeVry Amateur Radio Society 3300 North Campbell Avenue Chicago, Illinois 60618
10	PHD Amateur Radio Association P.O. Box 11 Liberty, Missouri 64068
11	Anchorage Amateur Radio Club P.O. Box 101987 Anchorage, Arkansas 99510
12	(Caribbean Insular Areas) Director, Military Affiliate Radio System P.O. Box 7388 Cidra, Puerto Rico 00639
13	(Hawaii and Pacific Insular Areas) Honolulu Amateur Radio Club 3251 Pakanu Street Honolulu, Hawaii 96822
	Koolau Amateur Radio Club 45-529 Nakuluai Street Kaneohe, Hawaii 96744
All call *regions:*	The W5YI Report P.O. Box 10101 Dallas, Texas 75207
	American Radio Relay League 225 Main Street Newington, Connecticut 06111

presstop de W9JUV

ACCESS TO TWO NEW AMATEUR BANDS AND CHANGES TO SEVERAL OTHERS are the subject of a WARC implementation Notice of Proposed Rule Making released by the FCC in October. The two new bands are "12 Meters" (24.890-24.990 MHz), and "33 cm" (902-928 MHz).

For 12 Meters, The FCC Proposes Dividing The Band at 24.930; RTTY would be permitted from 24.890 to 24.930, and SSB, AM, and NBFM phone plus facsimile and SSTV from 24.930 to 24.990. CW would be permitted throughout the 100 kHz-wide band. They also propose giving General and higher class licensees access to the band, with full legal power (1500 W PEP). On 12 meters, the Amateur Service would have Primary status.

For The 902-928 MHz Band, The Amateur Service Would Be Secondary to industrial, scientific and medical (ISM) operations. The Commission proposes no subbands for 33 cm. It would be available to all classes except Novice, to all emissions, with a 1500 W PEP power limit. No Amateur operation would be permitted in Colorado, Wyoming, or in U.S. possessions in Region 3. As Secondary users, Amateurs would also have to avoid interfering with Government, Automatic Vehicle Monitor, and ISM. They'd also have to avoid Region 1 or 3 users, since it's an Amateur band only in Region 2.

10.100-10.150 MHz Would Also Become A Permanent Amateur Band as a result of this NPRM. This supercedes the Special Temporary Authorization that's currently providing access to the band, putting Amateur operation Primary in the allocations tables and raising power limits to 1500 W PEP. It would be available to General and above, but limited to CW and RTTY. Effective immediately the 10.109-10.115 MHz protection slot has been deleted, but the 200 W PEP power limit will continue in effect until final action on the NPRM.

Not All The Effects Of This NPRM Are Beneficial. The Commission has further delayed implementation of the 18.068-18.168 MHz band on the grounds that U.S. Government fixed operations on these frequencies are likely to continue until 1989. The NPRM also proposes deleting the 420-430 MHz Amateur allocation along the Canadian border to protect Canadian commercial users of that band. The deleted area is a band extending 75 miles into the U.S. and Alaska from the border, and within this area Amateur operation would be permitted only with a waiver from the FCC and coordination of the Canadian Department of Communications.

Comments On The NPRM, PR Docket 84-960, Are Due at the FCC December 17; Reply Comments will be due January 16, 1985. An original and five copies are required to make a formal submission; six additional copies are required if you wish a copy to go to each of the Commissioners. Send them to The Secretary, Federal Communications Commission, Washington, D.C. 20554. As this docket covers a number of bands and issues, comments are expected to be heavy. As a result, action is unlikely until late next spring.

ALL MODES FOR 160 METER USERS HAS BEEN PROPOSED in yet another FCC NPRM released in October. Amateurs responding to ARRL's earlier petition to permit RTTY on top band had suggested adding SSTV and facsimile as well, and the Commission agreed. However, the radiolocation threat to 160's top half (November Presstop) remains a very real one.

Comments On The 160 Mode Relaxation Are Due at the Commission December 20, with Reply Comments due January 22, 1985. Refer to PR Docket 84-959.

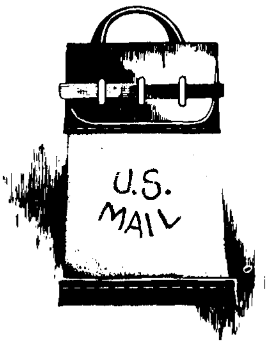
ARRL'S PETITION TO HAVE FCC PREEMPT LOCAL REGULATION OF ANTENNAS could trigger a problem for Amateurs. Amateur response thus far to the League's request, which was designated PRB-1 by the Commission, has been sparse, but it has generated a number of comments from municipalities and municipal organizations. Their response has been, understandably, that they need more—not less—control over what goes on within their borders. As a result they'd like the FCC to not only deny the League's request, but even consider giving some formal sanction to local ordinances limiting antennas and possibly even the operation of Amateur (and other) transmitters within their jurisdictions!

Such A Turnabout Would Be A Severe Blow To Amateur Radio, and strong Amateur support for ARRL's PRB-1 is needed to head it off. Though the Comment period on PRB-1 will close before this reaches print, Reply Comments (opposing increased local control) can still be filed through December 14. Send the original and four copies to the FCC Secretary, above.

THE U.S. AMATEUR POPULATION HAS DROPPED SLIGHTLY DURING 1984, from a total of 411,479 individual current licenses in effect at the beginning of the year to 409,923 at the beginning of October. Though there was a slight peak, to just over 412,000, in the spring, the subsequent decline has been at a steady rate. Though most license classes held their own or even increased slightly during the year the most dramatic change was in the Novice ranks, which dropped more than 5000 (from 85,823 to 80,461) during the nine-month period. Though many of these were upgrades, and others early recipients of the five year Novice license who lost interest, it's disturbing more new Novice licensees have not been brought in during the period.

Confirming This Concern, An Internal FCC Analysis of 1984 license statistics shows that while 18,800 people were joining the Amateur ranks, 19,964 dropped out. In addition, while 8,829 Novices upgraded during the period, 14,484 Novices let their licenses expire!

Since The Novice Exam Procedure Had Not Changed when the Volunteer Exam Program went into effect, that should not be a reason for the decline. The FCC, understandably concerned, is considering how license structure changes might influence this trend.



comments

SSB development

Dear HR:

"The Development of Amateur SSB" by John Nagle, K4KJ, (September, 1984, page 12) did a very good job of covering the history of SSB for the present day Radio Amateur.

Two names that should go down in any historical account — Don Norgaard, ex W2KUJ, and Wes Schum, W9DYV — were not mentioned. In 1948 Don wrote a series of articles on SSB for *QST*, using Bob Dome's (W2WAM) ideas to develop a phasing type of SSB transmission and reception. Then Wes came along with his Central Electronics 10A units and kits to make the nuts and bolts of phasing SSB easily available where filters were either difficult or impossible to obtain. As one of the first two hundred hams on the air on SSB, I found them a great help.

Some of those Western Electric HF SSB transmitters John mentioned as having been put into operation in the early 1930s were still in overseas service, though a little patched up, some 40 years later — when they were sold for scrap!

Wayne W. Cooper, AG4R, ex YN1WC, HR2WC
Miami Shores, Florida

loop antenna

Dear HR:

I think I can clear up any questions about the loop antenna (see WD6DUD letter, September, 1984, page 8) mentioned by Bill Orr in his February, 1984, column (see page 63). I've just returned from Germany with the original article that appeared in the May, 1983, issue of *CQ-DL*, the German ham magazine.

As you can see (**fig. 1**), there are actually three different methods of coupling to this antenna. C2 in Bill's diagram (**fig. 2**) is unnecessary as the feedpoint reactance can be tuned out with the loop tuning cap C1. According to diagram 13.3 of the original (**fig. 1C**), the tap should be made at 1/5 loop diameter with a spacing of 1/200 wavelength.

The length given in Bill's article is incorrectly shown in the diagram (**fig. 2**). The correct length is 0.2 wavelength at the *highest* frequency. The lengths given in the table (February, page 65) are OK.

The author states that the SWR is below 1.5 to 1 over an octave frequency range, but this will require motor tuning of the loop tuning cap because the bandwidth at the low frequency end of the range is very narrow (6 kHz on 80 meters). As for the groundplane dimensions, he recommends that it extend out at least twice the loop diameter, but further if possible. As with any small loop, the radiation resistance is quite low and ground losses can eat up a lot of signal if not minimized. This is also true of conductor losses, making the use of copper tubing a necessity. I would recommend 1/2-inch tubing because this size is readily available at plumbing and heating stores and will have less loss than the 3/8 size mentioned. Don't try to save money by eliminating the ground return line for the same reason.

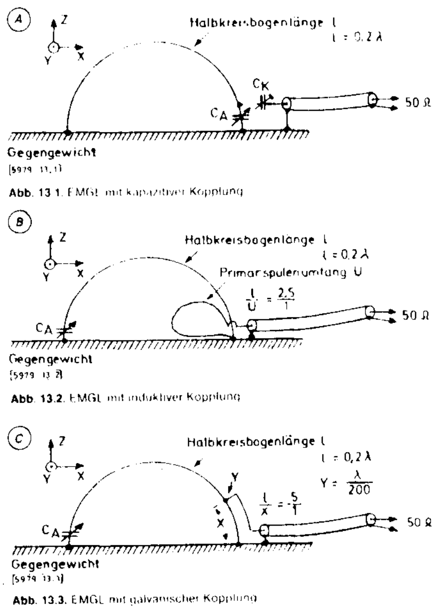


fig. 1. *CQ-DL* illustrations show three different methods of coupling.

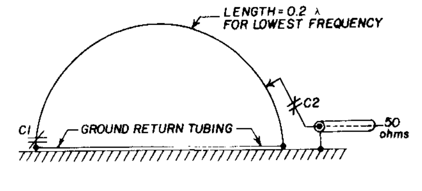


fig. 2. DL2FA groundplane loop as shown in W6SAI's February column.

Roger D. Johnson, AD1G
Caribou, Maine

spectrum history

In the interest of brevity, several lines of Joe Schroeder's (W9JUV) segment of the October editorial ("The Endangered Spectrum: 220 MHz Under Fire," page 6) were omitted. They are reproduced here in the interest of historical accuracy — **Editor.**

When the U.S. entered WW II the 1-1/4-meter band was just that: 224-230 MHz, a perfect harmonic of 2-1/2 meters (112-116 MHz), 5 meters (56-60 MHz) and so on down. A few years earlier, however, the entire spectrum above 110 MHz had been designated simply "Amateur and Experimental."

When the war ended and Amateur operations resumed, the 1-1/4 meter band was temporarily shifted to 235-240 MHz (probably to avoid some of the early radars still active between 180 and 220 MHz). Then, in the late 1940s, it was finally shifted to the present 220-225 MHz slot.

short circuit

Guerri Report

In "The Guerri Report," (November, 1984, page 54), the line that reads, "A complete SSB receiver could actually consist of 3000 individual receivers, each having 1 kHz bandwidth . . ." should be corrected to read "1 Hz." This suggests the possibility of 3000 discrete 1-Hz spectrum segments comprising a single voice bandwidth channel, similar to the technique utilized in spread spectrum applications.

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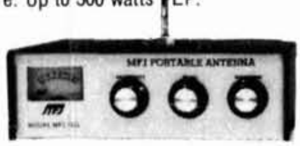

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TS-711A/TS-811A

Multi-function all-mode 2 m and 70 cm transceivers.

The TS-711A 2 m (142-149 MHz) and TS-811A 70 cm (430-450 MHz) all-mode transceivers are perfect base station units, designed to complement your present HF station. Both feature Kenwood's innovative D.C.S. circuitry. Built-in dual digital VFO's provide commercial-grade frequency stability through the use of a TCXO (Temperature Compensated Crystal Oscillator). The new fluorescent multi-function display shows frequency, RIT shift, VFO A/B, SPLIT, ALERT, repeater offset, digital code, call sign code, and memory channel. 40 multi-function memories store frequency, mode, repeater offset and tone. They have programmable scan, memory scan, and mode scan. The Auto-mode function automatically selects the correct mode for the frequency being used. When a mode key is depressed, an audible "beeper" announces mode identification in International Morse Code.

The TS-711A/TS-811A also feature all-mode squelch, noise blanker, speech processor (SSB, FM), IF shift, RF power control, alert, and a unique channel Quick-Step tuning that varies tuning characteristics from conventional VFO feel, to stepping action when CH.Q switch is depressed.

Combine all these features with built-in AC power supply and a hefty 25 watts RF output power and you have your ideal base station.

Optional accessories:

• CD-10 Call sign Display • TU-5 CTCSS Tone Unit • VS-1 Voice Synthesizer • MC-60A Deluxe Desk Mic • MC-80 Desk Mic • MC-85 Desk Mic • SP-430 External Speakers • MB-430 Mobile Mount • PG-2J DC Cable

TS-670

TS-670 All-mode "Quad Bander."

The TS-670 "Quad Bander" is a unique all-mode transceiver that covers the 6 meter VHF band and the 10, 15 and 40 meter HF bands. FM operation may be added with the optional FM-430. Key features include dual digital VFO's, 80 memory channels, memory scan, and programmable band scan. Direct keyboard frequency selection allows you to enter a frequency to either VFO or to a memory channel using the 10-button key-pad on the front panel. The 2-color fluorescent tube display indicates frequency to the nearest 100 Hz (10 Hz modifiable) and includes LED indicators that signal the specific functions in use. The optional GC-10 general coverage receiver unit allows continuous tuning from 500 kHz to 30 MHz. The VS-1 voice synthesizer unit is another popular option available. All this plus IF shift, all-mode squelch, CW semi-break-in with side tone, narrow-wide filter selection, noise blanker, and R.F. attenuator make the TS-670 "Quad Bander" the next transceiver you should own!

Optional accessories:

• GC-10 General Coverage Unit, 500 kHz to 30 MHz • VS-1 Voice Synthesizer • FM-430 FM Unit • YK-88C 500 Hz CW Filter • YK-88CN 270 Hz CW Filter • YK-88A 6 kHz AM Filter • PS-430 DC Power Supply • KPS-7A DC Power Supply • MC-60A Deluxe Desk Mic • MC-80 Desk Mic • MC-85 Multi-Function Desk Mic • VOX-4 VOX Unit

More information on the TS-711A/TS-811A and TS-670 is available from authorized dealers of Trio-Kenwood Communications, 1111 West Walnut St., Compton, CA 90220.

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complete solar power
for your ham station

Clever parts selection and careful siting deliver economical alternative power

We've all noticed, with increasing alarm, the steep rise in our monthly electric bills. Here in Hawaii, the rates have passed 13 cents per kilowatt hour, and show signs of continuing upward. Being of a prudent nature, I decided to make the effort to reduce consumption by all practical means. Switching to roof-mounted solar collectors for hot water heating was first. Converting my station to solar power was next.

Photovoltaics (PV) offer an attractive solution to the high cost of electric power if one can afford the initial capital investment. Friends in this industry claim that suitable panels shouldn't cost more than $8 to $10 per watt; this price, however, is based on a quantity order. I figured that a single panel of approximately 30 watts would be adequate for my purposes. The only problem would be to find a vendor who would provide a suitable panel for approximately $300.*

After contacting all the sources I knew, I found that such a unit might normally retail for as much as $500, more than I wanted to spend. Luckily, a California manufacturer agreed to provide one unit for $350. This particular panel (SOLEC No. 4136) provides about 2.5 amperes at 18 volts in full noon sun.

*For a list of manufacturers, send an SASE to *ham radio*, Greenville, N. H. 03048.

The next problem was to find a suitable 12-volt battery. I wanted one that would be able to stand the heavy transmitter load without balking, and could do so for a long time. A Sears Roebuck Deep Cycle Marine Battery ($76.00) was selected.

Because the PV panel generates more voltage than the battery needs for charging, a regulator is necessary. While the panel puts out 18 volts in full sun, as noted previously, it generates less power as the sun moves lower in the sky, thus making a "lossless" regulator desirable. (Series regulators usually lose a volt or more; this loss cannot be tolerated in this application.) An excellent on/off type regulator described by Millard[1] was selected. This circuit senses when the battery is fully charged and switches off the PV input. When the battery level drops, the regulator switches the input from the PV back on.

As depicted in **fig. 1**, extensive metering is used throughout. The PV panel, regulator, and battery output are protected by circuit breakers. While low voltage DC breakers are not easy to find, a variety of surplus units are available at a reasonable price (see parts list, **fig. 1**).

A series diode is used to isolate various elements of the total unit. We are all aware of the usual approximate 0.7 volt drop across the average silicon diode. As noted previously, we are scrambling for even fractions of a volt if the PV panel output is to be used effectively. Conventional diodes — just as conventional series regulators — have excessive losses, and therefore should not be used. Power Schottky diodes with their approximate 0.2 volt drop are preferred (see **fig. 2**).

Because a charging battery generates hydrogen gas, which can easily explode, the battery should be installed outside the house in a location in which air circulation is unimpaired. Of course, the further it is from

by William L. Schreiber, NH6N, 73-4327 Imo Street, Kailua-Kona, Hawaii 96740

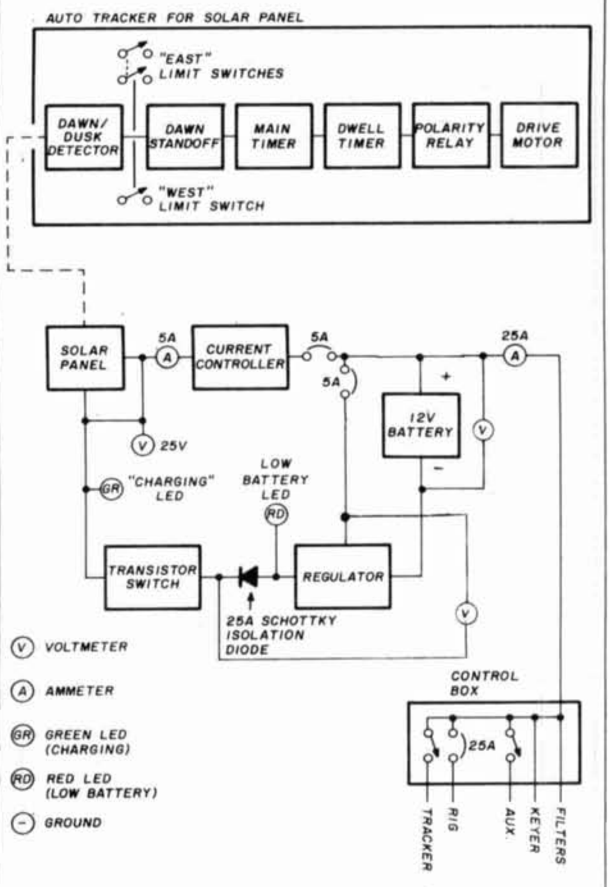
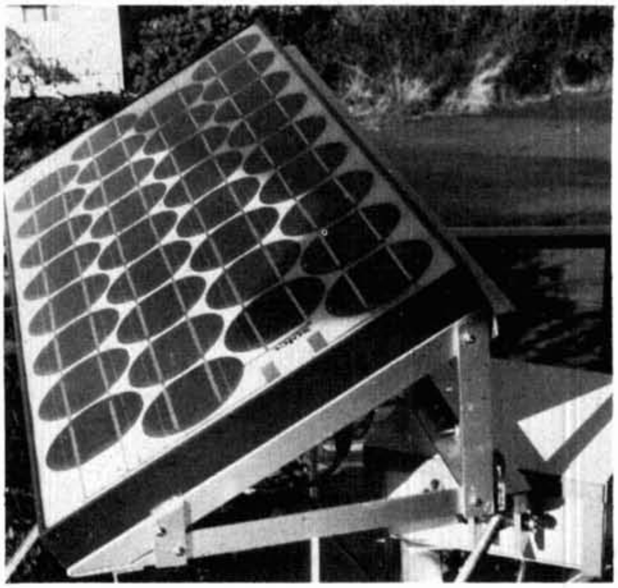


fig. 1. Block diagram of the solar power unit.

Parts suppliers

solar panel: 2.5 amp/18 volt	SOLEC International, 12533 Chadron Avenue, Hawthorne, California 90250
battery: 12 volt deep cycle Marine battery Catalog No. 28H9652N	Sears Roebuck
breakers: surplus low-voltage DC breakers	Fair Radio Sales, Lima, Ohio 45802
Schottky diode: SK 9151	Quement Electronics, 1000 Bascomb Avenue, San Jose, California 95128
bearings	Edmund Scientific, 101 East Gloucester Pike, Barrington, New Jersey 08007

the rig and the PV panel, the more critical the choice of cable becomes. I chose No. 10 copper wire which also proved to be quite expensive — because of its



Top front view of PV panel and drive control box.

tolerable IR loss. On a per-foot basis, a set of automobile jumper cables (copper) costs less than a third the price of regular wire. So I decided to use these between the battery, the control box, and the rig instead. But my PV panel is about 25 feet (7.62 m) from the battery — much longer than any jumper cables I could find. So I found, in my parts stock, a length of coaxial cable, similar to RG-8/U, that proved to have a No. 10 inner conductor. I used this for the run from the battery to the panel.

mounting the panel

Mounting the PV panel proved to be a problem. It was obviously important to be able to rotate it and even change its elevation if necessary. After perusing EPA documents on solar hot water panels, I decided that a good compromise for the elevation angle was to make it about 10 degrees plus local latitude. My location in the southern part of the Hawaiian Islands is about 18 degrees North latitude. So a 28 degree fixed elevation was selected. I built a frame out of 1 inch (2.54 cm) aluminum angle stock, and bolted it together with stainless steel hardware.

My first attempt to rotate the assembly involved mounting it on a wide-flange bicycle wheel hub. I bent a 1 × 1/4 inch (2.54 × 0.64 cm) sheet of flat aluminum stock to resemble a bicycle fork, and the PV frame was bolted to it. The thing worked, but it was an awkward arrangement and didn't seem sturdy enough to survive our occasional tropical storms.

I finally found "lazy Susan" bearing assemblies in various sizes. A 12 inch (30.5 cm) and a 6 inch (15.2 cm) bearing were secured. The larger one was bolted to two 12 × 12 inch (30.5 × 30.5 cm) 16-gauge aluminum sheets, with the PV frame bolted to the top sheet. A low-friction bearing resulted, and the whole assembly could be rotated with light pressure (**fig. 3**).

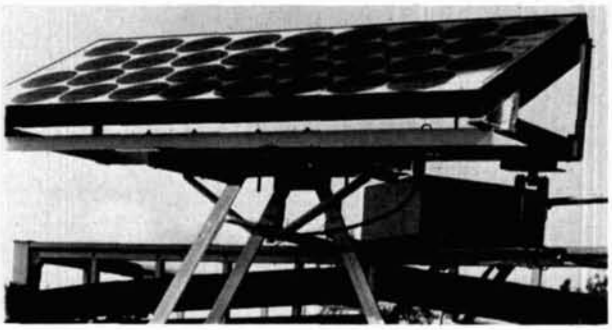
Total rotation angle is an additional consideration. Because very little current was developed by the PV panel for the first hour or so after sunrise, I decided to limit rotation between Southeast and West for approximately 135 degrees of rotation.

With these successes in hand, it became obvious that a motor-driven tracker should be added so that the panel would always face the sun. Two basic choices were considered. The first was a true tracker based on two phototransistors. It was breadboarded and found to work, but it had two basic flaws. A ±12 volt supply at 2 amperes was required. The battery would supply the +12 volts quite easily, but the negative voltage was not so easily provided. Since my goal was to be able to depend totally on the battery, no outside voltage could be considered. It's possible to generate −12 volts from the +12 volt battery via a power op amp inverter, but this seemed to be doing the job the hard way. In addition, an auto tracker gets into trouble when the sun goes behind the clouds for an hour or so. In order for the panel to find and face the sun, an "acquisition" system must be included in the tracker design. This was far too complicated and was summarily rejected.

The second choice was a timer-controlled system (see **fig. 4**). (Some circuits in Walt Jung's *IC Timer Cookbook*[2] are suitable.) My approach was to use an astable to handle the longer times — 40 to 60 minutes — and use its high output to trigger a monostable, which would have an adjustable duration of 0 to 30 seconds. This would turn the motor on via a relay.

For convenience, the timing resistors for the astable — eight 10-megohm 1/4 watt — were wired across a DIP switch. This allowed up to eight resistors to be placed in series. The timing capacitor was composed of five each 2.2 μF mylar capacitors in parallel.

At this point, it is important to consider the various pecularities of IC timers. My old standby, the 555/56, has serious limitations in applications in which time



PV panel in test position.

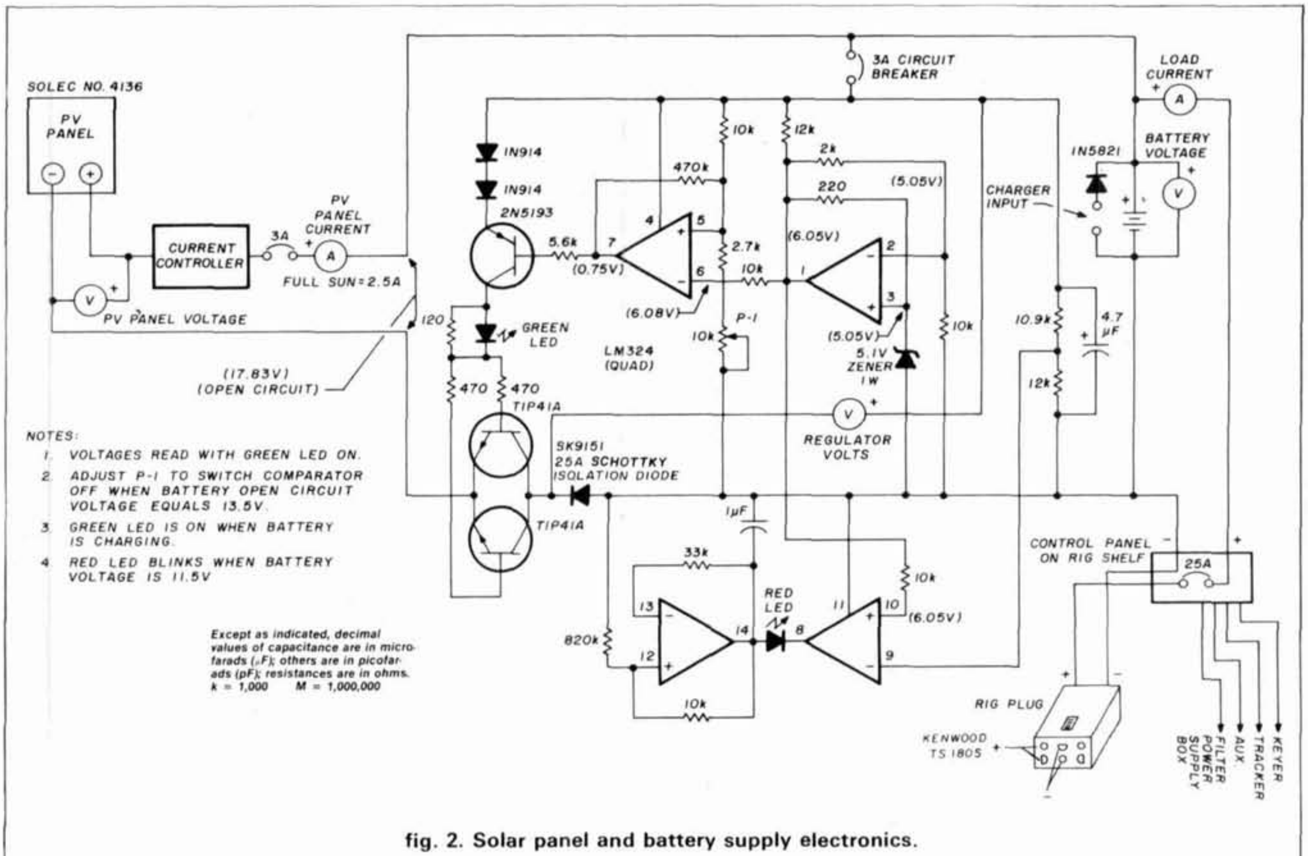


fig. 2. Solar panel and battery supply electronics.

When the FCC changed the rules, EIMAC was prepared for continuing HAM operations.

The FCC changed the allowable output power for linear amplifiers in amateur radio service. Hams can now run at 1500 watts PEP into an antenna. EIMAC was right there to meet requirements with its 3CX1200A7 tube.

Low-cost replacement for small spaces.

RF cabinets of many linear amplifiers currently use the EIMAC 3-500-Z tubes. The new 3CX1200A7 for design takes size into consideration and, by design, is recommended as a single, low-cost replacement for a pair of EIMAC 3-500-Z tubes for new amplifier designs.

General Specifications

The EIMAC 3CX1200A7 is a high-mu, compact, forced air cooled triode for zero-bias class AB2 amplifiers.

- 2.9″ dia. x 6.0″ long
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ICOM's three ultra compact mobiles...the IC-27A 2-meter, the IC-37A 220MHz and the IC-47A 440MHz... are the smallest mobiles available.

Even in such a small package the 25 watt mobiles contain an internal speaker which makes them fully self-contained and easy to mount.

Size. The ICOM compacts measure only 5½"W x 1½"H x 7"D (IC-37A is 9" deep)... which allows them to be mounted in various "compact" locations. Yet the compacts have large operating knobs which are easy to use in the mobile environment.

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9 Memories. The compact mobiles have 9 memories which will store the receive frequency, transmit offset, offset direction and PL tone. All memories are backed up with a lithium battery.

Speech Synthesizer. To verbally announce the receive frequency, an optional UT-16 voice synthesizer is available.

Scanning. The ICOM compacts have four scanning systems...memory scan, band scan, program scan and priority scan. Priority may be a memory or a VFO channel...and the scanning speed is adjustable.

Stacking Mobile Mounts. The IC-27A/37A/47A can be stacked to provide a three band mobile station. Each band is full featured and will operate even when another band is in use.

The IC-27A/37A/47A provide superb performance in the mobile radio environment. See them at your local ICOM dealer.

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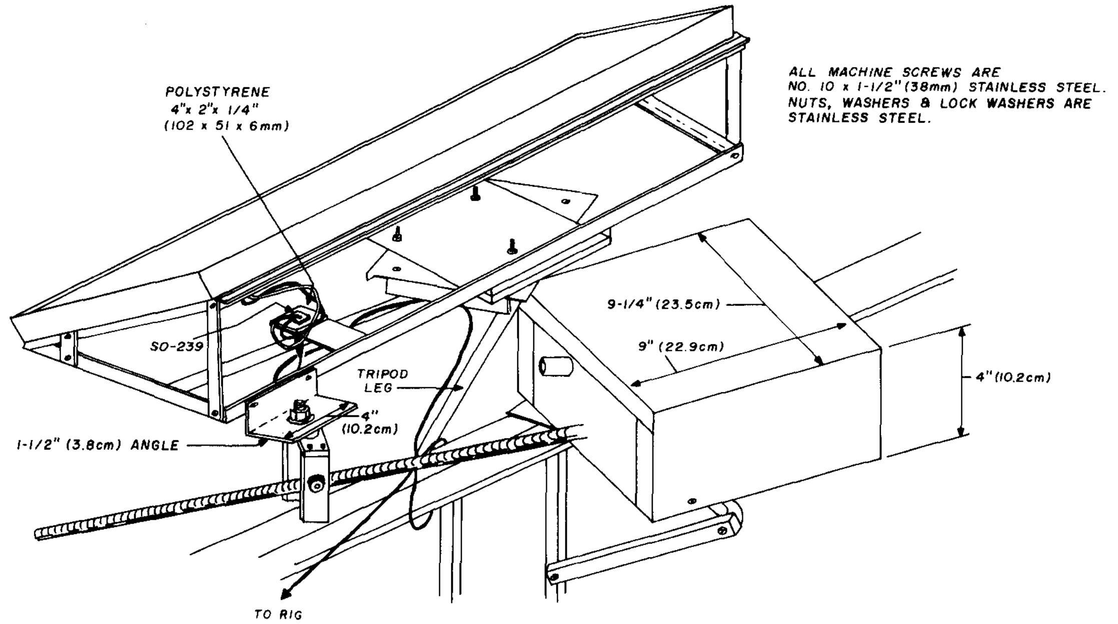


fig. 3. Low friction bearings ease rotation of the solar panel.

delays of long duration are desired. Also, the capacitors have to be carefully chosen to avoid leakage. Aluminum electrolytics are unsatisfactory, and even tantalums have to be used with caution. The low voltage film types (mylar, polystyrene, and polypropylene) are the best choice. If necessary, a quality tantalum may be used.

My first attempt at a 30-minute astable with a 555 proved to be very unpredictable. One of the problems was leakage within the IC. The other had to do with the inability of the 555 to time out with a resistor much larger than 15 megohms. Jung points this out in the *Cookbook* and suggests using the CMOS version of the 555/56 — the 755/56 made by Intel. This very low-leakage device easily handled a 100-megohm network, and was chosen for the astable.

When this was done, I noticed that after about 40 megohms were switched in, the circuit "timed out" the same even up to 100 megohms. The timing was erratic also. I learned a lesson here: the perf board was liberally splattered with rosin and fine solder splashes that are normally no problem. But throw in a lot of megohms, and the leakage both here and across the switch contacts will interfere with the timing. The solution was to clean the board with a flux solvent and get rid of the DIP switch. The bank of resistors now hangs in the air directly from the IC socket pins.

The monostable had no such problems because of the short timer delays involved. In the interest of simplicity, a dual IC — 7556 — was used for both purposes.

A practical problem incident to the 7556 is its low current sourcing capability — approximately 50 mA. I was unable to buy a multipole relay with coil currents below 80 mA here in Hawaii, so I added the Darlington power transistor as a relay driver.

dawn/dusk circuit

In the overall system concept, it was obvious that the tracker should stop when it finished its rotation toward the west. A limit switch did the job when an adjustable stop came in contact with the switch arm. Suitably connected, this turned off the motor. A means had to be provided to recognize this fact as evening approached, and the switching system allowed the PV panel to rotate back, face East, and then stop. The stopping was accomplished by two additional limit switches and an adjustable stop. At dawn, the limit switches had to be bypassed so that the timers could start the intermittent rotation.

The circuit chosen was based on the large change in resistance by a cadmium sulfide cell. Mine has a "dark" resistance of about 100K. Its "light" resistance was about 50 ohms. This characteristic was used in a simple comparator circuit. Again, the 741 op amp did not handle the relay coil current, and a Darlington power transistor was used as a booster. Should you find that the circuit "hunts" as it gets dark, it may be necessary to add hysteresis so that the op amp will have a "dead zone." This can be done by providing a very small amount of positive feedback by connecting a 6.8 Megohm 1/4-watt resistor on the comparator

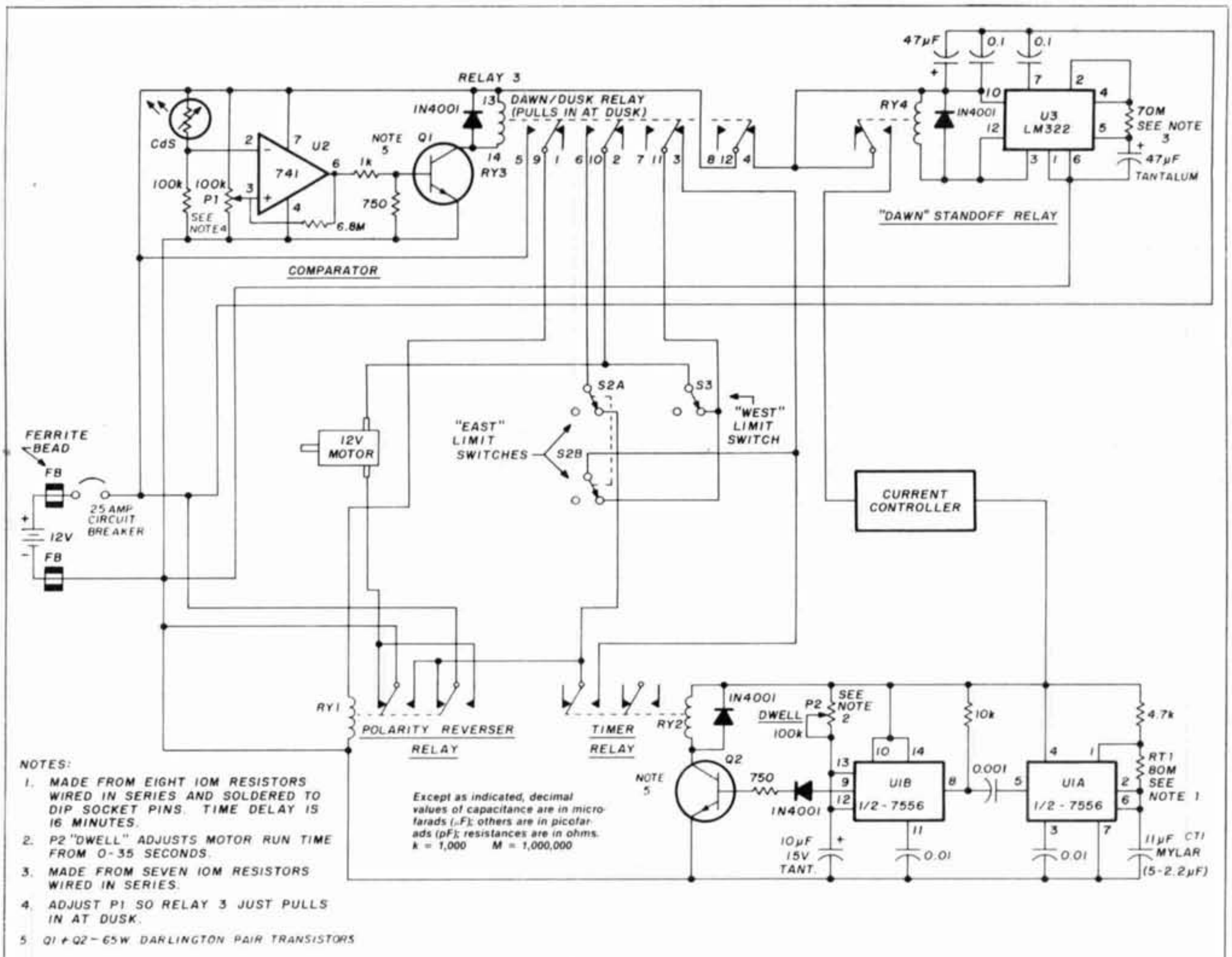
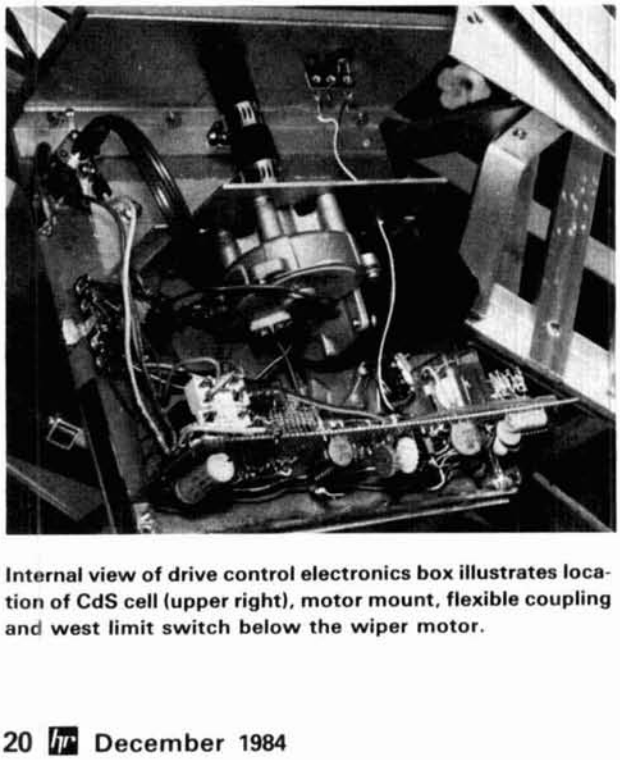


fig. 4. Autotracker circuits account for rotational limits, panel position reset and "dawn standoff" compensation for a low output condition at sunrise.



Internal view of drive control electronics box illustrates location of CdS cell (upper right), motor mount, flexible coupling and west limit switch below the wiper motor.

op amp output to its non-inverting input (pin 1 to pin 3). Another factor to consider is to match the dark resistance of the CdS cell with its series resistor. In my case, it's 100K.

dawn standoff

A number of tests revealed that the PV panel had very little output early in the morning, and reached its average output about an hour after sunrise. Because the "dawn" detector triggers at sunrise, the PV panel tends to begin rotating too early for optimum sun tracking. Adding the "dawn standoff" timer handled the situation nicely. Briefly, it uses a National LM-322 CMOS IC as a long-duration timer in a one-shot, power-up circuit.[3] The particular R_t C_t (80 Megohms, 47 microfarads) used here gives a delay of about 65 minutes. Triggered once by the "dawn" circuit, it times out, goes high, and actuates the relay that supplies power to the main timer. The delay can be adjusted over a wide range by increasing or decreasing

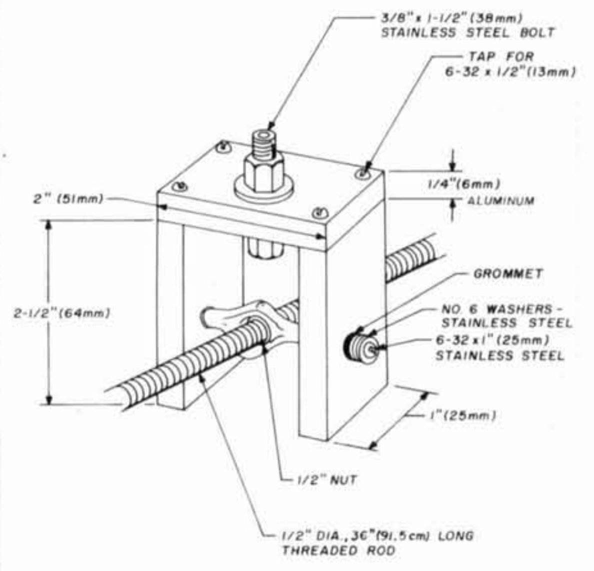
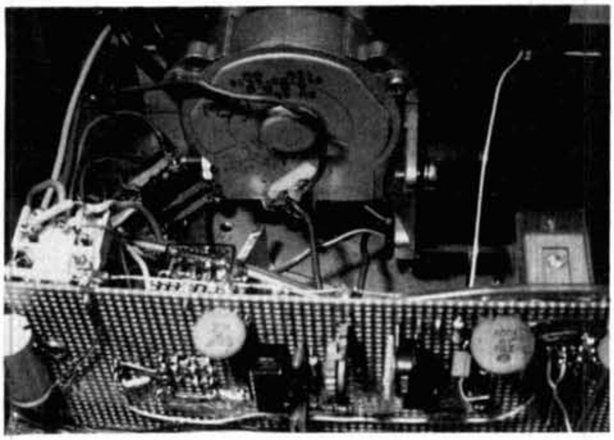


fig. 5. Mechanical details of the lead screw system.



Close-in view of drive control box electronics shows east limit switches below gear head.

C_t and/or R_t. As noted previously, the series resistor bank hangs in the air, and is soldered directly to the IC socket pins.

mechanical considerations

As can be seen in photos A and B, the electronics/drive motor box is offset to the rear on the "east" side of the PV panel. The exact location of this unit has to be determined by trial and error, and is based on the desired total rotation angle, as well as mechanical support practicalities. The center of my box is 13 inches (33 cm) from the center of the PV panel and is offset 7-1/2 inches (19 cm) to the side.

My choice of a lead screw system was limited to an available 3 foot × 1/2 inch (91.5 × 1.27 cm) (NC) steel drive rod (see **fig. 5**) which was the only threaded rod I could find in my local hardware store. An aluminum rod of the same dimension would have been preferable.

The only outside mechanical work needed was to weld a pair of threaded steel spacers (8/32 × 1 inch) (6.4 × 25.4 cm) to a steel 1/2 inch (1.27 cm) (NC) nut. A U-shaped bracket was made of 1 × 1/4 inch (25.4 × 6.4 mm) flat aluminum stock into which the nut and trunnions were fitted. The bracket was drilled at the top for a 3/8 × 1-1/2 inch (9.5 × 38 mm) stainless steel screw and nut that permitted it to be attached to an aluminum angle bracket. This in turn was attached to an extension in the PV panel. The whole arrangement resulted in a universal joint tolerant of my sloppy mechanical alignment.

Rotation angle to the west is limited by the threaded rod's contact with the PV panel structure. To the east, the system will not work beyond the point at which the rod comes into line with the center of the panel. This allows about 135 degrees of rotation. If more is required, a longer rod is needed, and the box would have to be moved out a greater distance from the PV panel.

I needed a reasonably powerful, geared down, 12 VDC motor that was both cheap and very small. Knowing that cars use all sorts of small electric motors, I went to the local wrecker and bought a nearly new 2-speed Mazda windshield wiper motor for about $3.00. It checked out at a slow speed of approximately 120 RPM — too fast for my purposes — but a lead screw system with a speed of much less than 1 RPM slowed the panel rotation down. I'll admit that it's a clumsy looking thing, but it works well and is very rugged. If you can find a small, powerful 1 RPM motor, you can probably drive the PV panel directly with a short vertical shaft and thus simplify things considerably.

motor

Since it's not possible for me to anticipate which particular motor you may use, there's no way I can indicate specific connections. In any case, the leads should be paired out with an ohmmeter. I used my bench power supply to try out my choices. A lower than normal voltage was used, and the built-in current limiter provided added protection. In any event, proceed cautiously, and determine which two leads coming from the motor (mine had six) are the proper ones.

electronics/drive box

This is made of double-sided circuit board soldered together for the top and sides. The front and bottom are 16 gauge aluminum. (The latter could have been made of the circuit board material just as well.) A small

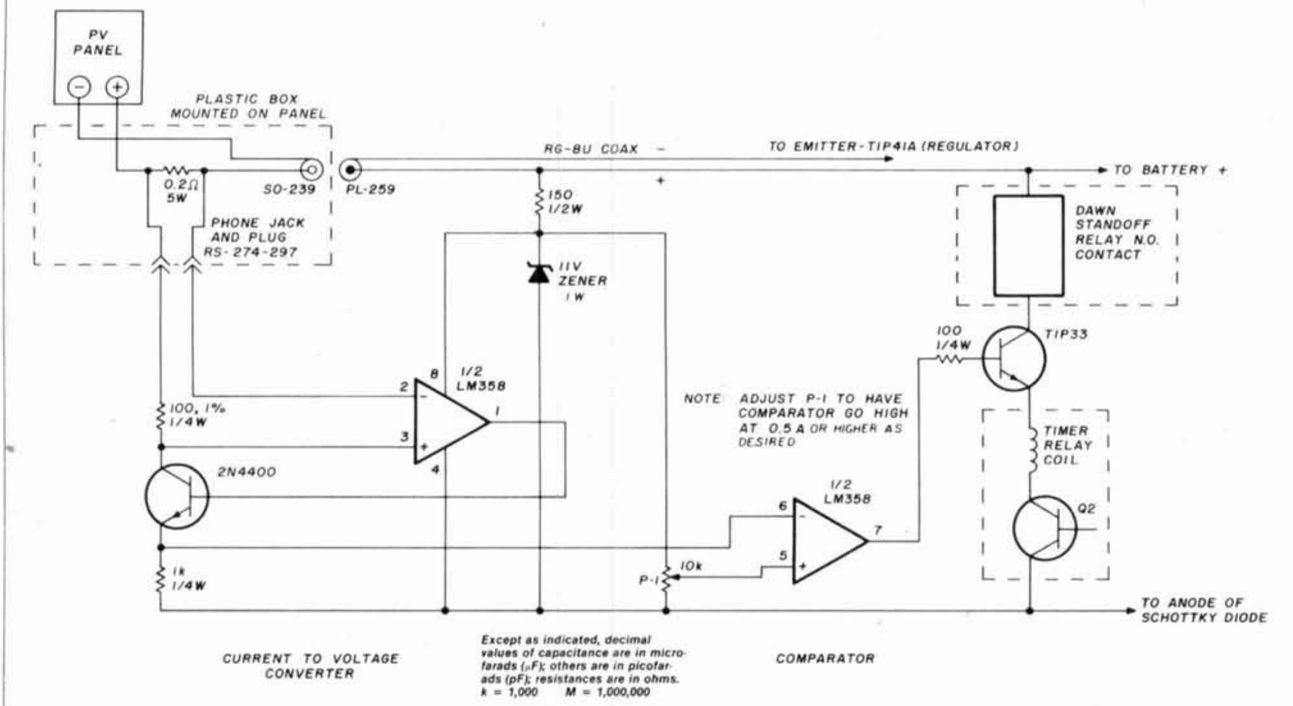
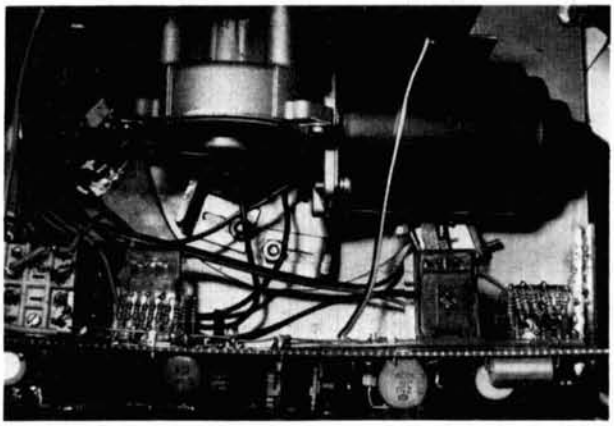


fig. 6. Current controller recognizes when a suitable charging level is achieved and stops panel rotation until the level drops.

"lazy Susan" bearing attached to the bottom panel allows rotation of the box as the panel swings.

The windshield wiper motor is mounted on an aluminum angle bracket and attached to the threaded rod with a piece of rubber hydraulic tubing that fits the motor shaft and rod tightly. The rubber tube is held in place with two hose clamps. This results in a flexible coupling which, together with the universally mounted nut and the rotating box, is very forgiving of inaccurate mechanical alignments (see photo).

A perf board is used to mount the relays, ICs, and assorted components for the timing circuits. The lead sulfide cell is mounted in a rubber grommet on the front of the box, where it has a good "view" of the ambient light.



Edge view of control box details limit switch stops, resistor banks ("in the air") and drive motor.

mounting

Mount the PV panel where it has an unimpeded "view" of the sky between southeast and west. Mine is mounted on a 10 foot × 3 inch (305 × 7.6 cm) plastic sewer pipe. Its base is set in concreted rocks, and the top has a flange that supports the 3/4 inch (1.9 cm) plywood base. Because the pipe is somewhat limber, I used guy wires to stabilize its position.

adjustments

A number of adjustments are necessary before the system is fully functional. Mechanical settings of the box position and rotation angle have already been mentioned. Electronics items needing adjustment include the following:

Voltage regulator. The comparator pot should be set to have the green LED turn off at a battery voltage of 13.5 volts. The ideal way to do this is with a metered power supply substituted for the battery wherein the supply is set to 13.5 volts and the pot turned until the LED goes out. Alternatively, the battery can be used with an accurate voltmeter. Once the battery reaches the desired voltage, this adjustment can be made. Of course, you'll have to wait for the battery to charge up to this voltage, which may take some time.

Main timer (first half of the 7556). Depending on how long a time delay is desired, both the timing capacitors and resistor bank can be added to or subtracted from. More resistance and/or capacitance increases the delay.

Dwell (second half of the 7556). The pot allows an adjustment of 0 to 30 seconds. In my case, the main timer times out at about 12 minutes, and the dwell is set for about 6 seconds. This combination is reasonably close to keeping the PV panel pointed at the sun ($\pm$15 degrees).

Dawn/dusk relay. The comparator's pot permits a wide range of light levels to be selected. Mine is set so that "dusk" is recognized just at sunset.

Dawn standoff. Again the RC combination controls this timer. Mine times out after approximately 1 hour after sunrise, which seems to be the point at which the PV panel produces a worthwhile output. If you want a shorter period, then decrease the resistor or the capacitor value.

Limit switches. Because it wasn't possible to find a double-pole double-throw micro switch for the "east" limit, two SPDT units were bolted together with 2-56 screws 1-1/2 inches (3.8 cm) long (photo D). The "west" limit required only one SPDT switch. Location of the stops for these switches took some careful planning to avoid mechanical interference (photo E). The whole assembly was rotated to the desired positions, and the bottom plate marked under the desired switch arm. It was then drilled and tapped for a 10-24 stainless screw 2 inches (5 cm) long. A jam nut was threaded on and tightened after the screw was adjusted up or down to make proper contact with the switch arm.

RFI

My whole assembly is quite close to the dipoles used in my ham set-up. RFI from the transmitter to the timer ICs could have been a problem. To reduce this possibility, ferrite beads are used on all wires going in and out of the control box; bypass capacitors are installed, and the whole assembly is grounded. Happily, all this worked and there is no interference problem.

a current-to-voltage converter

If you want to add some "smarts" to the system, you may want to provide a means of stopping the timing when the charging level is sufficient. I chose a level in excess of approximately 0.5 amp as the cutoff point.

This scheme was implemented by adding a series resistor of 0.2 ohm in the positive line coming from the PV panel (see **fig. 6**). One-half of a dual IC (LM-358) was arranged so that the non-inverting input received the full output of the PV panel. With a small signal transistor in an emitter follower connection, the combination gives a voltage output which is proportional to the drop across the 0.2 ohm resistor.

The other half of the LM-358 is used as a comparator. Its reference is the zener regulated panel voltage, and is adjusted by the 10K pot. It is set to change state when approximately 0.5 amp is provided by the panel. If the level is lower than this, the comparator goes high and switches on the TIP33 gate transistor, which allows the supply voltage to go to the main timer. Rotation of the panel then continues. If the level is in excess of 0.5 amp, it stops timing (and rotation) so that the system will work efficiently.

Should you want a higher or lower charging level, it is simple to arrange. Just reset the comparator's pot.

references

1. G.J. Millard, "Solar Powered Regulator Charges Batteries Efficiently," *Electronic Circuits Notebook*, McGraw Hill, 1981, page 337.
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the bicycle-powered station

All you need is some simple circuitry, a junkyard alternator, and a ten-speed bike

By combining human energy with a ten-speed bike, a used automotive alternator, and some simple circuitry, Amateurs interested in exploring alternative sources of power can put a station on the air with electricity derived from a low-cost system of only modest complexity.

The automotive alternator — readily available, inexpensive, and capable of producing up to half a kilowatt of power—is the perfect starting point for experimenters interested in producing their own electricity. This bicycle-driven power plant, designed around the common Delco-Remy alternator, provides 110 volts of regulated AC at levels up to 150 watts. Designed primarily to run a 100-watt transceiver, the setup is used several times a week as an entertaining form of physical exercise. For that reason I intentionally omitted any means of energy storage, either electrical or mechanical. (The alternator and its controlling electronics can easily be adapted to other sources of mechanical power if desired, so the circuitry should also be of interest to Amateurs interested in inexpensive wind-, water-, or gasoline-powered installations.)

The bicycle powered 100-watt Amateur station. The bicycle's rear wheel is chain-coupled to the automotive alternator, which in turn powers the transceiver. There's no electrical energy storage and no battery backup. Mounted over the operator's head are the control electronics and step-up transformer which produce the system's regulated 110-volt AC output. (*Photo courtesy W1GSL.*)

As a power source, the bicycle presents some interesting design challenges. Because the operator provides 100 percent of the station power, high efficiency is a must if over-the-air huffing and puffing are to be avoided. There is also a large variation in the amount of power delivered during each pedal revolution, so good regulator action is necessary to hold the output constant. These factors combine to make the bicycle a difficult source of mechanical power to apply successfully.

The performance data in **fig. 1** through **4** shows how well these problems have been solved. Mechanically, the system is somewhat noisy but still quite smooth. Its overall efficiency is such that a middle-aged adult in average condition should be able to produce 50 watts continuously for an hour without undue strain.* (A typical 30- to 40-minute radio contact leaves this operator damp but by no means exhausted.) The electronic regulator smooths out pedal-rate variations so they result in only a 2 to 3 percent modulation of the system output. In terms of transient response and spike suppression, the system is probably better behaved than the more expensive gasoline-powered generators so often used by Amateurs.

the alternator as a power source

Before discussing the design of this system, it's worth considering the alternator itself and several possible ways it can be used. The circuit diagram of the Delco-Remy unit I used (before and after modification) is shown in **figs. 5A** and **5B**. Basically, my alter-

*Consult your physician before undertaking any form of exercise program — **Ed.**

By **Penn Clower, W1BG**, 459 Lowell Street, Andover, Massachusetts 01810

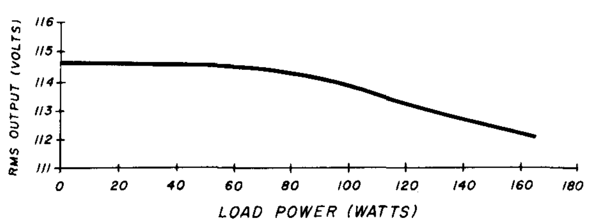

fig. 1A. Output RMS voltage regulation as a function of resistive load power. The total voltage drop is only 2 percent for a 160 watt increase in output power.

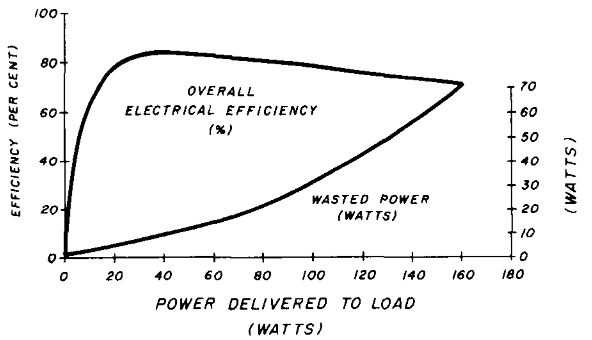

fig. 1B. Electrical efficiency. The power not dissipated in the load is lost as heat in the alternator stator and transformer resistances or is used to excite the field winding.

nator is a three-phase machine with a Y-wound stator driven by a rotating field coil. In normal use the three output phases are full-wave rectified to produce DC output with approximately 14 percent ripple. This output level is adjusted by controlling the field excitation current with an external regulator.

The windings are arranged so that seven electrical output cycles occur during one mechanical shaft rotation. This accounts for the relative advantage an alternator has over the common two-pole DC generator in low-speed applications. For a given field current, the unregulated output voltage of either machine is directly proportional to shaft rate, but the 7-to-1 electrical advantage of the alternator means it can reach the desired output level at a much lower mechanical speed.

In automotive use the regulator design can be quite crude because the car's storage battery does a good job of filtering the alternator's output. A more sophisticated regulator is necessary when the unit is used as the primary power source for a system having a minimum of energy storage capability. Overall regulation should limit the output level to within 5 to 10 percent of nominal, and care must be taken to limit the amplitude of any output spikes.

Lack of energy storage will also create another problem not found in the automotive environment. During normal operation the regulator circuitry will be powered from the alternator output. At startup there isn't any appreciable output, so a low-level source of regulator power must be provided along with a means of smoothly shifting back to internal power once the system is stabilized.

Typical car alternators can produce output powers on the order of 400 watts. It's worth noting that the internal resistance of the unit used here is 0.286 ohm, so at that output level the alternator I^2R losses are almost 300 watts. This explains why the unit is normally force-ventilated by a fan. Some care must be taken, particularly if the alternator is to be bicycle driven, to minimize that wasted power. Since the internal resistance is a given constant, the only way to increase efficiency is by lowering output current, and that means either lowering output power or increasing output voltage.

application techniques

With that as background, let's consider two possible ways of using the alternator. Since much of the newer radio equipment is capable of operating directly from 12 volts DC, one immediate suggestion is to simply use the alternator as a source of low-voltage DC power. Another approach, the one used here, is to tap into the AC portion of the alternator circuit with a transformer and bring the output level up to 110 volts. Each of these choices has its merits and disadvantages.

While I didn't explore the DC approach in depth, one problem that might be anticipated with that method is worth mentioning. When used without some sort of energy storage, the DC output of the alternator will have a ripple component equal to 14 percent of the peak level. Assume the decision is made to filter that ripple down to 0.5 volt peak-to-peak (3.8 percent) when the alternator is driving a 100-watt load. At the expected ripple frequency, an 8000μF capacitor will be required to do this, and the addition of that capacitor can complicate the control loop. The capacitor gets charged through the low output impedance of the alternator, but it must get discharged through the unpredictable and often variable load resistance. This means that the regulating control loop will respond differently to overvoltage and undervoltage transients. It might be difficult to design a circuit that would work well in all situations, and certainly some extra attention would have to be paid to the issue of loop stability.

The AC approach avoids this problem and has, in addition, the advantage of being able to run just about every piece of equipment in the station. The AC frequency will vary and is typically near 300 Hz, but that's

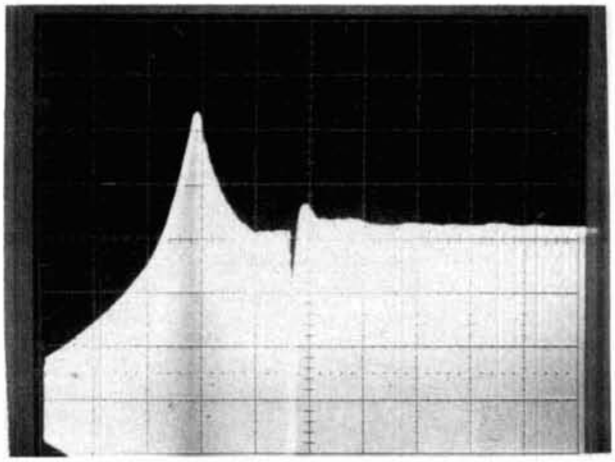

fig. 2A. The system startup transient. The AC output waveform is shown at 50 volts and 0.2 second per division; the zero level is 3 lines below the 'scope centerline. For this test the unloaded system was brought up to speed as rapidly as possible. Note the large overshoot which builds up before the regulator circuit can gather its wits. The small negative going spike at midscreen occurred when the battery driven startup current was removed.

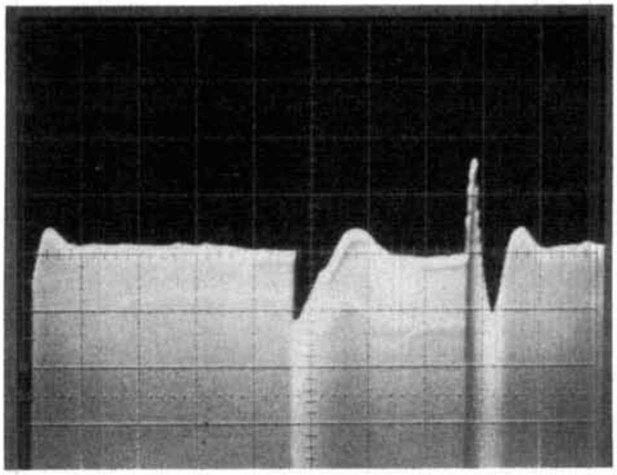

fig. 2B. Output voltage change caused by the sudden application (at midscreen) and removal (3½ divisions later) of a 75-watt resistive load. Scaling is 50 volts and 0.1 second per division, and the zero volt level is 3 lines below the centerline. This is the combined reaction of the regulator and bicycle operator; the regulator itself is even better behaved than the picture would indicate. The positive spike which occurs when the load is removed has been considerably shortened by the action of the neon bulbs.

of little consequence unless the equipment powered includes a clock or fan designed for 60 Hz. For most applications a valid argument can be made that higher input frequency will result in lower transformer core losses and improved power supply filtering. Furthermore the use of a step-up transformer allows the designer to select the alternator output voltage independently of the system's 110-volt output level. This means it's possible to set the system's output impedance for maximum efficiency at the expected power levels.

The biggest disadvantage of the AC technique has to do with the system output impedance and the nature of the load currents. Power supplies do not look like resistive loads because current is drawn only from the peaks of the rectified sine wave. Not only does this mean that current flows in pulses, but because of duty-cycle constraints the amplitude of those pulses is from four to six times the level of the DC current leaving the supply.

A small system like this is going to have a relatively large output impedance, about 24 ohms, and currents flowing through that impedance will cause a drop in the output voltage. Of course the regulator's job is to compensate for output voltage drops by increasing the field current, but unfortunately the regulator's bandwidth is only about 15 Hz. While it's perfectly capable of removing most of the effects of something like a resistive load being switched on and off at a few Hz, the regulator is powerless to do anything about the large, twice-per-cycle current pulses drawn by a transmitter's power supply. The frequency of that disturbance is simply too high. In such a case the controller may very well regulate the output RMS level as intended, but the wave form of that output will degenerate gradually from a sine to a square wave as the pulsed load currents increase.

That distortion might or might not have an effect on the radio used as a load. My transceiver is an older one with tube-type driver and amplifier stages. The filaments see no change because they react only to the constant RMS value of the supply voltage. The B+ does drop a little during a transmission though, and the peak input power to the transceiver is somewhat lower than when operated from commercial power. The lower B+ level is not harmful to the equipment, of course, and other than the decreased power level there is no noticeable effect on over-the-air performance.

design philosophy

Once the choice to produce 110-volt AC power is made, the design can proceed to the block diagram stage. At the start several measures can be taken to reduce the impact of the waveform distortion mechanism mentioned above. The first thing is to minimize the output impedance by selecting a relatively high alternator operating voltage.

The alternator windings normally produce a peak output after rectification of about 15 volts. That voltage is the vector sum of two sine waves phased 120 degrees apart and corresponds to an AC level before rectification of 10 volts RMS. The resistance of the stator winding is 0.286 ohm, so the transformation of 10 volts into 110 volts will create an overall

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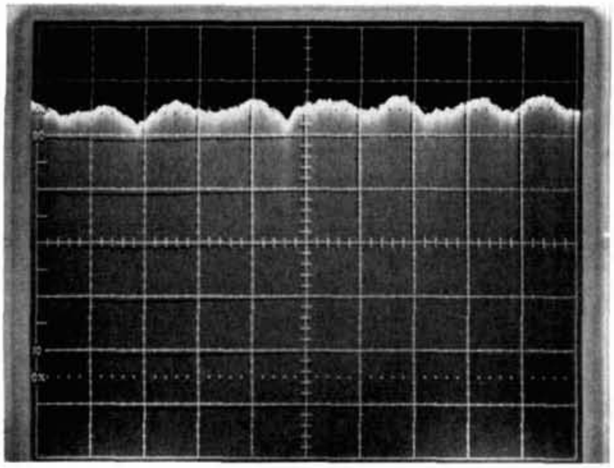


fig. 3. System response to an input disturbance. This 2 to 3 volt (rms) ripple on top of the AC envelope is all that's left of a 30 percent input power variation caused by pedal pumping. Scales are 10 volts and 0.2 second per division. Flywheel energy storage can filter the operator's power strokes enough to mask this problem, but that effect was minimized for this photograph by placing a 100-watt resistive load on the system's output.

output impedance of 34.6 ohms (0.286 ohm times the square of the transformation ratio). Operating the alternator instead at an output level of 12.6 volts RMS will lower that impedance to 21.8 ohms. That higher operating voltage will require a bit more field excitation power, but the lower output impedance is worth the expenditure. In fact, it might be worth trying an even higher voltage transformer if one is available. An 18 to 110 volt transformer would lower the output impedance to 10.7 ohms and possibly provide an overall gain in electrical efficiency as well.

The right transformer can, by lowering the output impedance, reduce the amount of distortion caused by a pulsed current (rectifier) load. The load can be protected against the harmful effects of any remaining distortion by designing the regulator to stabilize the RMS value of the output voltage.

The regulator could be designed to control the peak level of the output voltage, but in that case any distortion will cause the RMS output to increase as the load increases. That's because, as mentioned earlier, large pulse-type load currents will cause the output wave form to degenerate into a square wave. A sine wave with a peak level of 155 volts has an RMS level of 110 volts, but a symmetrical square wave with that same peak voltage has an RMS level of 155 volts.

Before the RMS level can be controlled, it must be measured, and that means an RMS detector is necessary. True RMS converters can be quite complex and expensive, but for this application there's an easy alternative. The high voltage waveform is full-wave rectified and the average DC value of that output is extracted with a suitable low-pass filter. The resultant voltage is not quite the same as an RMS measurement, but is close enough to do the job.

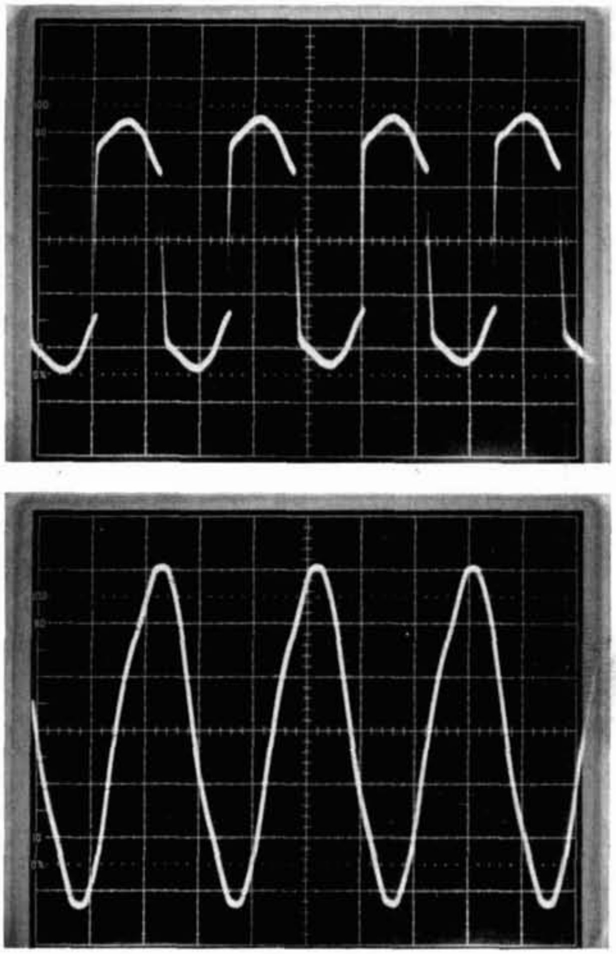


fig. 4. Waveform distortion caused by power supply loads. Scales for both photographs are 50 volts and 1 ms per division. Top photo shows the output AC waveform when the system is delivering 75 watts directly to a resistive load. In bottom photo, that load has been replaced with a transformer-rectifier-filter power supply, which is delivering the same power in DC form to a resistor. As discussed in the text, the peak output level has decreased but the RMS voltage is holding nearly constant.

Another consideration in the design of the regulator circuit is loop stabilization. As with any feedback loop, it's necessary to select components carefully in order to realize the desired response to a disturbance input. When, for example, a sudden 100-watt load is applied, it's expected that the output level will immediately drop about 20 volts (roughly 1 amp flowing through the 24-ohm output impedance). The regulator should then increase the field current so the output voltage rapidly and smoothly returns to 110 volts. In addition to being quick, this recovery should not involve any significant overshoot or ringing.

circuit design

The circuit diagram of the bicycle-powered system

A length of ordinary cycle chain drives a ten-speed sprocket cluster mounted on the alternator shaft. The rear bike rim is wrapped in black friction tape and drives the chain like a large pulley. Crude perhaps, but the drive system is also inexpensive and mechanically efficient. (*Photo courtesy W1GSL.*)

is shown in **fig. 6**. The alternator, shown inside the dashed rectangle, has been rewired slightly to optimize power use. This modification, as detailed in **fig. 5B**, separates one leg of the original Y winding to provide a source of low-voltage DC power for field excitation. Four of the original six alternator diodes are used for this purpose, while the other two are disconnected. The nominal output of this supply is about 10 volts, and that's a good level for Q4's collector supply, since it leaves only 3 or 4 volts to be dropped as waste across that power transistor.

transformer

The remaining two legs of the Y are brought out to drive the primary of the step-up transformer. That AC output is also connected to a bridge rectifier which provides operating potential for the low-level regulator stages. Since those stages don't draw much current, there's no power penalty in running them with this higher voltage, and the extra dynamic range is beneficial. With the regulator powered in this manner, it's possible to run the AC output all the way down to 50 volts before control is lost. The system isn't normally operated at that level, but having that much dynamic range improves the circuit's smoothness during start-up and transient conditions.

The main 12.6 to 110 volt step-up transformer was salvaged from an old black and white television chassis. Any similar transformer, or combination of transformers, having a 12.6 volt winding rated at 8 to 10 amperes will work perfectly well. The overall transformation ratio isn't critical, but primary voltages lower than 12 should be avoided for efficiency reasons.

regulator

The regulator samples the 110-volt output of the system with a small 1:1 transformer. This transformer drives a bridge rectifier through a series circuit made up of various resistors and neon bulbs. This circuit, along with the RC filter that follows the bridge, makes up the RMS (well, not quite RMS, but let's call it that anyway) converter. The idea is that the large effective value of series resistance between the transformer and bridge causes that combination to drive the capacitor like a current source. The capacitor charging and discharging rates are then very nearly equal and determined by the 16.6k load resistance. As a result, the capacitor voltage is free to follow the average value of the rectified voltage instead of its peak.

The neon bulbs are an overvoltage protective feature. Should the AC voltage exceed their firing level, they provide a low-impedance path that quickly charges up the filter capacitor, thus rapidly shutting off the field current. The resistors in parallel with the bulbs are carefully chosen unequal values which help minimize the difference between firing and sustaining voltages. This is about 15 volts per bulb, so the sequential firing encouraged by the resistor choice shaves 30 volts from the total trigger level. That sets the protective threshold at 190 volts peak or 135 volts RMS for a sine wave. Putting the resistors between the transformer and bridge lowers the inverse voltage seen across the diodes so that inexpensive small signal types may be used.

control system

A portion of the filtered RMS output is sampled with the 10K potentiometer and applied to one input of a differential amplifier. A 5.1-volt stabilized reference is applied to the other input, and the difference is amplified with a gain of 15.

The output at the collector of Q2 drives a common emitter stage, Q3, whose gain is partially determined by the impedance of the alternator field winding. It's a little unusual to have the gain of one stage influenced by that of another, but in this case the incomplete buffering is intended as part of the loop compensation. At currents in the 1-amp range the beta of power transistor Q4 is about 40, so the 5 ohm, 0.6 henry impedance of the field winding appears at Q3's collector as a 200-ohm resistor in series with 24 henrys. That impedance combines with the 3.3K collector resistor

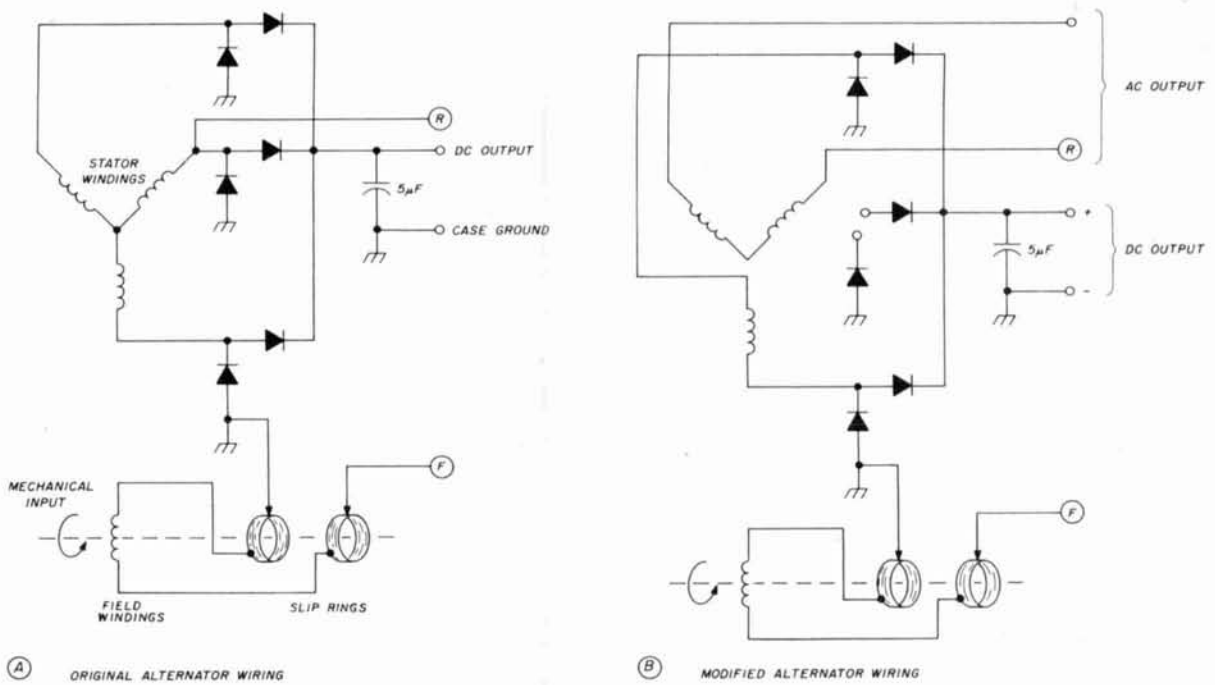
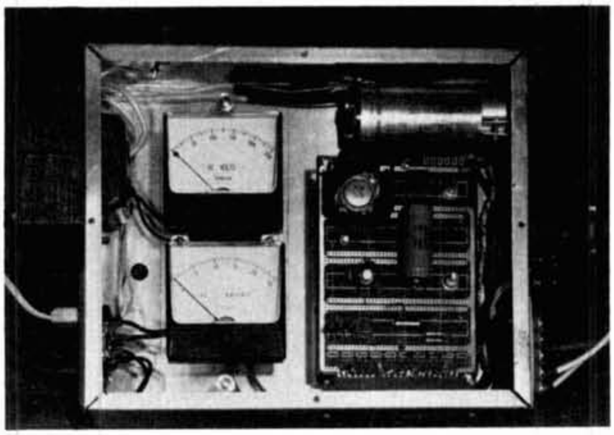


fig. 5. Electrical changes to the internal wiring of the Delco-Remy automotive alternator.



The controller electronics are mounted in and around a standard chassis. Metering is provided for alternator field current and AC output voltage. A set of four flashlight batteries, mounted out of sight behind the circuit board, provides field current through a push button when the system is first started. (*Photo courtesy W1GSL.*)

to create the lead-lag compensation necessary to stabilize the system.

The output transistor, Q4, operates as an emitter follower, and as always with emitter followers, some care is needed to prevent high-frequency oscillations. No instabilities were observed with this circuit, but some high-frequency loading (the series 0.001µF and 1K ohms) was added in any case. The 10µF capacitor is a termination for the AC voltage generated in the field winding by the stator current (back EMF).

One last circuit feature is startup field current provided by 4 D-size batteries through a momentary contact push-button switch and diode. When the setup is first started, this button is pushed and the alternator brought up to speed. The output generated by the battery current is enough to power the regulator, which can then bootstrap the system on up to 110 volts. When that happens, the diode in the battery circuit back-biases so the regulator can control the output level, and of course the operator can release the pushbutton.

mechanical design

Most of the mechanical details are visible in the photographs. The old and inexpensive 10-speed bike that drives the alternator is mounted at its rear axle to a large and stable base made from scrap lumber. The front wheel is removed and replaced with a wooden support. (While not strictly necessary, this adds stability to the setup by discouraging rotation of the front fork.) The alternator itself is mounted much as it would be in a car. Simple but sturdy aluminum brackets accept a mounting bolt that can be loosened to allow tension adjustments of the drive chain.

The alternator is chain-driven from the rear rim of the bicycle. The tire and tube were removed from the rear wheel, and the rim was trued up as much as possible with a spoke wrench. Any protruding spoke ends were filed flat and the inside of the rim was wrapped with several layers of black cloth friction tape. Mean-

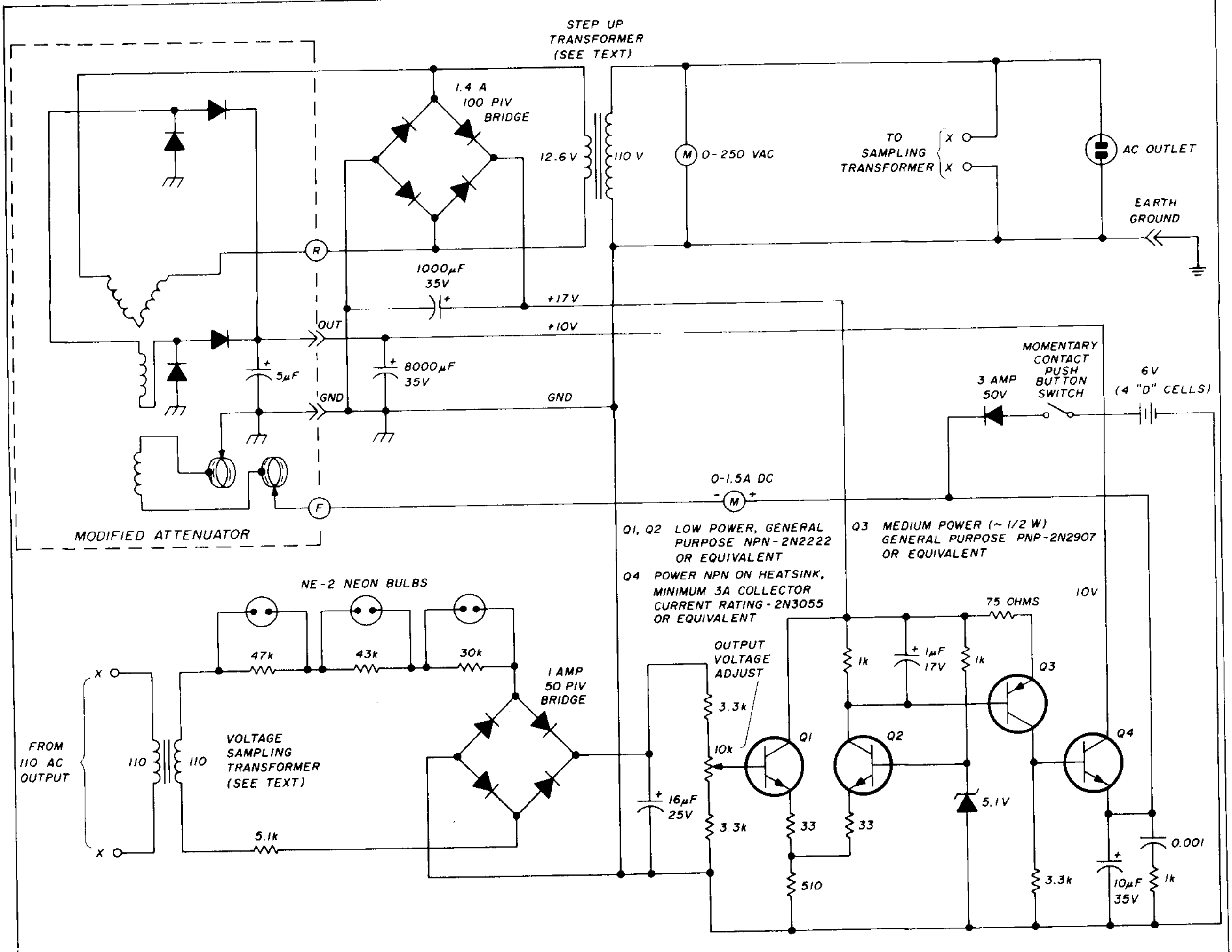


fig. 6. Schematic diagram of the bicycle-driven 110-volt power supply.

while, the alternator shaft had been fitted with a standard 10-speed rear sprocket cluster, so the bicycle rim and alternator could be mechanically coupled with a length of standard 10-speed bicycle chain.

This drive system may be simple, but it's also remarkably efficient, trouble-free, and inexpensive. The chain is friction-driven by the wheel rim, but the contact length is long (about 5 feet) and there is no noticeable slippage. The roller chain is quite flexible and has a smooth but positive grip on the alternator drive sprocket. The only thing at all fussy about the drive is that the bike rim must be reasonably circular since any radial runout will cause a variation in chain tension as the wheel revolves.

The drive chain used is custom-fitted to the mechanical setup by combining portions from two standard length chains. Ten-speed cycle chains have no master link and are instead adjusted with a special tool that disassembles links by pushing out the rivets. Available for about $5 in most cycle shops, that inexpensive gadget makes it possible to easily adjust the length of the drive chain.

The 10-speed sprocket cluster, as shown in **fig. 7**, is held on the alternator shaft with much of the original hardware. A simple spacer turned from scrap metal centers the cluster on the shaft and provides a flat surface for the clamp nut to seat against.

electrical construction

There are no sensitive areas in the regulator circuit, so layout and construction can really take any convenient form. The controller shown in the photographs was built in and on a 10 × 12 × 3 inch (25 × 30 × 7 cm) aluminum chassis. The leads running to the alternator (No. 8 AWG wire was used for the high-current AC wiring) connect to a barrier terminal strip on the right side of this chassis. The AC output leaves through a standard AC receptacle mounted on the left. Directly above the AC outlet is the step-up transformer. A small 1:1 isolation transformer, mounted on

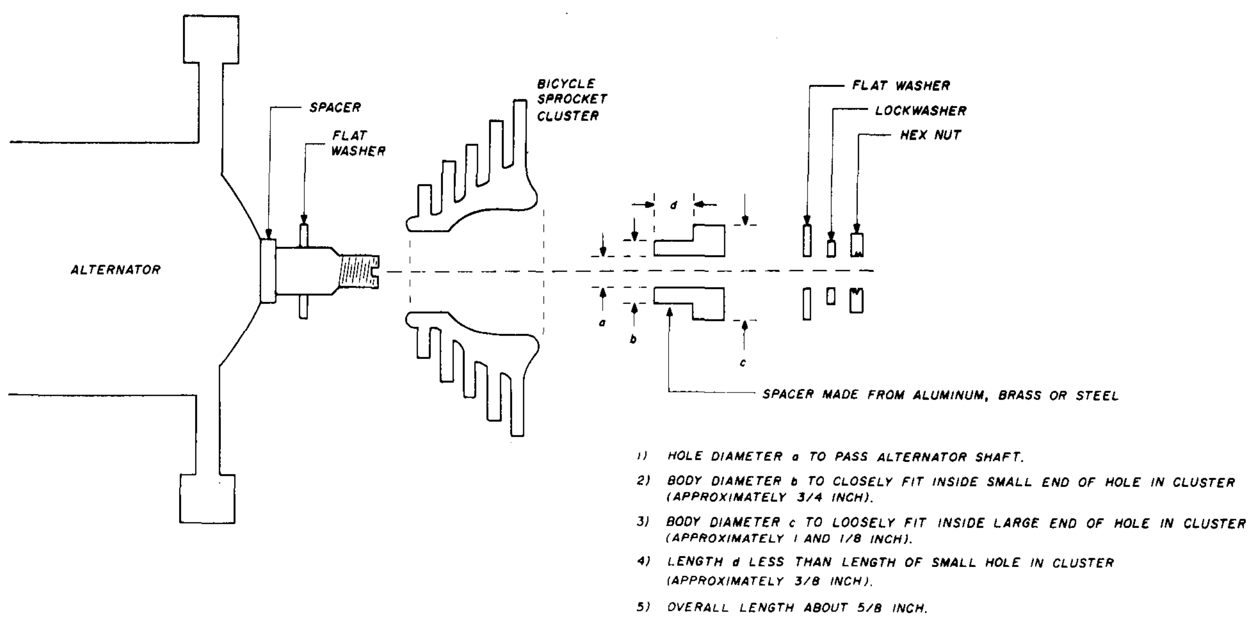


fig. 7. Method of mounting the sprocket cluster to the alternator shaft.

the right side of the box, samples the output; aside from being expensive if purchased new, it's bigger than necessary. A pair of very small filament transformers connected back-to-back would work just as well since the RMS converter circuit draws only a miniscule amount of power.

The smaller components of the regulator circuit are mounted on a standard 4.5 × 6 inch (11 × 15 cm) circuit board mounted inside the chassis. The plug-in feature isn't necessary, but was convenient to use during the development stage of the project. The 4 D-sized cells used to provide startup current are mounted out of view behind the circuit board.

In keeping with standard safety practice, one side of the AC output is tied to the chassis, which is in turn properly grounded. This step is necessary to prevent stray coupling and leakage paths from creating a large potential difference between the AC outlet and radio ground.

checkout and operation

Operation of this system is simplicity itself, but initial checkout should proceed with care. There is a surprising amount of power available here, and some expensive radio equipment is likely to be plugged into it. *Don't take any chances with your transceiver until you're sure the regulator is functioning properly!*

A lot of testing can be done without any load at all on the system. First set the voltage control potentiometer to mid-position. Press the field excitation pushbutton and start pedaling. The AC output voltage will slowly increase to 50 volts or so, then jump upscale to its final value. This level can then be set between 110 and 115 volts with the pot.

The neon bulbs may flash briefly when the output voltage first swings upscale. That's because the filter capacitors in the regulator power supplies take a moment to charge up. Until they do, the regulator isn't fully operative and the output AC level can build up rapidly. A worst-case example of this effect is shown in **fig. 2A**. The large overshoot in this photo resulted from jerking the alternator up to high speed as rapidly as possible. The size of the overshoot can be reduced a bit by smoother pedaling, but it always makes sense to have the output voltage up and stable at 110 volts before an electrical load is applied.

If everything seems satisfactory at this point, start loading the output. Begin by trying light bulbs of different wattage so you can develop an appreciation of how much effort is required to produce so many watts of power. An oscilloscope can be used to check how the system responds to suddenly applied loads. Light bulbs draw quite a bit of inrush current when first switched on, so they don't make a particularly good load for transient analysis. A soldering gun is better for this purpose because it represents an almost purely resistive load and can be switched on and off easily.

Other interesting loads include things like old five-tube table radios and small television sets. *Repeat: Only when you're convinced the circuit is working smoothly and reliably should you plug in a valuable* *transceiver.* The circuit of **fig. 6** is a mature design and works quite well, but I'm sensitive to all that can go wrong because I burned out a large number of 100-watt light bulbs in the early stages of this project!

performance

The measured electrical performance of the system is detailed in **figs. 1** through **4**. No effort was made to measure mechanical performance due to the lack of both suitable instrumentation and a data base to use as a yardstick.

Electrical measurements are much easier to make, and **fig. 1** shows two of the most important: steady-state regulation and overall electrical efficiency. To make these curves the system was pedaled at a constant rate while connected to several 110-volt light bulbs through a calibrated AC wattmeter and Variac. The output voltages and currents were then measured for a range of different load levels. The curve showing overall electrical efficiency is a measure of how much power was generated as compared to how much went to the load. The lost power was used either for field excitation or dissipated as waste in the alternator and transformer resistances.

The transient response photographs in **fig. 2** show how well the regulator responds in dynamic situations. The first picture shows the output AC envelope during the system power-up transient. The second part of **fig. 2** shows how the regulator handles the sudden application and removal of a 75-watt resistive load. It's difficult to maintain a constant pedal rate in the face of that load change, so the picture unfortunately shows the combined response of both the regulator and operator. Even with that complication the loop looks pretty good.

There is some flywheel energy storage in the moving components (most noticeably the high-speed chain drive), but when the electrical loading is heavy, that filter isn't enough to smooth the two-per-cycle power strokes coming into the system. **Figure 3** shows how well the regulator can hold the output level constant under those conditions.

Figure 4 shows the distortion effects in the output waveform caused by a power-supply load. The first trace was taken with 75 watts of resistive loading on the system. In **fig 4B** this was replaced with 75 watts of DC power delivered through a full-wave bridge rectifier to an RC load.

conclusions and suggestions

The system described has been in use for over nine months. I try to operate for at least half an hour every other day, so that nine months represents a lot of air time and a reasonable amount of physical conditioning. My transceiver draws 40 watts of power on receive and 80 watts of steady-state bias power on transmit. I can power the radio itself for well over an hour without tiring, so for exercise purposes the system is also loaded with a 40-watt light bulb.

With that loading the system produces 130 or 140 watts during transmissions and the output ripple due to pedal pumping becomes bothersome. As an easy cure, some 4 or 5 pounds (1.8-2.3kg) of flywheel weighting was added to the rim of the bicycle's rear wheel. That provides enough energy storage to operate the setup for several seconds, so even at high output levels the effect of pedal pumping is completely removed.

The only other modification under consideration at this time is changing the step-up transformer to an 18-volt unit. This would cut the system output impedance in half and so improve regulation, reduce waveform distortion, and increase overall efficiency in the 100 watt output range.

acknowledgement

Special mention must be made of the efforts of W1GSL, my trusty (and trusting) photographer. Not only does he make house calls, but he was willing to give the system a trial spin in his street clothes before getting down to photography.

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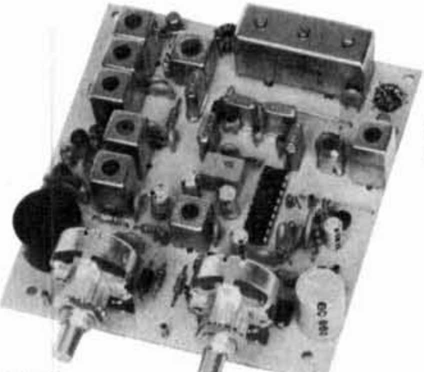

R144 Shown

- **R144/R220 FM RCVRS** for 2M or 220 MHz. 0.15uV sens.; 8 pole xtal filter & ceramic filter in i-f, helical resonator front end for exceptional selectivity, more than −100 dB at ±12 kHz, best available today. Flutter-proof squelch. AFC tracks drifting xmtrs. Xtal oven avail. Kit only $138.
- **R451 FM RCVR** Same but for uhf. Tuned line front end, 0.3 uV sens. Kit only $138.
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- **R110 VHF AM RECEIVER** kit for VHF aircraft band or ham bands. Only $98.
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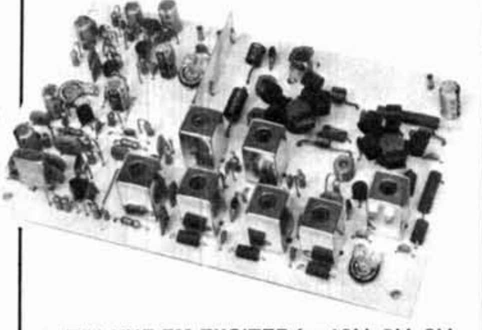

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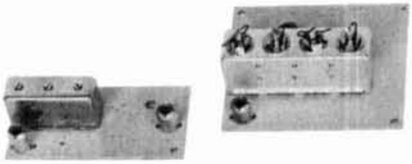

- **HELICAL RESONATOR FILTERS** available separately on pcb w/connectors.

HRF-144	for 143-150 MHz	$38
HRF-220	for 213-233 MHz	$38
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- **COR-2 KIT** With audio mixer, local speaker amplifier, tail & time-out timers. Only $38.
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- **NEW – SIMPLEX AUTOPATCH** Use with any transceiver. System includes DTMF & Autopatch modules above and new Timing module to provide simplex autopatch and reverse autopatch. Complete patch system only $200/kit. Call or write for details.

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MODEL	TUNES RANGE	PRICE
LNG-28	26-30 MHz	$49
LNG-50	46-56 MHz	$49
LNG-144	137-150 MHz	$49
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Our traditional preamps, proven in years of service. Over 20,000 in use throughout the world. Tuneable over narrow range. Specify exact freq. band needed. Gain 16-20 dB. NF = 2 dB or less. VHF units available 27 to 300 MHz. UHF units available 300 to 650 MHz.

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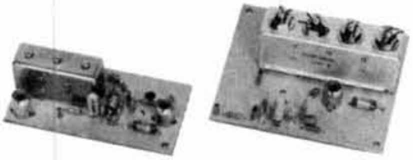
HELICAL RESONATOR PREAMPS



Our lab has developed a new line of low-noise receiver preamps with helical resonator filters built in. The combination of a low noise amplifier and the sharp selectivity of a 3 or 4 section helical resonator provides increased sensitivity while reducing intermod and cross-band interference in critical applications. See selectivity curves at right. Gain = approx. 12 dB.

Model	Tuning Range	Price
HRA-144	143-150 MHz	$49
HRA-220	213-233 MHz	$49
HRA-432	420-450 MHz	$59
HRA-( )	150-174MHz	$69
HRA-( )	450-470 MHz	$79

RECEIVING CONVERTERS

Models to cover every practical rf & if range to listen to SSB, FM, ATV, etc. NF = 2 dB or less.

	Antenna Input Range	Receiver Output
VHF MODELS	28-32	144-148
Kit with Case $49	50-52	28-30
Less Case $39	50-54	144-148
Wired $69	144-146	28-30
	145-147	28-30
	144-144.4	27-27.4
	146-148	28-30
	144-148	50-54
	220-222	28-30
	220-224	144-148
	222-226	144-148
	220-224	50-54
	222-224	28-30
UHF MODELS	432-434	28-30
Kit with Case $59	435-437	28-30
Less Case $49	432-436	144-148
Wired $75	432-436	50-54
	439.25	61.25

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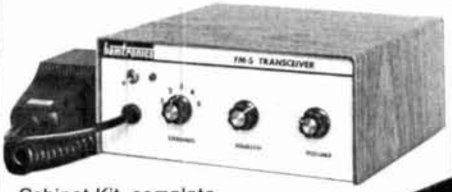

Cabinet Kit, complete with speaker, knobs, connectors, hardware. Only $60.

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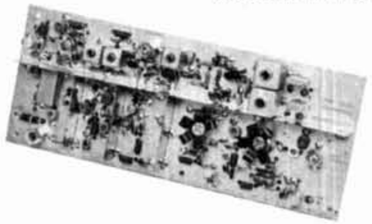
For SSB, CW, ATV, FM, etc. Why pay big bucks for a multi mode rig for each band? Can be linked with receive converters for transceive. 2 Watts output vhf, 1 Watt uhf.

	Exciter Input Range	Antenna Output
For VHF,	28-30	144-146
Model XV2	28-29	145-146
Kit $79	28-30	50-52
Wired $149	27-27.4	144-144.4
(Specify band)	28-30	220-222*
	50-54	220-224
	144-146	50-52
	50-54	144-148
	144-146	28-30
For UHF,	28-30	432-434
Model XV4	28-30	435-437
Kit $99	50-54	432-436
Wired $169	61.25	439.25
	144-148	432-436*

*Add $20 for 2M input



VHF & UHF LINEAR AMPLIFIERS. Use with above. Power levels from 10 to 45 Watts. Several models, kits from $78.

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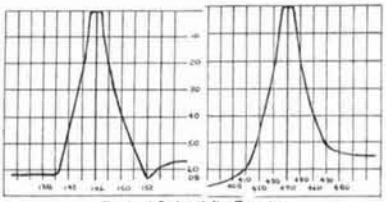

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the new industrial revolution: packet radio and local area networking

"I see no reason why intelligence may not be transmitted instantaneously by electricity."

— Samuel Morse, 1832

Because the traditional methods of point-to-point communications — CW and RTTY — relied on relatively low data rates, conventional systems have been slow to take advantage of available technologies and the bandwidths they offer. Yet the widespread use of digital computers has created an urgent demand for faster data transfer (burst communication) requiring large bandwidths giving a totally new meaning to Morse's words.

Now that personal computers have taken the place of comparatively "dumb" terminals, used to access large central computers (CPUs) via low-speed telephone lines, communication between and coordination of individual computers has presented a new and difficult challenge to system planners. Because there are differences between various computers, they have difficulty "talking" to each other, and problems are created whenever they are combined in a common network. In order to solve this problem, a new science, called LAN or *L*ocal *A*rea *N*etworking, was created, which employs packet-like communications to transmit data on dedicated wire (pairs), coaxial cables, fiber optics or "through the air" (see **fig. 1**).

The need for wideband communications has also affected traditional analog modulation schemes, which are now being replaced by digital ones. The search is on for techniques that will enable diverse types of electrical information — including voice and video signals — to be combined and transmitted. At the same time new higher speed analog-to-digital converters are replacing conventional processing techniques not only in RF processors but in large distributed networks and private branch exchanges (PBX). In addition, compressed full-motion digital television signals, including digital voice and computer information, can be distributed over networks for videoconferencing.

This phenomenon is not a science fiction fantasy, but rather a reality of the new industrial revolution, in which information is managed through automatic communications between and among "intelligent" machines.

As an example, consider a corporation with several manufacturing facilities in widely separated locations. Each factory acts as an independent unit with its own offices and production facilities. To operate effectively and maintain the competitive edge, each unit needs to have, at hand, easy access to the latest technology and techniques of information processing and communications. The answer to this need is a series of interconnected local area networks deisgned to instantly respond *electronically* rather than on paper.

communication path determines technique used

Most data communications over paths shorter than one hundred meters employ (hard-wired) parallel busses (see **figs. 2** and **3**.) Although this relatively fast method of data transfer is fully compatible with computer I/O's, use of busses over greater distances becomes impractical. Consequently, local area networks use serial digital packet formats at rates in the 50 Mbit/s range.* In communications this is referred to

By Cornell Drentea, WB3JZO, 7140 Colorado Avenue N., Minneapolis, Minnesota 55429

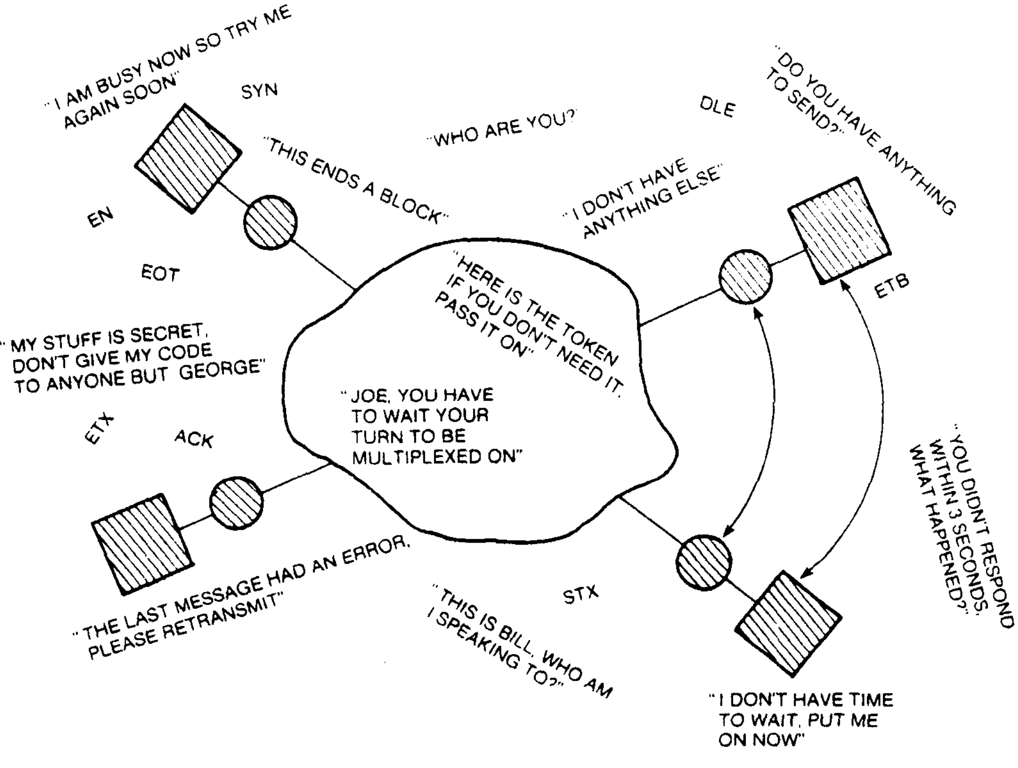


fig. 1. Some protocol requirements for local area networks.

as base-band information because it is the basic format that every device will eventually generate and accept for communications. Base-band can be used directly in point-to-point communications between data devices such as universal asynchronous receiver/transmitters (UART). However, without complex protocols and equipment, only two devices can communicate over the given medium at a given moment. For communication among more than two devices (e.g., computers), other methods must be used.

One method of communicating between several digital base-band devices over the same pair of wires is by multiplexing them synchronously at both ends. Known as time division multiplexing (TDM), this represents an early form of local area networks, as shown in **fig. 4**. TDM uses permanently dedicated time slots that can sometimes be "empty" of information; *statistical* time-division multiplexers (STATDM) eliminates this problem by dynamically allocating the time slots according to the activity on the channels.

- ***Computer Bus:***
  1-10 cm to 100 meters.
- ***Local Area Nets:***
  Several meters to several kilometers.
- ***Local Distribution Nets:***
  A few kilometers to approximately 100 kilometers.
- ***Long Haul Nets:***
  Hundreds of kilometers to thousands of kilometers.

fig. 2. Categorization of networks by geography.

*Packet is a block of data handled by a network in a well-defined format including a header (opening ID) and having a maximum size and data field. Consequently, a message may have to be carried as several packets.

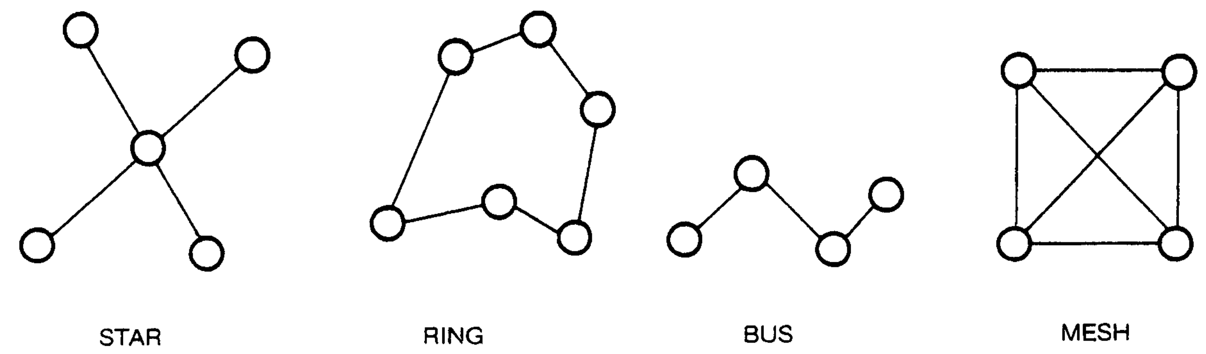


fig. 3. Types of local area networks.

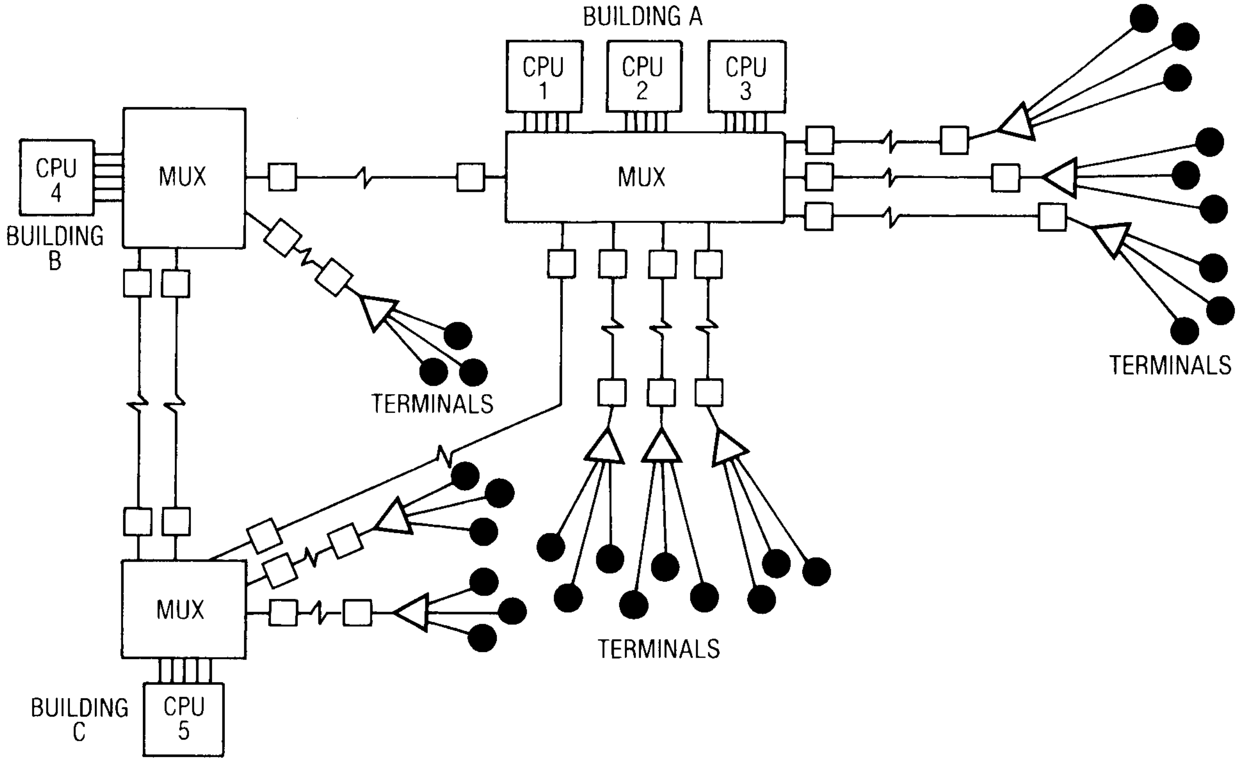


fig. 4. Block diagram of a campus local area network using multiplexers.

Other simple local area networks use an intelligent central switching facility in a star configuration that resembles conventional telephone PBX technology. By using base-band directly on wires and without multiplexing or central switching, we occupy 100 percent of that medium's bandwidth. (Also known as the percentage bandwidth factor.)

When base-band signals are translated to radio or optical frequencies, it becomes possible to pack more base-band channels into one single frequency, as shown in **fig. 5**. The higher the frequency, the more base-band channels can be transmitted.

Frequency division multiplexing, designated as the broadband approach to local area networks, can be further subdivided according to some interesting protocols, as shown in **fig. 6**.

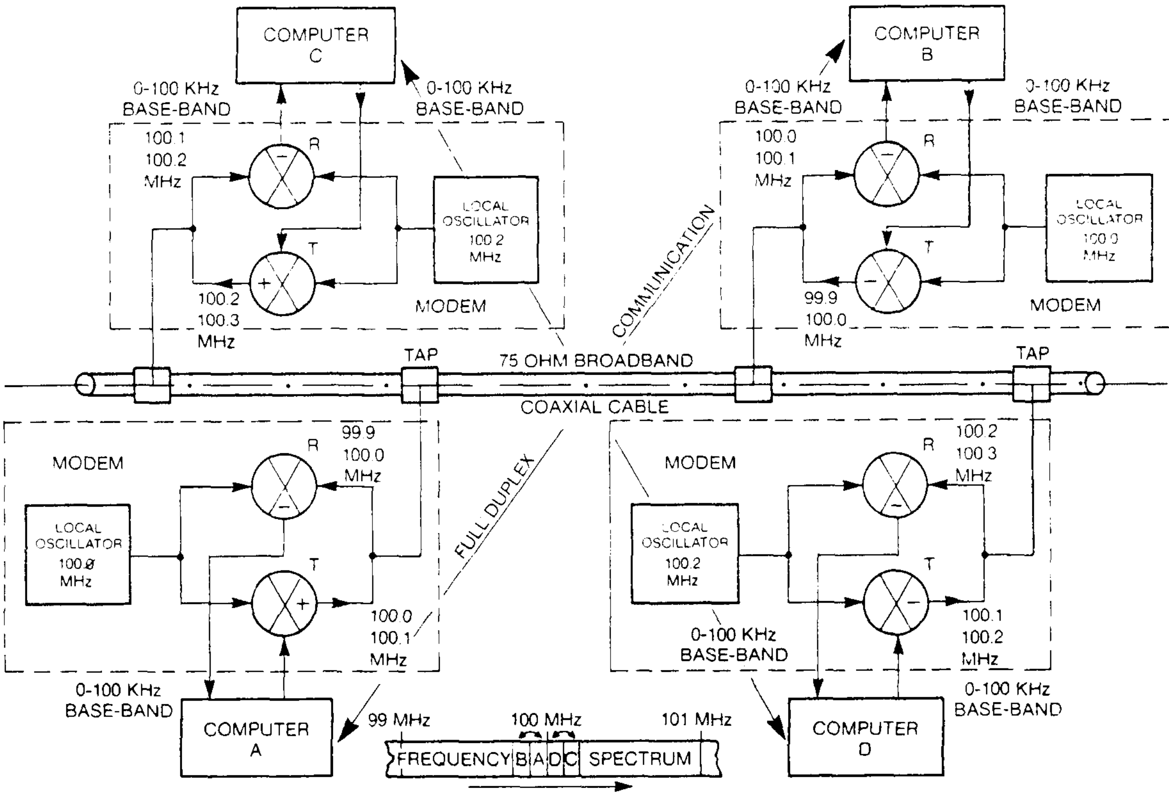


fig. 5. The concept of percentage bandwidth shows that the best medium utilization occurs when many base-bands are modulated on top of much higher frequencies as in frequency division multiplexing (FDM) techniques using radio-like processing.

listen while talking

In an effort to find better ways of utilizing the spectrum *without* switching and/or multiplexing new packet techniques using radio and coaxial cable networks have recently been developed and implemented. By borrowing ideas from real-life situations, designers have applied the characteristics of human interaction to communications between machines.

Imagine a business meeting in which several dynamic men and women exchange information and make decisions in one room. The process is rapid and contains several important elements common to all the speakers:

- A leading address to one or more participants preceeds any conveyed message.
- A message is conveyed.
- An end remark is usually followed by an invitation to answer.

If two persons begin to speak simultaneously their messages "collide," and one or both stops talking and tries again later. (In earlier packet radio, neither party would have "known" that a collision had occurred until they noticed that an acknowledgement had not been received at the end of the transmission.)

This relatively simple concept is the basis for a type of local area networking known as *Ethernet*. Ethernet allows the transmitting station to listen in on the "ether" while transmitting a packet, in a method known as "listen while talking." This method is similar to the situation that occurred in the hypothetical business meeting described above. It allows for any

collisions caused by another station that had decided to transmit to be immediately detected, which in turn causes the transmitter to cease transmission and restart at a later time. Ethernet can be used either at base-band over wire pairs and/or coaxial cables, or over channelized broad-band busses in CATV-like networks equipped with special modems as shown in **fig. 7**.

The concept was developed as a consequence of problems with simpler packet radio techniques that do not necessarily allow for the collision detection characteristic of Ethernet. These technologies evolved from the well-known ALOHA network. Developed at the University of Hawaii in 1970, the ALOHA system was essentially the first practical packet radio network, and covered the Hawaiian Islands, as shown in **fig. 8**.

The purpose of the ALOHANET was to provide inexpensive, statewide user-to-central computer communication for several hundred terminal users who were experiencing poor telephone modem connections at the time. Unlike previous networks, which used node-to-node communications, ALOHA network communication was accomplished in a radio broadcast mode, in which each station was heard and addressed by the central packet station known as the "Menehune", located in Honolulu, via two 100-kHz random-access channels at 407.35 MHz and 413.475 MHz. Although the system did not allow for direct user-to-user communications, information was nevertheless transmitted by transferring data to the central packet station and then forwarding it, after processing, to the destination user.

The packet transmission data rate was 9600 baud, with packets consisting of a 32-bit header, a 16-bit header parity check field, and up to 80 bytes of data followed by a 16-bit data parity check field. The maximum size packets were, therefore, 704 bits in length. Each took about 73 milliseconds to transmit; the entire network's propagation delay time was therefore, negligible in comparison to other systems. The random-access user-to-computer (407.350 MHz) channel allowed for packet headers containing user addresses to be identified by the Menehune central station. One natural consequence of the random nature of transmission was the probability of packet overlapping and collision. However, frequency division multiplexing was not considered necessary because of the burst nature of the computer traffic and the lack of additional radio channels. The collided packets were rejected by the Menehune station, and that fact was made known to the respective transmitting terminal by the absence of an acknowledgement signal on the computer-to-user return channel (413.475 MHz). This channel presented no random collision problem, since only the central processor's transmitter was broadcasting to all the other stations.

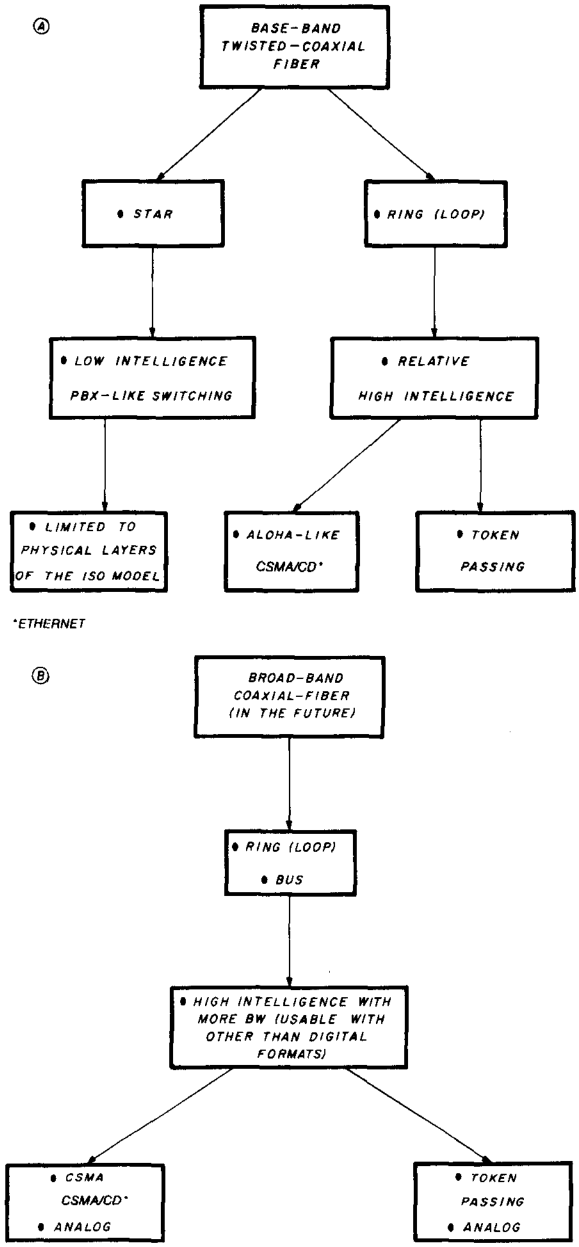


fig. 6. Characterization of networks by frequency domain and protocols; (A) using coaxial cables, (B) using fiber optic cables.

This scheme, known as pure ALOHA, presents some drawbacks as pointed out earlier; because the individual terminals have to wait until no acknowledgement is received, and then re-transmit the packet at random, with no guarantee that a collision will not occur again.

The relatively low throughput of pure ALOHA (see **fig. 9 A** and **B**) arises from the high practical probability of packet collision, which in turn is the result of

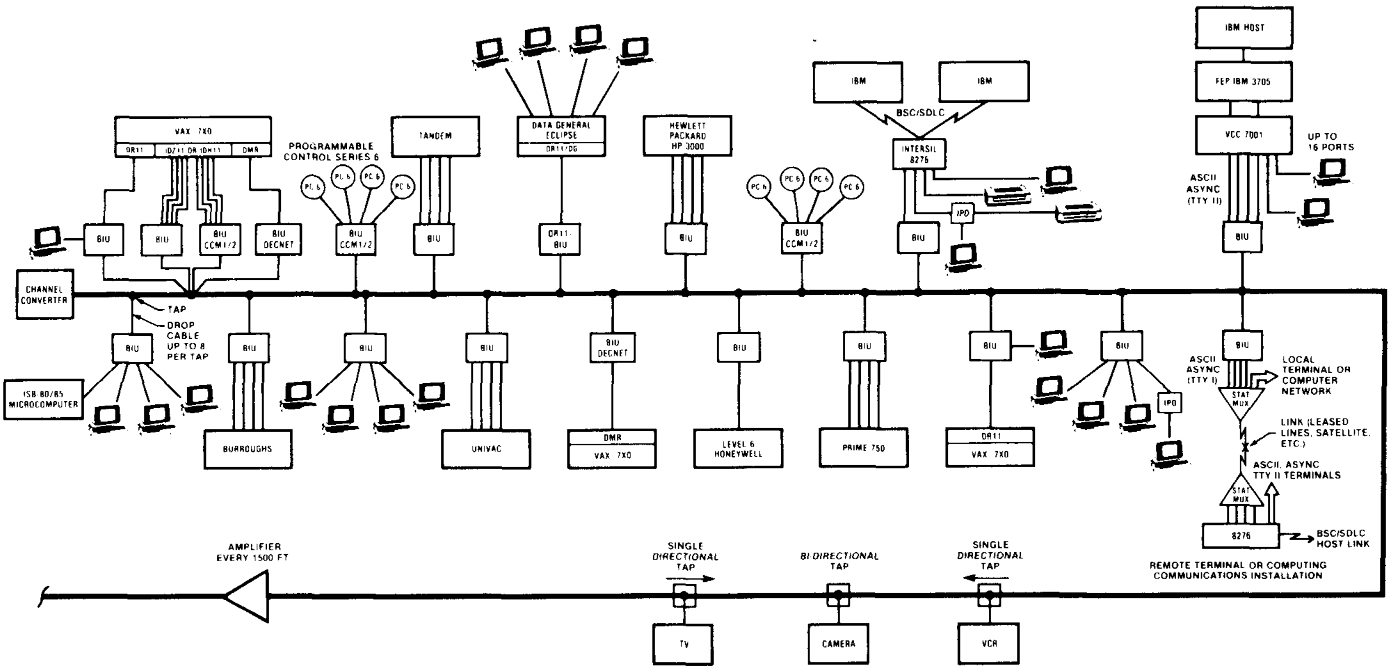


fig. 7. The GE-NET is an example of Ethernet implemented on a broad-band CATV-Network.

the total lack of discipline in transmissions from the terminals.

To correct the situation, various proposals were made to improve the bandwidth utilization. Among them were slotted ALOHA, which restricts and synchronizes terminal transmissions for better throughput and the various carrier sense multiple access (CSMA) techniques presently used.

The present AX.25 protocol used in Amateur packet-radio networks is superior to the ALOHA protocol because it listens before it transmits; however, the characteristic "listen while talk" feature of Ethernet is not available on these networks. The more advanced Ethernet concept was pioneered by the Xerox Corporation in November, 1978, as part of a proposed nationwide network known as the Xerox Telecommunications Network (XTEN).

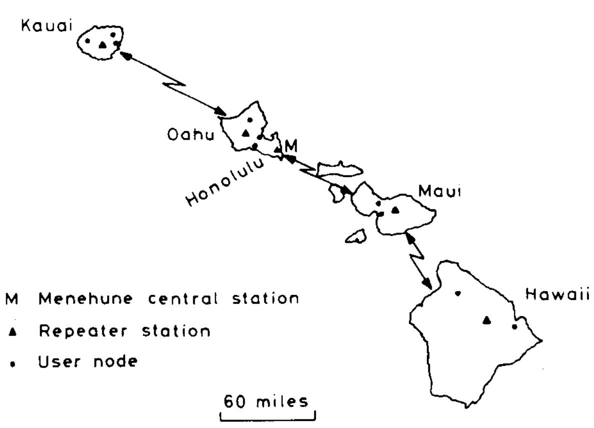


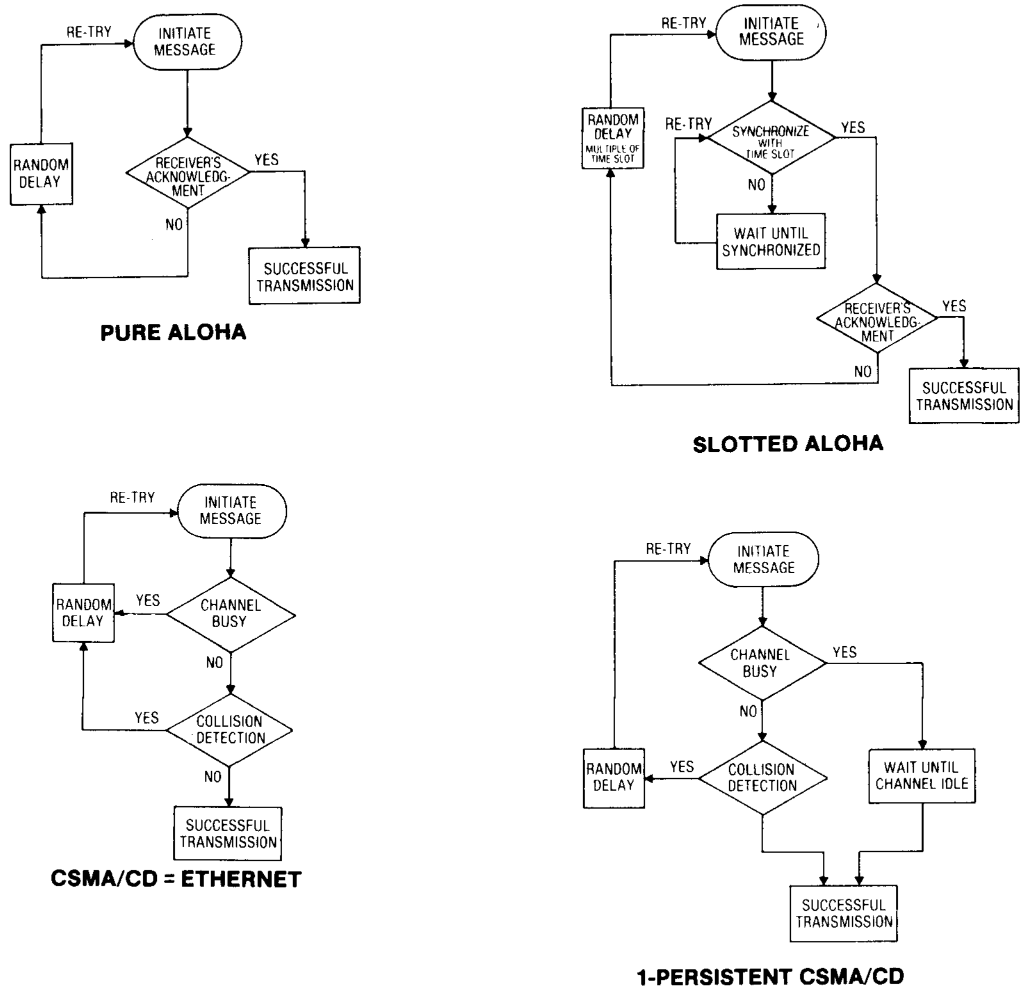
fig. 8. Map of the ALOHA network.

satellites would provide the link

The overall XTEN approach required satellite communications to link together through shared earth stations located in all the major cities around the country. Microwave radio, and/or infrared fiber links were envisioned to connect the local users to the earth stations, and finally, individual buildings and/or campuses were to be connected through a 3-Mbit/s packet network.

In its modern form, Ethernet is a base-band local area network allowing a theoretical rate of 10-Mbit/s over coaxial cable. It can connect as many as 1024 stations over a span of 7600 feet. Each station is designated by a 48-bit code and collision detection (CSMA/CD) is accomplished through the previously described techniques. In spite of the many improvements since the ALOHA days, the opponents of Ethernet still maintain that when network traffic is especially heavy, real-time applications such as those on aircraft or in automated factories could fail to be completed within the limits of their prescribed times, resulting in long collision recovery time, and even the risk of catastrophic failure.

Consequently, new methods have been developed to provide more discipline in accessing networks. One



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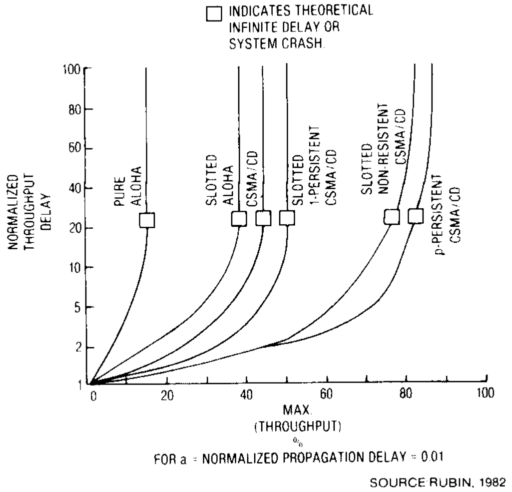

PROTOCOL	MAX. THROUGHPUT (%)
PURE ALOHA	18.4
SLOTTED ALOHA	36.8
CSMA/CD	40.0
1-PERSISTENT CMSA/CD	52.9
SLOTTED 1-p CSMA/CD	53.1
0.1-p CSMA/CD	79.1
NONPERSISTENT CSMA/CD	81.5
0.03-p CSMA/CD	82.7
SLOTTED NONPERSISTENT CSMA	85.7
IDEAL SCHEDULING	100

B

fig. 9. (A) Variations of the ALOHA concept, and their functional flow diagram, (B) system crash is indicated by the maximum throughput delay for the different methods shown.

such method is the newly introduced *token-passing* technique, which consists of passing an accessing token from terminal to terminal, which in turn provides a well-ordered method of accessing the network for the participating users. The relatively complicated error recovery algorithms of token-passing make this approach difficult to implement, however. (Most base-band, token-passing networks can also be implemented in broad-band.)

A compromise between the CSMA and *token-passing* is now being examined, with products to follow in the near future.

applying what we've learned

Shown in **fig. 10** is a radio transceiver with a built-in computer terminal; it could also be considered a computer terminal with a built-in radio transceiver. Behind its friendly look lies a packet radio network of unprecedented complexity.

Imagine a nationwide dispatching operation involving thousands of roving service or supply vans, with drivers. Obviously such an operation would require foolproof, effective communications and positive inventory control of all parts used in the field (with automatic reordering). A combination of radio paging, telephone dispatching, and paperwork would be necessary.

In such an application increased productivity and elimination of delays associated with the paperwork means improved communications. Equipped with this portable radio terminal, otherwise known as a PCX, a fleet can reliably access a distributed database network that keeps track of all inventories even as they are being continously modified from the field by all other portable radio terminals. This — and dispatching and ordering functions as well — happens in real time, in all locations, and without the inevitable delays associated with paperwork.

These developments are expected to open new doors in professional and Amateur packet radio activities, with products such as hand-held transceivers expressively made for data networking applications. Networks using satellite packet gateways will allow electronic mail to be carried from one side of the world to the other, and facilitate the handling of emergency traffic on a worldwide scale.

what we can expect

Today's personal computer is a small first step in the eventual connection of diverse communication technologies in terms of interactive thinking of machines. This new science has been named *telematique* or *telecybernetics*.

Theoretically, the ideal local area network should allow any computer or office/factory device to communicate with any other device within its geographical

fig. 10. The portable radio data terminal (PCX) can provide interactive data communications with large data bases geographically dispersed. Thousands of such devices can modify inventories in real time, while roving around cities as part of area-wide packet networks operating in the 800-MHz business band. The local networks are in turn connected with a nationwide data network which provides updating of information to all other PCXs located elsewhere on the network. (*Photo courtesy of Motorola Inc.*)

boundaries. However, current local networks are limited in that they perform only limited functions in specific settings. This approach may or may not survive the present technological revolution. Such is the case with some inexpensive base-band systems that can work with only certain brands of computers. *The problem is essentially one of standardization.* The age of the throwaway computer is here; the average life time of a design appears not to exceed two years. And although the new machines are increasingly powerful, with built-in LAN capabilities, combining them into networks remains one of the most challenging tasks ever encountered.

To help minimize these problems, the International Organization for Standardization (ISO) introduced the Open System Interconnection (OSI) model in 1978, a concept for developing communications between dissimilar devices. The model has since been adopted by the IEEE and is now reflected in the IEEE-802 standard, as shown in **fig. 11**. Upon completion, this standard is expected to ensure compatibility between local area networks with only a minimum amount of work required of the user.

conclusion

Today's local area networks cannot be expected to provide communications between dissimilar devices without additional customized hardware and software interfaces. At the same time, standardization is not expected to resolve the fundamental incompatibility of computers that use different operating systems, languages, and syntaxes. Technical experts seem in

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The MX-15 is a 15-meter band SSB/CW hand-held transceiver. It measures only 1½'' (D) × 2⅝'' (H) and offers 300mW for SSB and CW operation. A single-conversion receiver employing a MOS/FET front-end offers clear and sensitive reception. As a base or portable station, the MX-15 offers an unlimited challenge in QRP operation. Additional accessories are available to extend your operation.

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- **6 pc. AAA Batteries**
- **DC Cable**
- **Instruction sheet**

ACCESSORIES AVAILABLE

MX Channel crystal.... (Standard Frequency)	$7.00
MS-1 External Speaker-Microphone	$23.50
Noise Blanker Kit	$6.50
NB-1 Side Tone Kit	$11.50
SP-15 Telescoping antenna	$19.50
2M2 DC-DC Converter set	$17.50
PR-1 Mobile Rack Kit	$23.50
VX-15 External VXO (one crystal supplied)	$53.50
PL-15 10W Linear amplifier	$89.50

Photo shown MX-15, VX-15, PL-15, SP-15, MS-1 and PR-1

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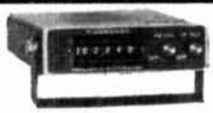

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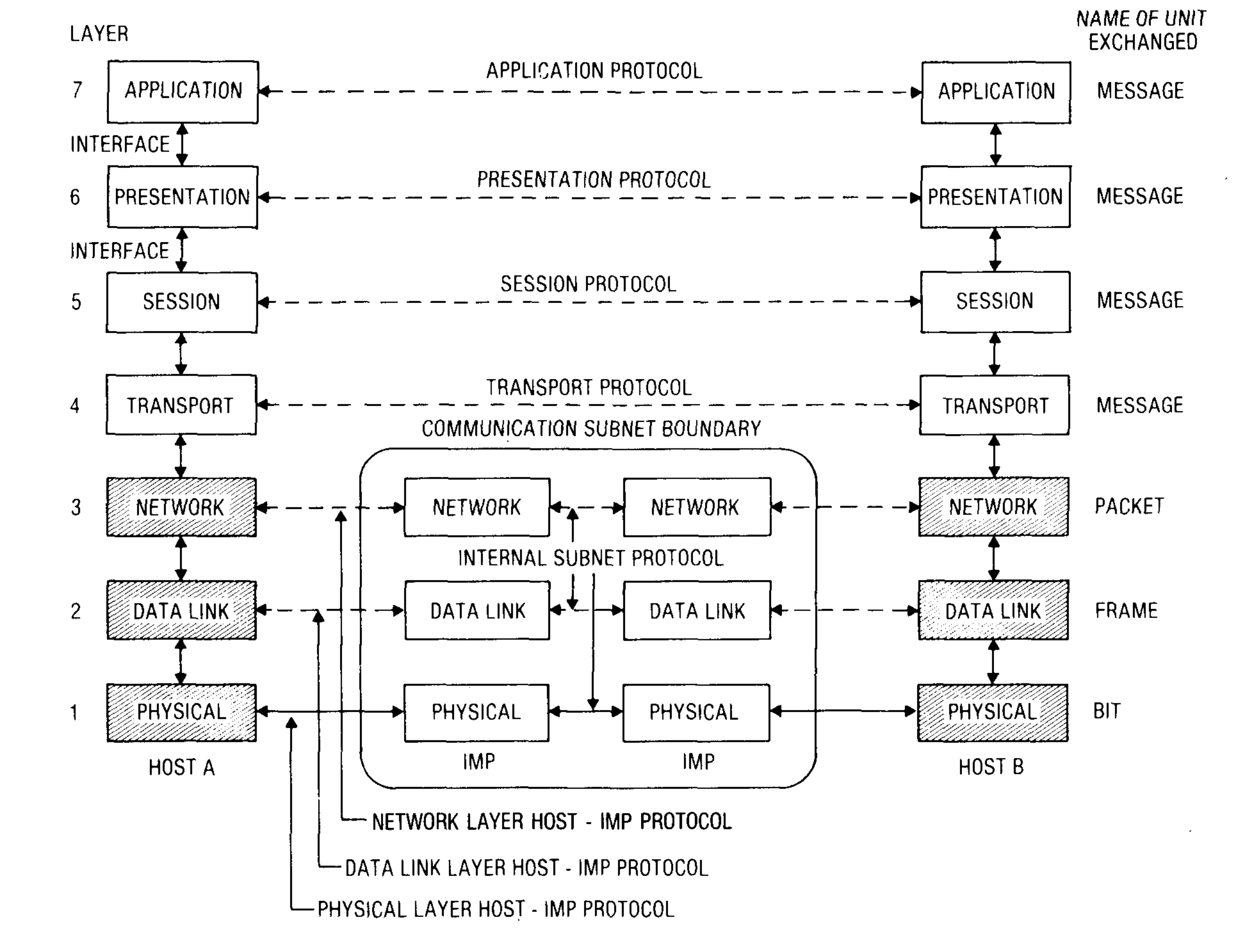

fig. 11. The seven-layer ISO OSI model has been adopted by IEEE's Project 802 Committee and a set of protocols is now available for the bottom layers.

doubt as to whether the processing imposed by the protocols will still permit real-time applications such as digitized voice or compressed video-conferencing — services that have been part of the promised land of LAN. The question still remains whether or not electronic mail belongs within the ISO OSI structure, outside of it as a separate service, or somewhere in between. It is well known that few true distributed database products are available today. Not to be confused with file sharing, these systems involve similar databases that are geographically dispersed. The intent is to provide each LAN containing such a database with updated information in real time. The problems associated with finding, updating, and verifying this information at several locations calls for traffic-intensive communications not yet compatible with the ISO OSI model because of the current lack of "intelligent" software.

The hardware-oriented issues with packet communications seem small when compared with the above issues; progress remains to be made in frequency division multiplexing technologies required for broad-band optical communications and in the design of fiber-optic power splitters (the optical equivalent of RF-power splitters). At light frequencies the process becomes impractical because of temperature instabilities. Consequently, achieving broad-band communications over fiber optics remains a laboratory experiment. Fibers are usually used in base-band modes; applications using Bragg technology are possible, however, in real time broadband-to-base-band converters (patent pending by the author and Honeywell, Inc.).

acknowledgements

My thanks go to Pat Snyder, WA0TTW, of the University of Minnesota Computer Center, and board member of the Tucson Amateur Packet Radio Corporation, for carefully reviewing this work, and the Honeywell Corporation, which provided time to make it possible.

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Polyvinylidene fluoride (PVF_2) is a flexible polymer film that exhibits both piezoelectric (mechanical-electrical) and pyroelectric (heat-electrical) activity. Unlike the conventional ceramic piezoelectrics found in microphones and record player pickups, PVF_2 film is thin and flexible and can be cut to any desired shape. PVF_2 film has been used in applications ranging from microphones, headphones, and keypads to transducers that reduce the buildup of algae on ships' hulls.

Seeing the potential for its use in both Amateur Radio and non-Amateur related applications, I recently spent $35.00 for a sample package of PVF_2 film and began experimenting. This article relates what I've learned over the course of my experiments.

how it works

To understand PVF_2 film it's necessary to examine its construction. **Figure 1** represents a single sheet of PVF_2 film. Its thickness varies from 0.240 mil (6 μm) to 4.4 mil (110 μm), depending upon the manufacturing process used. The nickel metallization layer is only 0.002 mil (500 Å) thick.

As **fig. 1** shows, the molecules in a PVF_2 sheet are arranged in a zigzag pattern much like the folds of an accordion. Two fluorine atoms and one carbon atom form a molecular dipole, causing a net cigar-shaped charge redistribution within that region. Likewise, two hydrogen atoms and one carbon atom form a molecular dipole with an opposite charge redistribution. Following the pattern, the dipoles are arranged in a head-to-tail fashion throughout the entire length of the material. Collectively, a net electric field polarization exists within the material in the direction indicated. It is the physical movement of these dipoles that cause a voltage to be produced or a physical displacement to occur whenever a voltage is applied. Because *voltage and molecular motion are interrelated in PVF_2* film, the material can be stretched to produce a voltage or a voltage can be applied to the material to cause it to move.

When PVF_2 film is stretched in the direction shown in **fig. 1**, a voltage with the indicated polarity is shown for this particular action. As the material is relaxed, the molecular dipoles return to their original position and a voltage of opposite polarity exists until the material once again stops moving. For PVF_2 film, the voltage output per unit displacement is typically 65×10^7 V/m (volts per meter); thus a molecular displacement of 0.0004 mil (0.01 μm) (a typical value) produces a voltage output of 6.5 volts.

stretching speed determines output voltage

One topic that requires further attention is that of the apparent "rate effect" present in producing voltage from mechanical stress. Although a voltage is produced by physically moving the molecular dipoles, the charge present in the material itself is mobile and redistributes itself to maintain an overall charge neutrality. To produce a substantial voltage output it is necessary to rapidly move the material along its stretch axis by applying a quick mechanical impulse. The voltage output is greatest for a quick, sharp tug (maximum impulse) and lowest for a gradual pull. Also, if the material is relaxed quickly, a sharp, opposite-polarity pulse is generated. The voltage output which results from the mechanical displacement is related not only to the actual displacement but also to the *rate* of displacement. The voltage output is not a steady DC voltage; rather it is a voltage pulse occurring initially with one polarity and then (upon relaxation) the other.

Figure 2 illustrates the effects of these actions. In this case, the material is pulled taut along the stretch direction and is maintained under tension, much like the material stretched over the head of a drum. The initial positively-sloped line at the left hand side of the figure (between points 1 and 2) indicates the maximum impulse region and thus the voltage produced is a maximum. Between points 2 and 3 the force is con-

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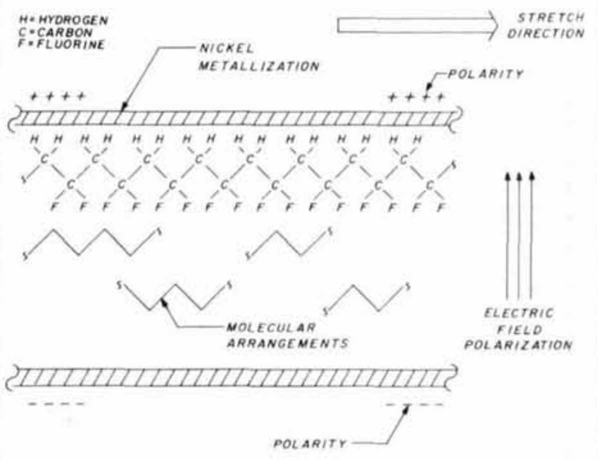


fig. 1. Cross-section of a single sheet of PVF_2 film. Nickel metallization provides electrical contact for the material. Zig-zag chain of molecules is responsible for the material's properties. Polarity is shown for one stretch direction. Note the polarity of the electric field.

stant and no output voltage is produced since charge redistribution has occurred. The declining impulse slope at the right hand side of the figure (between points 3 and 4) produces a corresponding negative-going voltage pulse.

Since voltage pulses, rather than a static DC voltage, are produced when PVF_2 film is stretched, the voltage pulses must be observed on an oscilloscope. Using one, I observed peaks as high as ± 10 volts when a sample of 52-μm thick material was given a sharp tug. The output impedance of the material is quite high (over 10 megohms) since the actual current flow in the material is extremely small.

PVF_2 sheets are available in various thicknesses. Certain thicknesses are better suited for specific applications: for *mechanical-to-electrical* conversions, *thick*, uniaxially stretched material works best because the greater the material's thickness, the greater the voltage output. For *electrical-to-mechanical* conversions, *thin*, uniaxially stretched material is preferred because the thinner the material, the greater the mechanical displacement.

PVF_2 is easy to work with

To cut PVF_2 film to the required dimensions, begin by taping the edges of the sample flat. Using a metal straightedge as a guide, cut the film with an X-ACTO™ knife, using a steady motion. *Don't use scissors*; they tend to deform the cut edges.

Electrical connection to the material can be made in several ways. The method I used, certainly one of the most economical, is to use a copper foil tape with conductive adhesive. I used a 3M® tape, No. 1181, which sells for approximately $12.00 for a roll 18-yards

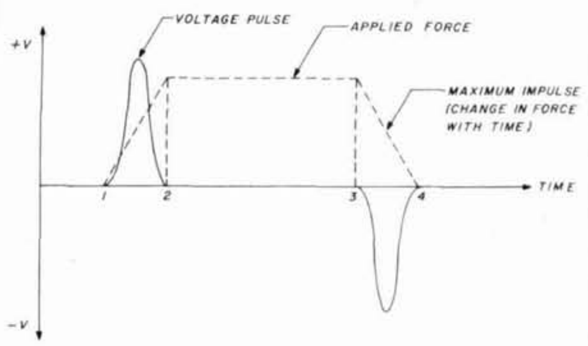


fig. 2. Rate effect of stretched PVF_2 film. Voltage pulses are maximum for maximum rate of change in applied force. No voltage output exists for a steady-force condition.

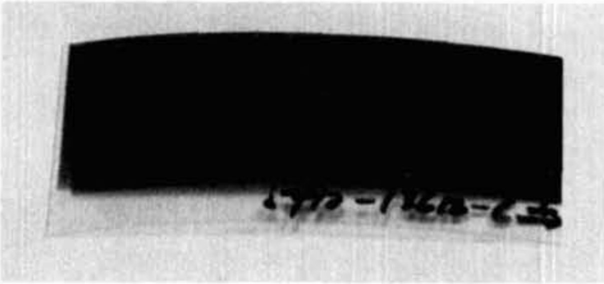


fig. 3. A sample of 110 μm PVF_2 film. The serial number indicates the film type. Note the + sign and the right arrow indicating the instantaneous polarity and stretch direction of the top metallized surface.

(16.2 meters) long and 3/8-inch (0.94 cm) wide. Because very little foil is required, a single roll can be used for many applications and should last for quite a while.

I found that No. 30 wire-wrap wire was the easiest to use because it is thin enough not to interfere with any mechanical actions and is very flexible. When soldering, don't attempt to solder directly to the PVF_2 film; this will destroy the material and could liberate toxic hydrogen fluoride gas. Instead, first solder the wire to the copper foil tape and then attach the foil tape to the material. Generally it's best to tin the foil tape first and then bring the wire to the tape. Once the lead is connected to the foil tape, remove the protective backing and apply the foil tape to the material, using finger pressure. While the tape's adhesive is quite strong when the tape is pulled along its length, it can be carefully peeled off if necessary.

Keep the wire leads at a 90-degree angle to the material's "stretch direction." In some instances I found myself tugging on the wires and generating a voltage when I was not expecting it. *The right angle*

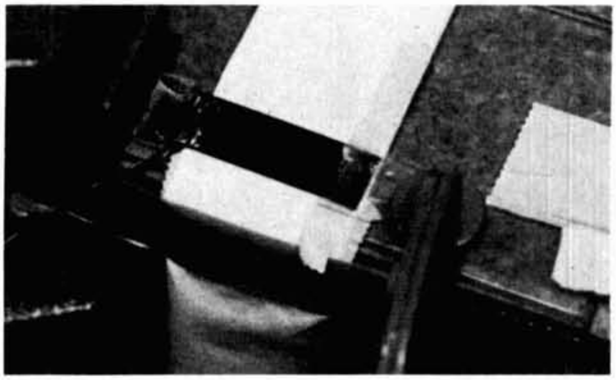


fig. 4. View of the keyer paddle arrangement. Note that the wire leads are oriented 90 degrees to the material's stretch axis. The facial tissue acts as a shock absorber. The two pieces of drafting tape are visible at the right edge of the material.

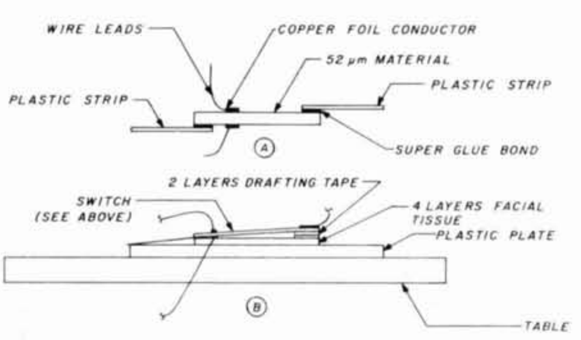


fig. 5. Side view of keyer paddle arrangement. (A) shows switch (note that plastic strips are glued at top and bottom for maximum elongation); (B) shows sidetails of entire set-up. The facial tissue acts as a shock absorber; drafting tape elevates one end of film.

approach prevents force from being applied along the stretch axis.

Figure 3 illustrates a typical sample of the material. Note the serial number indicating the film's thickness (110 μm in this case) and the markings on the lower right hand corner. The + sign indicates the positive polarity side of the material; the stretch direction is indicated by the arrow at the right. When cutting, be sure to keep track of polarities and stretch directions at all times. I found it best to use drafting tape (such as Scotch® brand No. 230) to label and hold down all cut pieces because the tape adheres well but can be easily removed without wrinkling the film. Masking tape is too sticky and too difficult to work with, especially for thin samples of material.

Morse code keyer paddle

After experimenting with PVF_2 film for a few months I decided it was time to try to produce something useful. Intrigued by a sketch that showed the positive- and negative-going pulses of a piece of stretched film when it is struck, I immediately thought of using the film in a Morse code key or keyer paddle. With the material, a lightweight, nonmechanical device could be constructed to fit any size constraint. Its performance could be varied through electronic, rather than mechanical adjustments. Because I didn't have enough spare film to construct the keyer paddle I had first envisioned, I decided instead to build a workable circuit and mechanical configuration that could be easily modified and extended by any Amateur interested in pursuing this topic.

Because thicker PVF_2 films are better suited for mechanical-to-electrical conversions than thinner films, I decided to try working with the 52 μm piece supplied with the sample package. **Figure 4** shows the experimental keyer paddle (actually one side of a keyer paddle arrangement) and **fig. 5** shows construction details.

A strip of 52 μm material 1-1/8 inch (2.81 cm) long and 1/2 inch (1.25 cm) wide was glued with Super Glue® (an Eastman 910® type of adhesive) to two plastic strips cut from an old blister pack. I found that only a small amount of Super Glue is needed to bond the plastic to the material and that air is necessary for proper drying. Facial tissue is used as a shock absorber to prevent damage to the underside of the film. The two layers of drafting tape raise one end of the material to allow room for movement. The plastic plate serves merely to protect the table from the vise and C-clamp used to secure the prepared 52 μm strip. The wire-wrap wires are then routed to the paddle's electronic circuitry.

Because the keyer paddle produces only positive and negative pulses as it is depressed and released, the electronic circuit shown in **fig. 6** is needed to produce the steady key-up and key-down conditions necessary for Morse code transmission. The circuit amplifies and buffers the input transitions, detects the polarity of the transitions, and sets or resets a flip-flop at the beginning and end of a keying interval. A 555 timer serves as an audio oscillator to monitor the keying.

U1 is a LF353 BIFET op amp characterized as having an extremely high input impedance and an extremely low input offset current. Because PVF_2 film is unable to produce even moderate amounts of current, it is necessary to use the BIFET-type op amps with their low input offset currents. At first I tried using a conventional 741 op amp and found that both the input impedance and input offset current were incompatible with the material. R1 and R2 set the gain of U1A to approximately −12.2, and U1B uses R3 and R4 to set its gain at approximately +121. Resistors R5 through R9 adjust the op amp's offset voltage. Adjust R7 and R8 for zero volts at the junction of steering diodes CR1 and CR2 with no input applied.

U2 is an LM339 voltage comparator. U2A acts as a detector for positive transitions and U2B as a detector for negative transitions. The respective voltage thresholds are set by resistors R12 and R15. By trial and error I found that setting the voltages at pins 4 and 11 of U2 to + and − 1.5V seemed to work best for keyer applications. In general, I found that the threshold voltages should not be set below 0.6V because if they are, the circuit will become susceptible to false triggering, either through keyer paddle vibrations or through other electrically-induced signals. Because the thresholds are completely variable, it's possible to vary the paddle's "weighting" or how hard the film must be struck to produce the required voltage pulses. Compare this to conventional mechanical keyer paddles where the "feel" is almost impossible to vary and the advantages of using PVF_2 film are immediately obvious.

U3 is a LM1489 EIA-to-TTL level converter commonly used for RS-232 applications. It serves here as a bipolar-to-unipolar converter for operating the logic circuits of U4 and U5. Because U3's output is at a TTL level, R16 and R17 are needed to pull up the output voltage during a logic "1" condition.

U4A and U4B are configured as inverters to compensate for U3's logic inversion. U5A acts as an SR flip-flop, set by positive-going transitions at the material's output leads and reset by the negative-going transitions once the material is released. Because the transition polarities are dependent upon the direction the material is stretched, it may be necessary to either reverse the input leads or turn the piece of material over in order to find the correct triggering polarity. (If your keying pattern sounds "inverted," you'll know that the wires need to be switched.)

U6 acts as an astable multivibrator when the top of R19 is at 5V due to U5A pin 1 being held at a logic "1." R19, R20, and C5 form the IC's timing circuit and determine the tone's frequency. R18 acts as a pull-up and load resistor, and C6 acts as a coupling capacitor. If you're going to be driving other circuits, U6 is not needed.

When working with the BIFET op amp, keep all component leads as short as possible. Keep the connecting leads from the keyer paddle shielded and as short as possible. Given the high input impedance of the BIFET op amps, it is very easy to turn the keyer paddle into an antenna if you're not careful. Here again, the variable comparator thresholds do give you some freedom in preventing false triggering. Use whatever bypass capacitors are required for your particular application. Remember, keep the op amp bypassed!

Once the circuit was breadboarded, the keyer paddle leads were connected. When I first touched the material's surface I started to pick up a 60 Hz hum, indicating that some type of isolation would be necessary. Since the keyer paddle was partially obscured by the vise and C-clamp I was forced to use a cotton swab as a cushioned striker for the material. Its cotton tip protected the films' surface from scratches and puncture, but was sufficiently pointed to produce an adequate impulse for keying purposes. Although holding the cotton swab was slightly awkward, I was able to realize reliable keying at speeds of 10 to 15 WPM. In a normal keyer paddle configuration a "clean" layout should allow you to operate the paddle as you would any other keyer. In an actual keyer paddle it would be best to use a piece of protective plastic over the paddle's surface to prevent scratching the material and to place a "dimpled" or rounded object between the plastic sheet and the keyer paddle to simulate the impulse actions of the cotton swab.

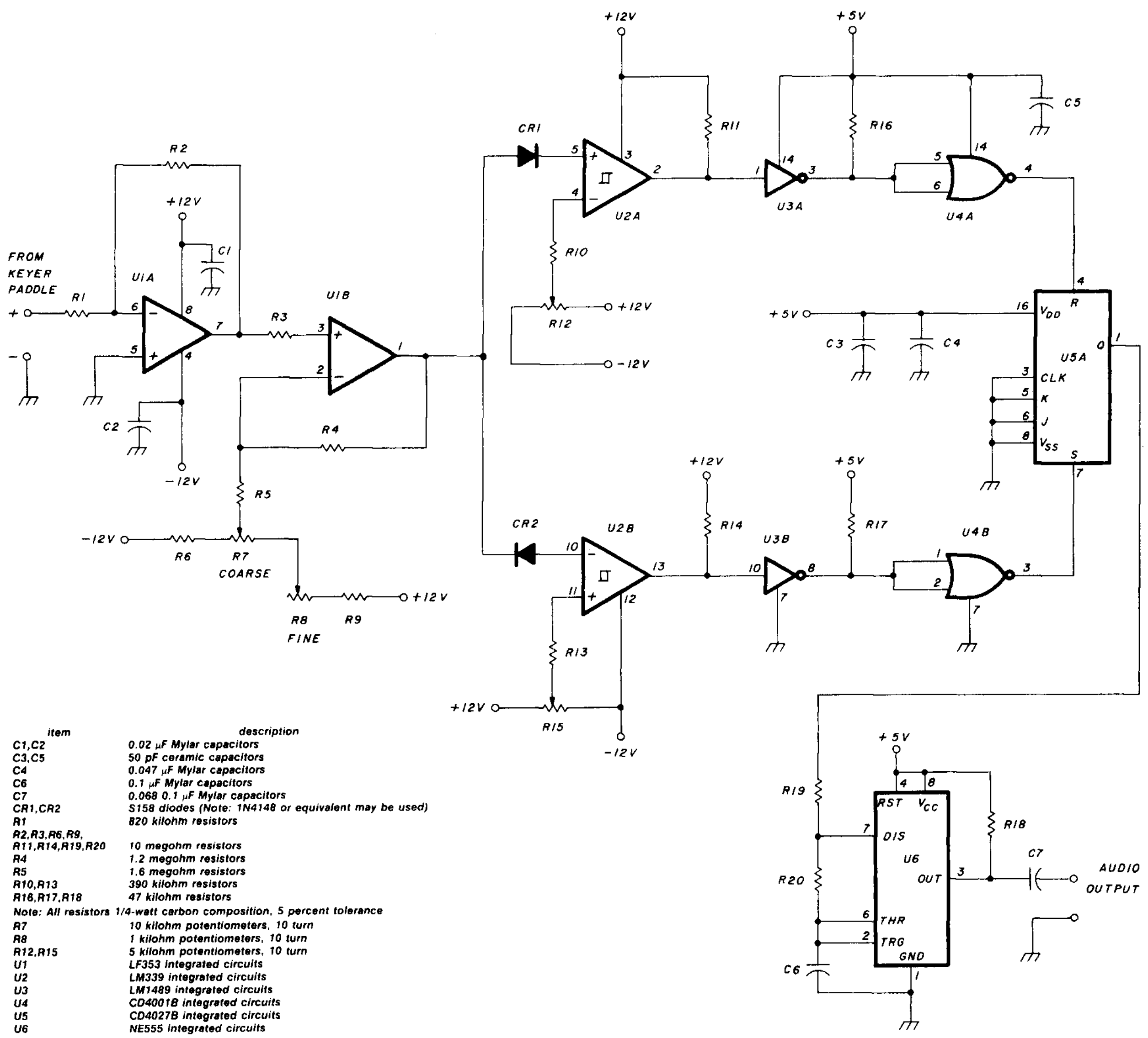
I did try using a sample of 110 μm material for the keyer paddle but found that, in general, it was too sensitive for keyer paddle applications. In fact, the 52 μm sample used in the keyer paddle was so sensitive (with 0.6V thresholds) that I could knock on the table and trigger the circuit to the "on" state. I'm sure you'll find 52 μm material to be more than sufficiently sensitive for any keyer paddle applications you may want to investigate.

film works as heat sensor, too

Although mechanical-to-electrical and electrical-to-mechanical conversions have been my primary topic for research, it is also possible to use PVF_2 film to sense infrared (heat) radiation. In this case the material produces a voltage when a changing heat source is brought near the material. Waving a match in front of a sheet of PVF_2 will cause the film to produce a voltage. Should you decide to experiment with PVF_2 film, it is important to remember not to burn the PVF_2 film or expose it to temperatures in excess of 572 degrees F (300 degrees C) because toxic hydrogen fluoride gas will be liberated.

further applications for ham radio

I hope by now I've gotten your mental motor running and you're thinking about other possible applications for PVF_2 film. As you ponder, keep in mind the cost of the material and the type of mechanical-electrical conversion you want to perform. Because the cost of PVF_2 film ranges from 17.5 cents/cm^2 for 9 μm material to 75 cents/cm^2 for 110 μm material, the average ham might find large sheets of this material too expensive for use in any one application.



item	description
C1,C2	0.02 μF Mylar capacitors
C3,C5	50 pF ceramic capacitors
C4	0.047 μF Mylar capacitors
C6	0.1 μF Mylar capacitors
C7	0.068 0.1 μF Mylar capacitors
CR1,CR2	S158 diodes (Note: 1N4148 or equivalent may be used)
R1	820 kilohm resistors
R2,R3,R6,R9, R11,R14,R19,R20	10 megohm resistors
R4	1.2 megohm resistors
R5	1.6 megohm resistors
R10,R13	390 kilohm resistors
R16,R17,R18	47 kilohm resistors
Note: All resistors 1/4-watt carbon composition, 5 percent tolerance	
R7	10 kilohm potentiometers, 10 turn
R8	1 kilohm potentiometers, 10 turn
R12,R15	5 kilohm potentiometers, 10 turn
U1	LF353 integrated circuits
U2	LM339 integrated circuits
U3	LM1489 integrated circuits
U4	CD4001B integrated circuits
U5	CD4027B integrated circuits
U6	NE555 integrated circuits

fig. 6. Schematic diagram of the keyer paddle circuit. The paddle's output signal is amplified by U1, threshold-compared by U2, converted to unipolar logic levels by U3, inverted by U4, and held in a key-up or key-down condition by U5. U6 acts as an audio oscillator.

Take heart; most applications require only a small amount of material, so one purchase should last quite some time. For larger purchases, sharing a purchase with a friend or with a club might make acquisition more affordable. The 52 μm keyer paddle described in this article required a piece measuring only 0.56 inch^2 (3.5 cm^2), at a cost of \$1.23. Because the material is fairly new, the cost per cm^2 should decrease steadily with time.

For hams who have projects in which space and weight are important (such as QRP rigs and hand-held radios) the advantages of using PVF_2 film in microphones or speakers may more than offset the cost of the film. In fact, for QRP applications, for example, where CW is the primary medium of communications, a PVF_2 film speaker could be built inexpensively. This speaker would be physically small and lightweight and would consume very little power. It could be custom-tailored to operate over any desired band of audio frequencies. Other applications such as "do-it-yourself" keypads for equipment are also possible.

PVF_2 film has been used in transducer applications at frequencies up to 9.36 GHz. One can easily envision different types of antennas and RF pick-up schemes that Amateurs could easily build. The literature does make reference to vidicons or video "tubes" for cameras using PVF_2 film pieces to form an imaging area so SSTV and ATV enthusiasts may wish to read more about this. Believe me, I could easily ramble on about potential applications for pages on end, but my advice to you is read all of the applications literature available and try your own experiments.

Applications literature for PVF_2 film is available from KYNAR Piezo Group, Pennwalt Corporation, 900 First Avenue, P.O. Box C, King of Prussia, Pennsylvania 19406-0018.

acknowledgements

Special thanks to J. Victor Chatigny at Pennwalt for his comments and assistance and to my friend Bill Drake, for his help with the photos.

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Garth Stonehocker, K0RYW

absorption affects signal level

Last month's column included a discussion of the effect of absorption on signal levels. Under *normal* conditions a transmitted wave leaving the antenna at an effective radiated power level of 1000 watts, traversing the atmosphere twice and the ionosphere once, might arrive at the receive location over 170 dB down (weaker) and still be copyable. During the winter a lower absorption level and a quieter (or undisturbed) geomagnetic field results in enhanced conditions for the lower frequency bands. Occasionally, however, a condition called the *winter absorption anomaly* appears to reverse the trend.

This anomaly causes higher than usual attenuation of signals crossing a 500 to 2500 km wide strip between 60 and 70 degrees North latitude (or more precisely, 40 to 55 degrees *geomagnetic)*. The region is known as the mid-latitude trough, and the effect lasts for several days at a time. The trough is an actual depression in electron density of the F region and a higher than usual level of D and E region ionization. It is the latter condition specifically that accounts for greater wintertime absorption levels on the lower bands. One can expect it to occur from November through February and to be more severe during times of greater geomagnetic activity. However, there is no *short* term (i.e., hour or day) relationship to solar flux or geomagnetic activity indices. The extra ionization is related to electron particles entering the auroral zone. No applicable data collection or warning system is currently available for forecasting the anomalous absorption.*

last-minute forecast

The 27-day solar (activity) cycle variations which have increased over those of the summer and fall, are expected to be even greater in December. Look for a maximum in activity to occur on approximately the sixth of the month, with a corresponding minimum on the twentieth. A solar flux of 80 represents an average value for this part of the solar cycle. The effect of this increased solar activity means that the 10 through 30 meter bands will be useful for communicating between North America and South Africa, South America and Australia over trans-equatorial paths during the first two weeks in December. The latter weeks will favor the lower frequency bands — 30 through 160 meters. Tune in to WWV at exactly 18 minutes after the hour to receive their daily broadcasts of solar activity and geomagnetic field information useful in determining HF/VHF band conditions. Also during this month, both atmospheric noise and ionospheric absorption will be down, resulting in useful *daytime* DX conditions for the 40 and 80 meter bands.

The Geminid meteor shower, which reaches its peak on December 13-14, provides the richest and most reliable display of the year, with rates of 60-70 per hour (measured mainly by radio because of the poor weather in December). Also, a smaller portion of the shower (15-20 per hour) is observed on December 22. Lunar perigee and full moon occur on December 7th and 8th, respectively. Winter solstice — the longest night — occurs on December 21.

*Reports of this absorption would be useful to me. Please write to Garth Stonehocker, Route 2, Box 45, Earlysville, Virginia 22936.

band-by-band summary

Ten, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of the bands will be shorter and will occur closer to local noon. Trans-equatorial propagation on these bands will be more likely towards evening during conditions of high solar flux and a disturbed geomagnetic field.

Thirty and forty meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 20 meters but shorter skip and signal strength may decrease during midday on some days — days coinciding with high solar flux values or anomalous absorption. Nighttime use will be good except after days of very high MUF conditions. Generally the usable distance is expected to be somewhat greater than that achieved on 80 at night.

Eighty and one-sixty meters are the nighttime DXer's bands, especially this time of year. The bands open just before sunset and last until the sun comes up on the path of interest. Except for daytime short-skip signal strengths, high solar flux values don't affect these bands much. Some days the anomaly will affect day and night signal strength.

DECEMBER

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	30	30	40*	40	40	40	40	40	40	40	40	30	30	30	30	30	30	30	20	20	20	20	20	30	N
EUROPE	30	30	30	30	30	30	30	30	30	30	40	40	40	40	40	40	40	40	40	30	30	30	30	30	NE
S. AFRICA	15	15	15	15	15	15	15	15	15	20	20	20	30	30	20	20	20	20	20	20	20	20	15	15	E
S. AMERICA	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	15	15	15	15	10	10	10	10	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	15	S
NEW ZEALAND	10	15*	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	15	10	10	SW
OCEANIA AUSTRALIA	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	10	10	15	W
JAPAN	20	20	20	20	30	30	40	40	40	40	40	40*	30	30	30	30	30	30	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	30	30	20	20	20	20	20	20	30	30	30	30	30	40	40	40	40	40	40	30	N
EUROPE	30	30	30	30	30	30	30	30	30	30	30	40	40	40	40	40	40	40	40	30	30	30	30	30	NE
S. AFRICA	15	15	15	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	E
S. AMERICA	15	15	15	10	10	10	10	10	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	15	S
NEW ZEALAND	15	15	15	15	15	15	15	15	20	20	20	30	30	20	20	20	20	20	20	20	15	15	15	15	SW
OCEANIA AUSTRALIA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	W
JAPAN	20	20	20	20	20	30	30	40	40	40	40	30	30	30	30	30	30	30	30	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	30	30	30	20	20	20	20	20	30	30	30	30	40	40	40	40	40	40	40	40	40*	N
EUROPE	30	20	20	20	20	20	20	30	30	30	30	40	40	40	40	40	40	40	40	30	30	30	30	30	NE
S. AFRICA	15	15	15	15	15	15	15	15	15	15	15	20	20	30*	20	20	20	20	20	20	20	20	15	15	E
CARIBBEAN S. AMERICA	15	15	10	10	10	10	10	15	15	15	15	15	20*	20	20	20	20	20	20	20	20	20	20	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	S
NEW ZEALAND	15	15	15	15	15	15	15	15	20	20	20	30*	20	20	20	20	20	20	20	20	15	15	15	15	SW
OCEANIA AUSTRALIA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	W
JAPAN	20	20	20	30	30	30	40	40	40	40	40	40	40	40	30	30	30	30	30	30	30	20	20	20	NW

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

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Bill Orr, W6SAI

huff-duff revisited

"Huff-duff" is a World War II term for *high-frequency direction finding,* a technique vital for locating enemy transmitting sites. A considerable amount of money and energy were expended in maintaining this essential service.

HF-DF (the modern term) was developed about 1938, principally by the Australian Ministry of Civil Aviation. Many direction finding stations were installed on that continent. But as aeronautical radio communication in the United States gradually shifted over to VHF bands, the extensive use of high-frequency direction finding became unnecessary. America's entry into World War II, however, changed this, and accelerated development of new equipment was quickly undertaken by the FCC and the United States Army Air Corps. After the war, some of this equipment showed up in the surplus markets as the SCR-291, which was largely ignored by Radio Amateurs.

Amateur direction finding today

Interest in HF-DF is slowly growing among Radio Amateurs. Those who experiment with this interesting hobby soon learn that a new set of ground rules applies, especially with regard to long distance ionosphere-reflected signals.

Direction finding involves the determination of a *line of direction* between a receiver and a distant transmitter. The intersection of two lines of direction from widely separated receivers establishes the location of the transmitter. This technique works well on ground wave signals, but "skip" signals reflected from the ionosphere undergo alterations that greatly increase the difficulty of determining a line of direction. As long as ground wave signals are used for direction finding, accurate bearing readings are relatively easy to obtain. DF work has been accomplished with good results on VHF (the 2-meter band, for example) and on ground wave signals on the 160-meter band. But Amateurs have had little experience with DF work on the DX bands with long-distance signals.

DF principles

The loop antenna is the device that forms the basis of all low- frequency DF work, such as demonstrated by the radio ranges operating in the 200 kHz to 500 kHz portion of the spectrum. Nondirectional (transmitting) ground stations are used as a signal source on which bearings are taken by a special receiving antenna. In most instances, a simple loop antenna is used (**fig. 1**). If the dimensions of the loop are small compared to wavelength, the pickup pattern of the loop is as shown in **fig. 2**. Maximum pickup occurs when the radio wave travels parallel to the plane of the loop, cutting first one side then the other. If the loop is at right angles to the direction of wave travel, the radio wave cuts both sides of the loop simultaneously and the resultant output voltage of the loop is zero. The resulting pickup response of the loop is thus a figure-eight pattern and there are two loop positions for which the output is zero and two for which it is a maximum. Note that the loop pattern is opposite to that of a Quad loop, or other large loop, wherein the plane of maximum signal response is at right angles to the plane of the loop.

Note also that the nose of the loop pattern is extremely broad and the pattern nulls are very sharp. It is the pattern nulls which are useful for direction-finding work.

Operation of the loop direction finder is simple. The receiver is tuned to the desired station and the loop is rotated until the signal disappears. The line of direction is then compared with either magnetic or true north, and the results plotted on a map. Two such bearings, from separated locations, constitute a *fix*, or the location of the signal source.

the HF-DF problem

The most important alteration to the radio wave as it is reflected from the ionosphere is a change in polarization. The electric field of a radio wave exerts a force on electrons in the ionosphere, in addition to the force exerted on them by the magnetic field of the earth. The resulting action of the two forces causes an elliptical electron motion that causes the polarization of the exciting wave to be elliptical. Measurements made in 1931 showed that regardless of the initial polarization of the transmitted wave, the reflected wave polarization was elliptical for medium distances, with the possibility that polarization could be elliptical or horizontal in the case of long- distance waves striking the earth at a low angle of incidence.

In addition to polarization changes, the reflected radio wave exhibits fading, probably as a result of two or

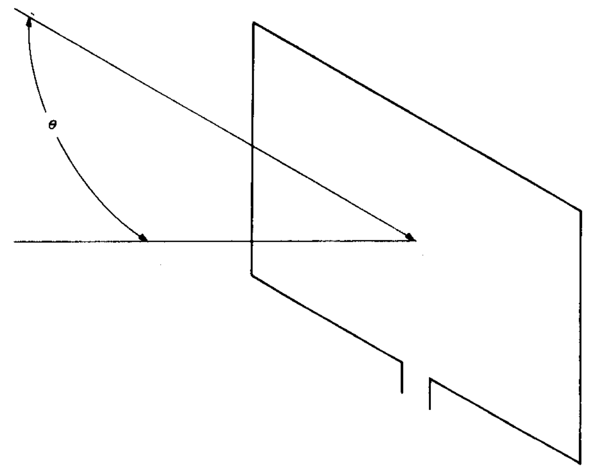

fig. 1. Output voltage of small loop is proportional to the cosine of the angle θ between plane of the loop and signal vector.

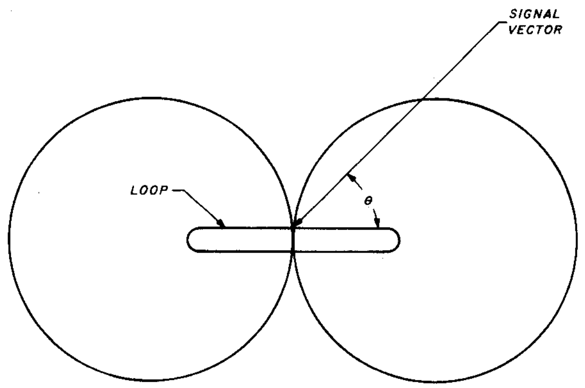

fig. 2. Pickup pattern of small loop. Note nulls at right angles to plane of loop. Maximum pickup occurs when signal vector is parallel to plane of loop.

more reflection paths having slightly different lengths. The different paths come about because of irregularities in the refracting layers of the ionosphere. More often than not the paths are sufficiently separated so that the received signal seems to be coming from two separate directions.

deviations from the great-circle path

It is widely assumed that long-distance signals follow a great-circle path from transmitter to receiver. Amateurs frequently place considerable faith in great-circle maps when aiming their directive antennas at an exotic DX station. Experienced operators, however, have long known that this confidence is not always justified. With a small beam antenna having a "fat" forward lobe, azimuthal deviation of the DX signal is of little importance. Users of highly directive beams, on the other hand, have found that deviations of 10 to 20 degrees from the great-circle path of DX signals are common over short periods of time, and deviations as much as 90 degrees have been noted for signals appearing from the antipodal area of the world.

In the case of sporadic-E layer reflected signals, reflection can occur from many "clouds" of reflective regions and multiple echoes are present on a signal propagated in this fashion. Experimenters have noted that bearing deviations up to 20 degrees are present on echo signals.

the loop antenna for DF work

"Top Band" DXers have used small loop antennas for reception for many years. Generally speaking, the loop is not used to DF the received signal, but to null out local interference (such as power line noise, or TV oscillators) to achieve improved signal-to-noise ratio. *However, most users of loop antennas have found out that they are useless as direction finders for ionospherically reflected signals.*

I have used a simple loop antenna for tracking thunderstorm interference on 160 meters, but the directional nulls were broad and pointing only to a very general area of thunderstorm activity.[1] No success was had DFing long-distance signals on this band.

the Adcock DF antenna

In 1919 a patent was granted to F. Adcock, an Englishman who developed an HF direction finding antenna that was not susceptible to polarization error of the reflected radio wave. The antenna was sensitive only to the vertically polarized component of the wave (**fig. 3**). Amateurs will recognize this device as similar to, though smaller than, the famous W8JK beam antenna. If the spacing between antenna elements is small compared to the wavelength, the voltage output of an Adcock antenna is a function only of the height above ground and the angle between the vertical plane containing the elements and the plane of polarization of the radio wave. Until World War II, most practical direction finders used various versions of the Adcock antenna. The old SCR-291 DF equipment used a version of the Adcock antenna, plus a "sense" antenna that removed the 180-degree ambiguity caused by the Adcock array. An oscilloscope was used to view the SCR-291 output, which provided a DF signal picture as illustrated in **fig. 4.**

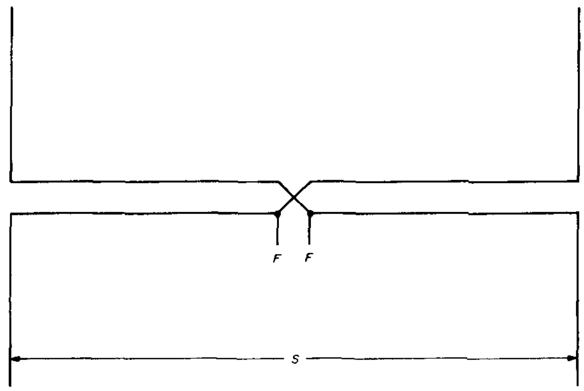

fig. 3. Balanced Adcock antenna system which responds to vertically polarized waves. Spacing S is small in terms of wavelength.

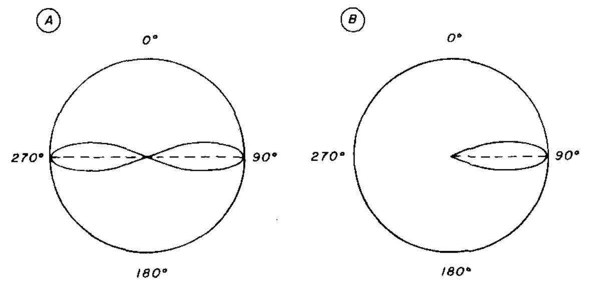

fig. 4. (A) Elongated figure-eight pattern provides line of direction in SCR-291 direction finder. (B) Sense antenna changes "leaf" pattern to cardioid to resolve signal ambiguity.

For accurate DFing, up to 100 bearings were taken at various times during a 24-hour period, with particular attention to bearings taken at sunrise and sunset. Bearings from two or more widely separated DF receivers were taken, smoothed, and then plotted on a map, thereby indicating an area in which the transmitter was located with a reasonable degree of probability.

The Adcock system has been supplanted by superior devices, but these are large and beyond the cost of the average Amateur. But when properly built and used, the Adcock antenna is a valuable addition to the Amateur interested in HF direction finding.

building an Adcock array

I've seen only one article dealing with the construction of an Adcock antenna.[2] In brief, the Adcock array consists of two vertical antennas, closely spaced with respect to wavelength and connected out-of-phase, as shown in **fig. 5**. Each antenna can be shorter than its physically resonant length, with system resonance established by placing a simple tuner at the operating position. Using a 0.1 wavelength design for a 20-meter Adcock array provides element spacing of about 7 feet 6 inches. Signal pickup depends upon element length. In order to make a compact array, 0.1 wavelength elements were chosen.

The array elements are cross-connected with a short length of open-wire TV "ladder line." The feedpoint is taken at the center of the line. A second length of ladder line leads from the array down to the antenna tuner at the operating position. Care should be taken to keep the line clear of nearby metallic objects.

The tuner consists of a resonant circuit with an inductive pickup link coupled to the receiver via a coaxial line. Balancing capacitors are connected across the tuning capacitor to enhance the signal null. It's best to locate the tuner directly below the antenna, if possible, to ensure a good signal null.

The first model antenna was built on a wood frame which proved to be rather unsteady. A new model, in which aluminum tubing is used for the framework supporting the antennas, is under way.

Antenna system resonance can be established with a dip meter or by adjustment of the tuner on a local signal. When resonance is found, the antenna is rotated for the best null and the balancing capacitors (C2,C3) are adjusted to enhance the null. The antenna exhibits a figure-eight pattern, with the nulls at right angles to the plane of the antennas.

It may be difficult to find an exact null when trying to determine a line of direction to a signal because of background noise obscuring the null. In this case, the null can be estimated by noting the position of the antenna at an equal signal position on each side of the null. The angular difference between the readings is divided in half to find the actual null.

In my case, the experimental Adcock array is turned by a small TV rotator because the wind resistance of the antenna is quite low. The array and read-out of the rotator are adjusted so that "North" on the indicator corresponds to true North (not magnetic North).

using the Adcock array

It's best to practice on local signals and one-skip signals. (I use stations in the Denver area for DF practice.) Beware of local reflections from hills and from guy wires, gutters, and other nearby metallic objects. Use a map to determine the true direction of your practice stations, and make multiple

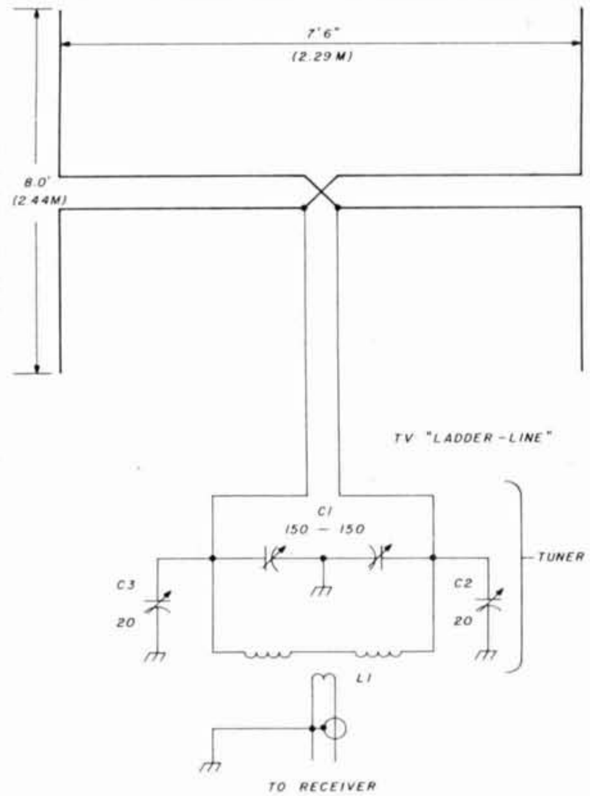

fig. 5. Simple Adcock array for 14-21 MHz operation. C1-L1 resonates system to listening frequency and C2, C3 are used to enhance signal null. (Patterned after design by W1YNC, *QST*, August, 1975, page 30.)

readings to obtain meaningful results. Remember, the loop provides an ambiguous reading with an error factor of as much as 180 degrees if you don't keep your wits about you. Pretty soon you can "spot" a station with reasonable directive accuracy before you hear its location or look it up in the Callbook.

If you have a general coverage receiver, you can venture out of the ham bands. The 15-MHz broadcast band is a good one on which to try DF experiments. A prime choice is WWV in Colorado, as are the Voice of America outlets, the BBC, HCJB in Quito, Ecuador, and Radio Japan. DFing requires a certain amount of expertise if your efforts are to produce meaningful results, but with practice you'll become more skillful. Then when 6L6GX opens up from Lower Bulltravia, you can tell your DX friends, "Don't waste time on him. His heading doesn't check with where he says he is!" (This reminds me of the story of an infamous DXer who appeared to have worked the world from many exotic locations until his adventures were finally branded as fraudulent. DXCC credit was revoked for the thousands of "new country" contacts he had provided for gleeful DXers. After the excitement had died down, he resurfaced one day, working portable from New York City. While in QSO with someone, a hard-bitten DXer zero-beat the fraud and very slowly sent: "Gee! Do you really think he's there?!")

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2. Tony Dorbuck, W1YNC, "Radio Direction Finding Techniques." *QST*, August, 1975, page 30.
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4. Peter C. Sandretto, *Principles of Aeronautical Radio Engineering*, McGraw-Hill Book Co., New York, 1942.

ham radio

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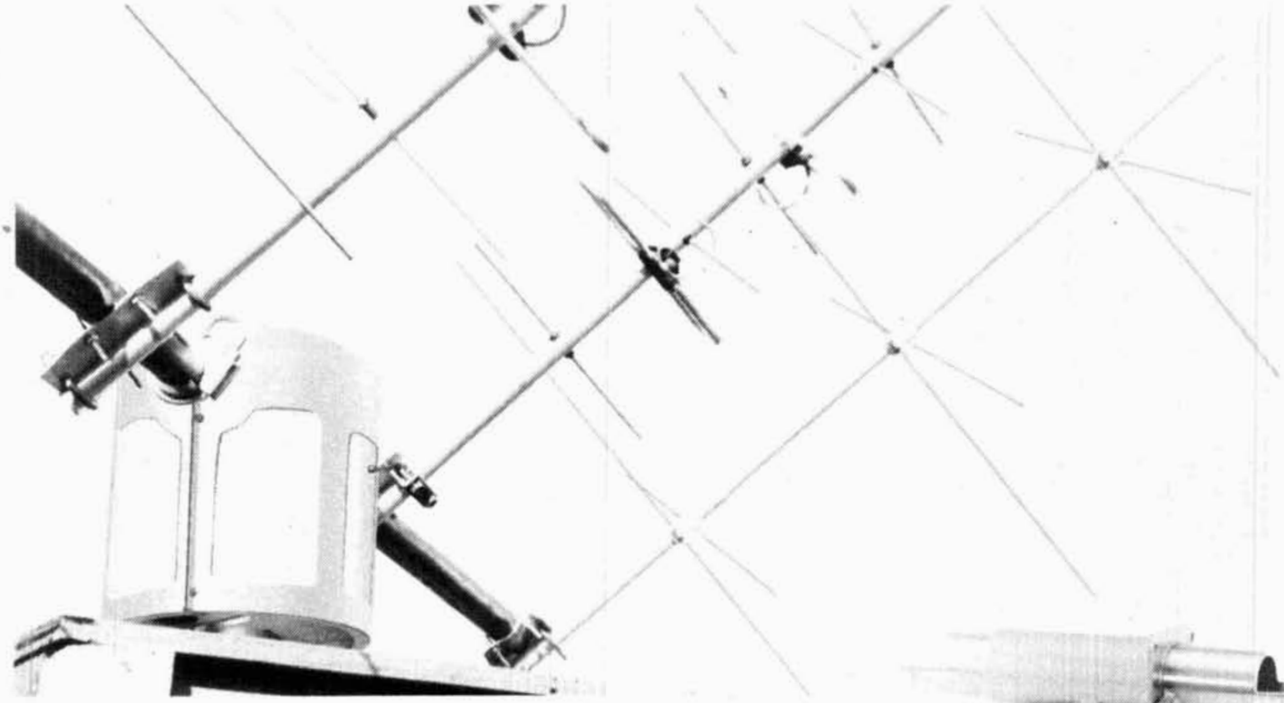

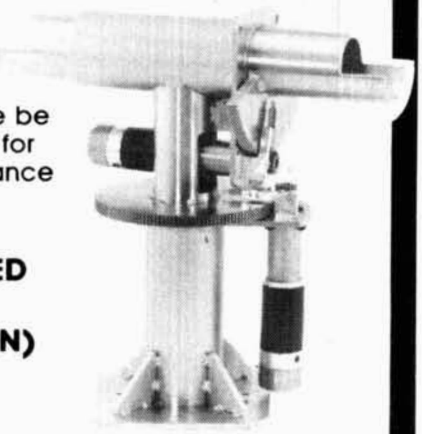

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quasi-bilateral IF for transceivers

Try this old technique to save space in new circuits

Oldtimers will remember the bilateral IF circuits that were popular some years ago. The intent of their designers was to avoid duplication of circuitry in the receiver and transmitter stages in compact transceivers. Although a few manufacturers, notably Side-Band Engineers, made use of bilateral IF stages, the technique never fully caught on with homebrewers, possibly because of the overall complexity and parts count involved with the early bipolar circuits, and the tendency towards instability exhibited by some designs. Orr's *Radio Handbook* carried a brief description of a bilateral IF arrangement similar to the SBE circuit.[1] The bilateral IF has gone out of vogue in recent years with the advent of modern IC technology, high density PC board layouts, and inexpensive semiconductors.

When I decided to build a compact 160-meter transceiver the use of such a scheme seemed ideal to conserve space. Unfortunately, all of the designs appearing in my various handbooks left much to be desired and were in fact quite dated. The final circuit that was used successfully in my project is presented here.

My objectives for a simple, effective circuit were met through the use of IC and passive doubly balanced diode ring mixer components available at low cost and small in physical size. While my circuit is not a true bilateral design, it represents a unique approach to a shared receive/transmit IF that I haven't seen used. (Benjamin Vester, W3TLN, was on this track when he described his pocket-sized solid-state 20-meter transceiver in the ARRL's *Sideband Handbook*.[2] I had chosen an IF frequency of 455 kHz to make use of some old Collins mechanical filters I found in my junkbox. Crystal filters of more recent vintage should work as well or better because of high insertion loss associated with the early mechanical filters I used. My 160-meter transceiver is single conversion, and a low IF frequency was not a concern.

The first stage of the IF, a Mini-Circuits Lab DBM, does double duty as the balanced modulator for transmit and as the receiver first mixer. During receive, the preselector output (1.8 MHz) is coupled to the IF port of the DBM while the LO port is driven by the VFO, with the 455 kHz IF resulting. During transmit the IF port is driven with amplified microphone audio for SSB operation. The LO port is driven by the BFO. Note for CW generation during transmit a DC bias on the mixer IF port allows BFO energy to feed through the mixer; of course the BFO frequency must be shifted to within the passband of the filter for proper transmitter operation. SSB audio is coupled to the IF port through an RF choke so as not to affect receive operation, while the value of the coupling capacitor from the IF port to the receiver preselector has insufficient reactance to adversely load the microphone audio.

This confirguration admittedly compromises the dynamic range of the DBM. Termination impedance of the IF port of diode ring mixers is critical and directly affects the insertion losses and dynamic range. Care was taken to provide a diplexer at the IF output (the manufacturer's designated RF port for the DBM), and with the use of a 3-dB pad at the mixer LO port to insure reasonable terminations in my IF circuit. The 160-meter transceiver I built is entirely passive (negative gain) from the antenna to the first receiver mixer, so some loss of ultimate mixer performance was permissible. The first active stage is the dual-gate MOSFET amplifier used to provide gain, matching, and filter switching. The Collins filters used (FA series) follow and were operated parallel-resonant for high impedance to match to the MOSFET amplifier output. The filter's output coils were coupled in the series-resonant mode to achieve a 50-ohm impedance, allowing efficient use of diode switching between the SSB and CW filters and to match the low impedance input of the MC1349 IF amplifier.[3] A simple circuit (**fig. 1**) allows simple selection of alternate bandwidth filters.

The heart of the IF stage is the MC1349 IC, an improved version of the Motorola MC1350. Besides the MOSFET preamp for the filter, it is the only other active component required for the IF stage. Notice that pin 5 is the AGC input for the device. In my unit I used audio-derived AGC with good results, but simple RF-derived AGC circuits have been published for this

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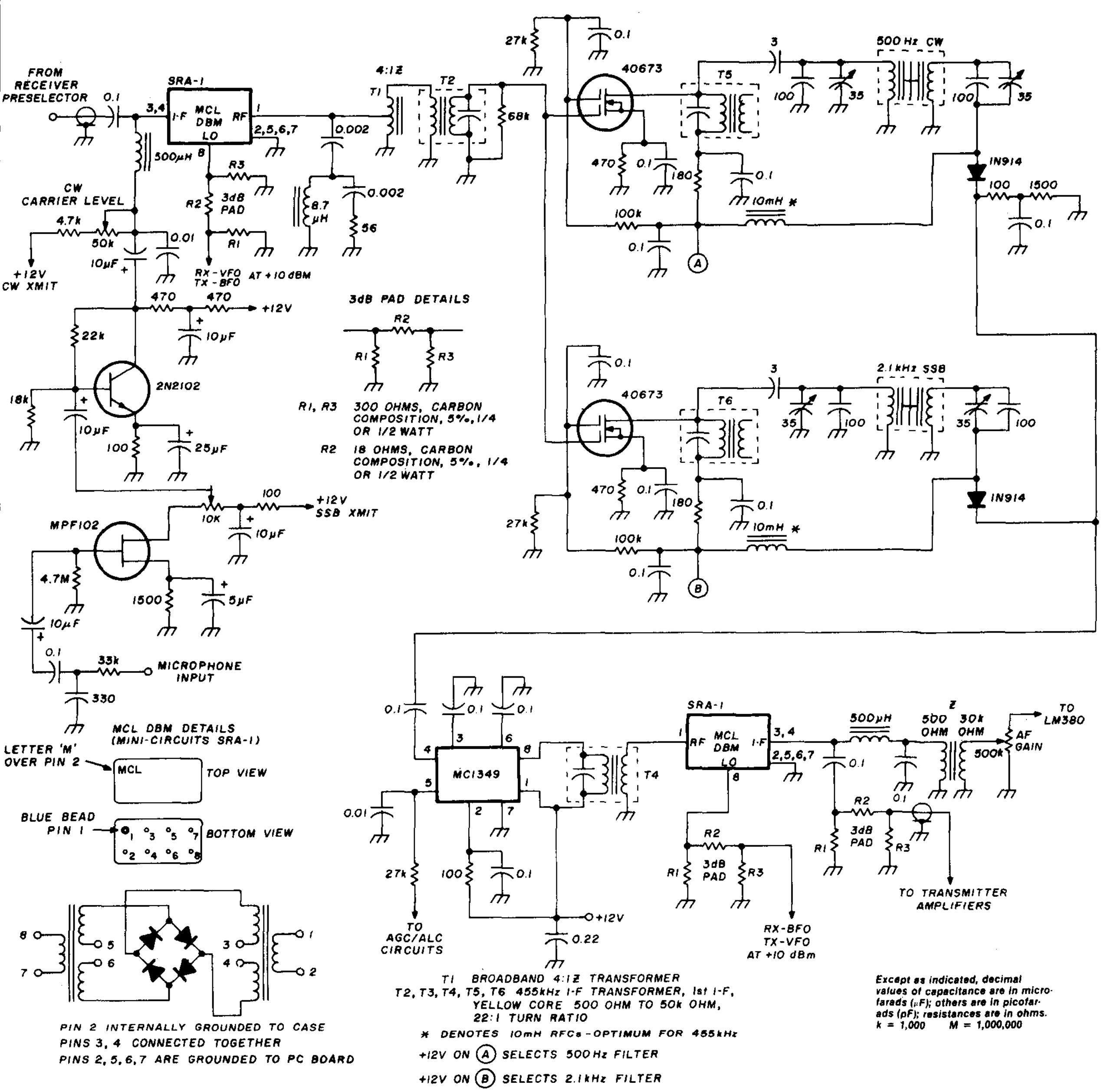

fig. 1. Quasi-bilateral IF for transceivers. Schematic shows simple method of switching between filters of different bandwidths. More filters can be added by duplicating the dual-gate FET stage for each additional filter desired. Author's transceiver incorporates one additional filter for AM reception.

device.[4] The transmitter ALC signal, also common to the AGC buss, was developed in my transceiver by diode sampling to the transmitter PA output. If an MC1350 is used, pin 3 should be grounded; for the MC1349 pin 3 should be bypassed to ground. In the MC1349, pin 3 is connected to the emitter of the input transistor and sets the device input impedance and gain. *I used pin 3 in my unit to couple the IF signal* to a separate AM IF stage and detector. The MC1349 output is transformer coupled to the second MCL DBM RF port. In receive this DBM serves as the product detector, with the BFO energy coupled to the LO port. Note that the recovered audio is through a three-element low-pass filter to remove any IF RF from reaching the audio stages.

The first bypass capacitor of the filter does not go directly to ground but is connected to a 3-dB attenuator. During transmit, the BFO is switched to the first DBM, and the VFO to the LO port of the second DBM. This arrangement allows the desired 160-meter product from the second DBM to be passed onto the transmitter preselector and amplifiers with minimum losses and good termination to the mixer port.

BFO and VFO signal switching between the LO ports may be achieved through miniature relays or by using solid-state circuitry. No relays were used in my transceiver; the preselector tracked the main VFO and was switched for common duty for both the transmitter and receiver. Recovered audio from the diode ring detector was sufficient to drive a single LM380 audio IC. The AGC range is 70 dB.

Although I haven't tried it yet, the performance of this circuit at 9 MHz should be better. The newer crystal filters will have less insertion loss, and the DBMs I used were not rated for best operation below 500 kHz at the LO and IF ports. Of course the values for the diplexer[5] and for the audio coupling RF chokes would have to be changed accordingly as well as the transformers used if another IF is chosen. While the circuitry presented here could certainly stand improvement, it was my intent only to outline my solution for a novel IF arrangement. As this circuit best lends itself to a single conversion transceiver, the use of a 455-kHz IF is probably precluded above 7 MHz to avoid image problems. The FA series Collins filters I used are the same as those used in the Collins 75S series receivers; they are expensive if purchased new, but can often be found at flea markets for reasonable prices. Several dealers regularly offer imported 455-kHz mechanical filters and surplus crystal filters in the 5-9 MHz range at very attractive prices.

While no actual gain or noise figure measurements were made, the losses between the antenna and first mixer are excessive in my transceiver. The input and output of the lightly coupled preselector are diode switched for a total path loss exceeding 8 dB. For operation on 80 or 160 meters this is acceptable if a full-size antenna is used. My objective in doing this was to allow the noise floor of the receiver to be no less than required for everyday use on 160 meters, thereby improving the capability of the receiver to handle strong signals. With a carefully designed bandpass filter, such as the low loss units designed by Wes Hayward, W7ZOI, operation of this IF on any HF band should be possible without additional front end gain.[6]

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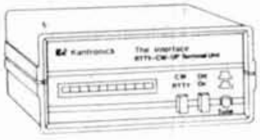

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B1016	2M	Yes	10W	160W	20A	$249
B3016	2M	Yes	30W	160W	17A	$199
C22	220	No	2W	20W	5A	$ 79
C106	220	Yes	10W	60W	10A	$179
C1012	220	Yes	10W	120W	20A	$259
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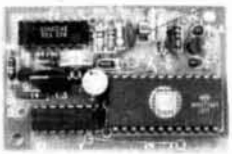

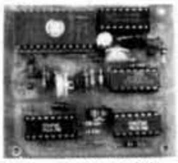

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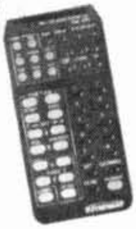

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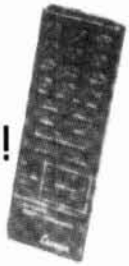

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microstrip impedance program

Use your TRS-80, HP-41, or HP-67/97 to determine stripline width and other parameters

Even though programmable calculators and computers can easily be used to perform microstrip impedance line calculations, I haven't seen many suitable programs in print.

The two synthesis programs listed here were written for use on the Radio Shack TRS-80™ computer (**fig. 1**) and the HP-67/97 or HP-41 (**fig. 2**). Both may be adapted for use on other programmable calculators and personal computers.

The programs allow the user to input a desired microstrip impedance value, frequency, relative dielectric constant and thickness to calculate the correct linewidth for that impedance. It will also output the effective dielectric constant, relative velocity factor, and 90-degree microstrip line length.

The mathematics used represent a good closed form approximation for line impedances between 20-110 ohms and will deliver results that fall within 3 percent of Wheeler's line impedance values.

Table 1 contains a list, printed out using a TRS-80, of microstrip line impedances from 10 to 100 ohms for *a 0.025 inch (6.35 mm) thick dielectric* of 10 at 2.0 GHz. It also shows the effective dielectric constant for lines, relative phase velocity, and 90 degree length.

The program listing in **fig. 1** is a short closed form approximation. No adjustments (such as a conductor thickness correction or variation of equations used at different impedance values) are made to improve accuracy. The ± 2 percent accuracy is sufficient for most applications.

table 1. Microstrip line impedances from 10 to 100 ohms for a 0.025 inch (6.35 mm) thick dielectric of 10 at 2.0 GHz.

ER = 10 Thickness = 0.025 inches Frequency = 2 GHz
Enter Z_0 impedance wanted (LOW, HIGH, STEP)?
10,100,10

impedance	width (inches)	ER-effective	velocity factor	90-degree (inches)
10.0	0.2446	8.7	0.340	0.500
20.0	0.1026	7.9	0.355	0.523
30.0	0.0572	7.4	0.367	0.540
40.0	0.0355	7.1	0.376	0.553
50.0	0.0233	6.8	0.383	0.564
60.0	0.0156	6.6	0.390	0.574
70.0	0.0106	6.4	0.395	0.582
80.0	0.0072	6.3	0.400	0.589
90.0	0.0049	6.1	0.404	0.595
100.0	0.0033	6.0	0.408	0.601

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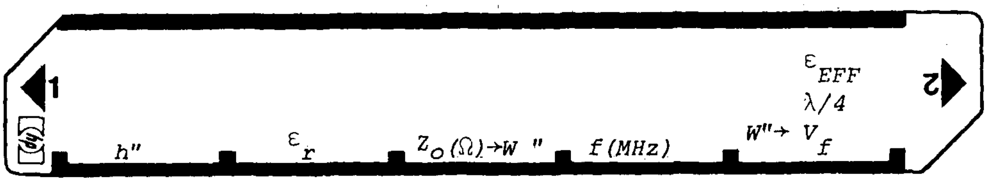
By Dennis C. Mitchell, K8UR/1, 1 Cider Mill Lane, Upton, Massachusetts 01568

```
10 CLEAR 100
20 A$="###.#":B$=".####"
30 H2=.06:ER=4.8:ZO=50
40 CLS:PRINTCHR$(23):PRINT:PRINT:PRINT:PRINT"     MICROSTRIP CALCULATOR"
50 PRINT:PRINT
60 PRINT"      D.C. MITCHELL-K8UR"
70 FORI=0TO500:NEXT
80 CLS:PRINT"              DIELECTRIC CONSTANTS"
90 PRINT STRING$(55,"-")
100 PRINT"ER OF FIBERGLASS-EPOXY P.C.=4.8"
110 PRINT"ER OF TEFLON OR SIMILAR FLUROCARBON POLYMERS=2.5"
120 PRINT "ER OF 99% ALUMINA DIOXIDE CERAMIC = 10"
130 PRINT"THE DIELECTRIC THICKNESS (H) OF 1/16'' P.C. STOCK DOUBLE CLAD W/1 OZ./SQ.FT. COPPER=.0597"
140 PRINT"FOR 1/32'' THICK P.C.B. USE .0285";:PRINT" (DEFAULT ZO=50,ER=4.8,H=.0597)"
150 PRINT
160 INPUT"ENTER THICKNESS OF P.C.B";H2
170 INPUT"ENTER ER OF P.C.B.";ER
175 INPUT "ENTER FREQUENCY IF 90 DEG LENGTH IS DESIRED (GHZ)";FQ
180 CLS:PRINT "ER=";ER,"THICKNESS=";H2;"INCHES","FREQ=";FQ;"GHZ"
190 INPUT"ENTER ZO IMPEDANCE WANTED (LOW, HIGH, STEP)";ZL,ZH,ZS
200 PRINT "IMPEDENCE   WIDTH(IN.)     ER-EFFECTIVE       VR      90DEG.(IN)"
210 FOR ZO=ZL TO ZH STEP ZS
220 W1=H2
230 W=377*H2/(SQR(ER)*ZO*(1+1.735*ER[-.0724*(W1/H2)[-.836))
240 IF INT(W*10000)/10000=W1 THEN 270
250 W1=INT(10000*W)/10000
260 GOTO 230
270 PRINT USING A$;ZO;:PRINT "        ";:PRINTUSING B$;W1;
280 EF= ((ER+1)/2 + (ER-1)/2 * (1+10*(H2/W1))[-.5)
290 VR=1/SQR(EF)
295 L=11.785/4/FQ*VR
300 PRINT "          ";:PRINTUSING A$;EF;:PRINT "          ";: PRINT USING "#.###";VR;:PRINT"     "
310 NEXT ZO
```

fig. 1. TRS-80™ microstrip line program listing.

User Instructions



STEP	INSTRUCTIONS	INPUT DATA/UNITS	KEYS		OUTPUT DATA/UNITS
1	*Input thickness of dielectric in inches*	h"	A		
2	*Input relative dielectric constant*	ε_r	B		
3	*Input desired impedance* Z_o	Z_o Ω	C		*Width "*
4	*Input frequency in MHz*	f MHz	D		
5	*Input W, width of line in inches*	W in	E		ε_{EFF}
					λ/4 (inch)
					V_f

fig. 2A. User instructions for microstrip impedance line program for HP-67/97 or HP-41 calculators.

STEP	KEY ENTRY	KEY CODE	COMMENTS
001	f LBL A	31 25 11	
	STO 2	33 02	
	1	01	
	0	00	
	0	00	
	0	00	
	STO 3	33 03	
	•	83	
	8	08	
010	3	03	
	6	06	
	CHS	42	
	STO 5	33 05	
	•	83	
	0	00	
	7	07	
	2	02	
	4	04	
	CHS	42	
020	STO 6	33 06	
	1	01	
	•	83	
	7	07	
	3	03	
	5	05	
	STO 7	33 07	
	h RTN	35 22	
	f LBL B	31 25 12	
	STO 4	33 04	
030	h RTN	35 22	
	f LBL C	31 25 13	
	STO 0	33 00	
	RCL 2	34 02	
	STO 1	33 01	
	f LBL 0	31 25 00	
	RCL 0	34 00	
	RCL 6	34 06	
	h y^x	35 63	
	RCL 7	34 07	
040	X	71	
	RCL 1	34 01	
	RCL 2	34 02	
	÷	81	
	RCL 5	34 05	
	h y^x	35 63	
	X	71	
	1	01	
	+	61	
	RCL 0	34 00	
050	X	71	
	RCL 4	34 04	
	$\sqrt{x}$	31 54	
	X	71	
	1/X	35 62	
	RCL 2	34 02	
	X	71	

STEP	KEY ENTRY	KEY CODE	COMMENTS
	3	03	
	7	07	
	7	07	
060	X	71	
	RCL 3	34 03	
	X	71	
	f INT	31 83	
	RCL 3	34 03	
	÷	81	
	RCL 1	34 01	
	g x=y	32 51	
	R/S	84	
	h x<>y	35 52	
070	STO 1	33 01	
	GTO 0	22 00	
	R/S	84	
	f LBL D	31 25 14	
	STO 8	33 08	
	h RTN	35 22	
	f LBL E	31 25 15	
	RCL 2	34 02	
	÷	81	
	•	83	
080	6	06	
	h x<>y	35 52	
	g x > y	32 81	
	GTO 1	22 01	
	•	83	
	0	00	
	2	02	
	9	09	
	7	07	
	f COS	35 63	
090	X	71	
	GTO 2	22 02	
	f LBL 1	31 25 01	
	•	83	
	1	01	
	2	02	
	5	05	
	5	05	
	h y^x	35 63	
	•	83	
100	6	06	
	3	03	
	X	71	
	f LBL 2	31 25 02	
	RCL 4	34 04	
	1	01	
	−	51	
	X	71	
	1	01	
	+	61	
110	STO 9	33 09	
	f $\sqrt{x}$	31 54	
	h 1/x	35 62	

STEP	KEY ENTRY	KEY CODE
	1	01
	1	01
	8	08
	1	01
	1	01
	RCL 8	34 08
	÷	81
120	RCL 9	34 09
	f $\sqrt{x}$	31 54
	÷	81
	4	04
	÷	81
	RCL 9	34 09
	f -x-	31 84
	h RV	35 53
	f -x-	31 84
	h RV	35 53
130	R/S	84
140		
150		
160		

REGISTERS

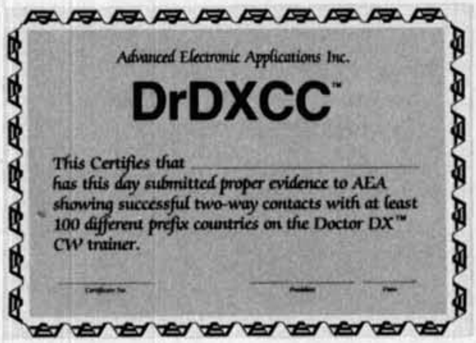
0	1	2	3	4	5	6	7	8	9
Z_o Ω		h	1000	ϵ_r	-.836	-.0724	1.735		
S0	S1	S2	S3	S4	S5	S6	S7	S8	S9

A	B	C
a	b	c
0	1	2

fig. 2B. Program listing for microstrip impedance line program for HP-67/97 or HP-41 calculators.

Doctor DX™ Challenge

BY AEA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back to basics: transistor biasing

Add a bias network to minimize the effects of temperature change

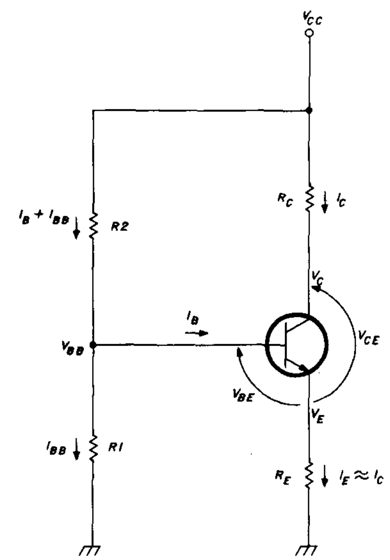
This article has been adapted, with permission, from Chris Bowick's ***RF Circuit Design,*** *published by Howard W. Sams & Company, Indianapolis, Indiana, and available from Ham Radio's Bookstore, Greenville, New Hampshire 03048, for $25.45, postpaid.*

In most RF amplifier designs, very little thought is given to the design of bias networks for the individual transistors involved. Sometimes this lack of interest is justified: if, for instance, the amplifier was expected to operate only at room temperature, there would be little need to spend much time developing an extremely temperature-stable DC operating point. If, on the other hand, the amplifier was expected to operate reliably and maintain certain specifications (gain, noise figure, etc.) over large temperature extremes — as in outdoor mounted preamplifiers, for example — the DC bias network would have to be carefully considered. A quick look at the Y and S-parameter curves for most transistors reveals that a change in the transistor's bias point does in fact change all of its RF operating characteristics. This means that the DC operating point must remain stable under specified operating conditions or the RF characteristics may change drastically.

factors affecting operating point

Two basic internal transistor characteristics — ΔV_{BE} and $\Delta\beta$ have a profound effect upon the transistor's DC operating point over temperature. The object of a good temperature-stable bias design (see **fig. 1**) is to minimize the effects of these parameters.

As the operating temperature *increases* the base-to-emitter voltage, V_{BE}, of a transistor *decreases* at the rate of about 2.5 mV/°C from its nominal room temperature value of 0.7 volts for a silicon device. As



step

1. Choose the operating point for the transistor:
   $I_C = 10\ mA \quad V_C = 10\ V \quad V_{CC} = 20\ V \quad \beta = 50$
2. Assume a value for V_E considering bias stability (see text):
   $V_E = 2.5$ volts
3. Assume $I_E \approx I_C$ for high beta transistors
4. Calculate R_E, knowing I_E and V_E:
   $$R_E = \frac{V_E}{I_E} = \frac{2.5}{10 \times 10^{-3}} = 250 \text{ ohms}$$
5. Calculate R_C, knowing V_{CC}, V_C, and I_C:
   $$R_C = \frac{V_{CC} - V_C}{I_C} = \frac{20 - 10}{10 \times 10^{-3}} = 1000 \text{ ohms}$$
6. Calculate I_B, knowing I_C and β:
   $$I_B = \frac{I_C}{\beta} = 0.2\ mA$$
7. Calculate V_{BB}, knowing V_E and V_{BE}:
   $V_{BB} = V_E + V_{BE} = 2.5 + 0.7 = 3.2$ volts
8. Assume a value for I_{BB}, the larger the better (see text)
   $I_{BB} = 1.5\ mA$
9. Calculate R1, knowing I_{BB} and V_{BB}:
   $$R1 = \frac{V_{BB}}{I_{BB}} = \frac{3.2}{1.5 \times 10^{-3}} \text{ ohms} = 2133 \text{ ohms}$$
10. Calculate R2, knowing V_{CC}, V_{BB}, I_{BB}, and I_B:
   $$R2 = \frac{V_{CC} - V_{BB}}{I_{BB} + I_B} = \frac{20 - 3.2}{1.7 \times 10^{-3}} = 9882 \text{ ohms}$$

fig. 1. Base resistor divider plus emitter resistor determines transistor's operating point.

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V_{BE} decreases, more base current is *allowed* to flow, in turn producing more collector current, which is exactly what we would like to prevent. The total change in V_{BE} for a given temperature change is called ΔV_{BE}. The primary external circuit factor that tends to minimize the effects of ΔV_{BE} and over which the circuit designer has control is the emitter voltage V_E of the transistor. A decrease in V_{BE} with temperature causes an increase in emitter current and a subsequent increase in V_E. The increase in V_E is a form of negative feedback which tends to reverse bias the base-emitter junction and, therefore, *decrease* the collector current. A decrease in V_{BE}, therefore, tends to be counteracted by the increase in V_E and the collector current does not increase as much with temperature. If these observations are put into equation form we have:

$$\Delta I_C \approx -\frac{\Delta V_{BE} I_C}{V_E} \qquad (1)$$

where ΔI_C = change in collector current
I_C = quiescent collector current
ΔV_{BE} = change in base-to-emitter voltage
V_E = quiescent emitter voltage

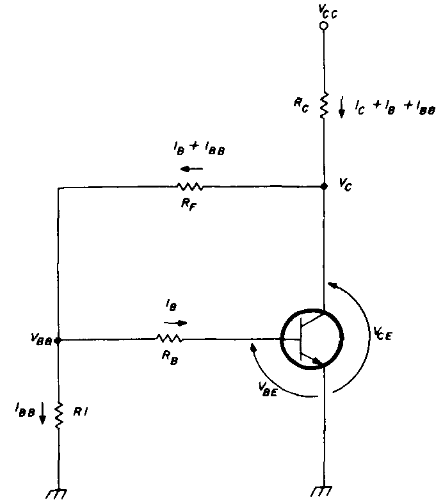
Thus, if V_E were made equal to 20 times ΔV_{BE}, then the collector current would change only 5 percent over temperature due to ΔV_{BE}. It is the value of the emitter voltage (V_E) and not the value of the emitter resistor (R_E) which is the important bias design criteria.

From **eq. 1** it appears that the higher V_E is, the better. This would be true if we only had to bias the transistor at a specific operating point. Obviously, there are other things which must be considered in the design. A high emitter voltage, for instance, does tend to waste power and decrease the AC signal gain. A bypass capacitor across R_E at the signal frequency is usually used to prevent the loss in gain, but the wasted DC power may still be a problem.

If the amplifier is to operate over a temperature change not exceeding ±50°C, then an emitter voltage of 2.5 volts will provide a ±5 percent variation in I_C due to ΔV_{BE}. In fact, you will find that the majority of the transistor bias networks similar to **fig. 1** will provide a value of V_E from 2 to 4 volts depending upon the values of V_{CC} and V_C chosen. Higher values are, of course, possible depending upon the degree of stability you need.

varying current gain affects operating point

The change in a transistor's DC current gain, β, over temperature is also important to the circuit designer. Any variation in β produces a corresponding change in quiescent collector current and changes the transistor's designed operating point. The β of a silicon transistor typically *increases* with temperature at the rate of about 0.5 percent/°C. A ±50°C temperature variation causes β and consequently collector current to vary as much as ±25 percent.



step

1. Choose the operating point for the transistor:
   $I_C = 10$ mA $V_C = 10$ V $V_{CC} = 20$ V $\beta = 50$
2. Assume values of V_{BB} and I_{BB} to supply constant current I_B:
   $V_{BB} = 2$ volts $I_{BB} = 1$ mA
3. Calculate I_B, knowing I_C and β:
   $I_B = \frac{I_C}{\beta} = 0.2$ mA
4. Calculate R_B, knowing V_{BB}, $V_{BE} = 0.7$ V, and I_B:
   $R_B = \frac{V_{BB} - V_{BE}}{I_B} = \frac{2 - 0.7}{0.2 \times 10^{-3}} = 6500$ ohms
5. Calculate R1, knowing V_{BB} and I_{BB}:
   $R1 = \frac{V_{BB}}{I_{BB}} = \frac{2}{1 \times 10^{-3}} = 2000$ ohms
6. Calculate R_F, knowing V_{BB}, I_{BB}, I_B, and V_C:
   $R_F = \frac{V_C - V_{BB}}{I_{BB} + I_B} = \frac{10 - 2}{1.2 \times 10^{-3}} = 6667$ ohms
7. Calculate R_C, knowing V_{CC}, V_C, I_C, I_B, and I_{BB}:
   $R_C = \frac{V_{CC} - V_C}{I_C + I_B + I_{BB}} = \frac{20 - 10}{11.2 \times 10^{-3}} = 893$ ohms

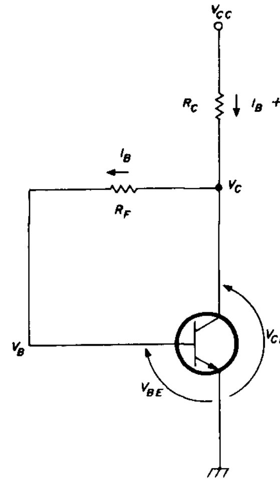
fig. 2. Collector feedback resistor minimizes temperature effects.

Not only does β vary with temperature, but the manufacturing tolerance for β among transistors of the *same* part number is typically very poor. It is not uncommon, for instance, for a manufacturer to specify a 10 to 1 range for β on the data sheet (such as 50 to 500). This, of course, makes it extremely difficult to design to bias network and achieve consistent performance.

Collector current for a corresponding change in β is approximated by:

$$\Delta I_C = I_{C1}\left(\frac{\Delta\beta}{\beta_1\beta_2}\right)\left(1 + \frac{R_B}{R_E}\right) \qquad (2)$$

where I_{C1} = collector current at $\beta = \beta_1$
β_1 = lowest value of β
β_2 = highest value of β
Δ_β = $\beta_2 - \beta_1$
R_B = parallel combination of $R1$ and $R2$ in **fig. 1**
R_E = emitter resistor



step

1. Choose the operating point for the transistor (V_C, I_C): $I_C = 10$ mA $V_C = 10$ V $V_{CC} = 20$ V $\beta = 50$
2. Calculate I_B, knowing I_C and β: $$I_B = \frac{I_C}{\beta} = 0.2 \text{ mA}$$
3. Calculate R_F, knowing V_C, V_B, $= V_{BE} = 0.7$ V and I_B: $$R_F = \frac{V_C - V_B}{I_B} = \frac{10 - 0.7}{200 \times 10^{-6}} = 46.5 \text{ K}$$
4. Calculate R_C, knowing I_B, I_C, V_{CC}, and V_C: $$R_C = \frac{V_{CC} - V_C}{I_B + I_C} = \frac{20 - 10}{10.2 \times 10^{-3}} = 980 \text{ ohms}$$

fig. 3. Simplest bias network employs collector feedback resistor.

This equation indicates that once a transistor is specified, the only control that the designer has over the effect of β changes on collector current is through the resistor ratio R_B/R_E. *The smaller this ratio, the less the collector current varies.* Again, however, some compromise is necessary. As you decrease the ratio R_B/R_E, you also produce the undesirable effect of decreasing the current gain of the amplifier. Also, as the ratio approaches unity, the improvement in operating point stability rapidly decreases. As a practical rule of thumb for stable designs, the ratio R_B/R_E should be less than 10.

Figs. 1, 2, and **3** indicate three possible bias configurations for bipolar transistors in order of *decreasing* bias stability. Complete step-by-step design instructions and an example are also included with each circuit configuration. Note that the bias networks of **fig. 2** and **3** do not contain the emitter resistor (R_E), which provides the negative feedback to counteract collector current variations over temperature. Instead, resistor R_F is connected from the collector to the base of the transistor to provide the negative feedback. Obviously, for these two designs, the designer neither has control over the ratio R_B/R_E, nor the voltage V_E of **fig. 1**. Surprisingly, however, R_F works quite well in minimizing the effects of transistor parameter variations over temperature.

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More Details? CHECK—OFF Page 160

VHF/UHF WORLD

Joe Reisert W1JR

low-noise GaAs FET technology

Over the last 20 years state-of-the-art low-noise receiver design has advanced at an incredible rate. Noise figures are now approaching 0 dB on all Amateur bands from 2 meters through 23 cm (1296 MHz), and aren't that much higher on 13 cm (2300 MHz). Most of this progress is due to advances in GaAs FET (Gallium Arsenide Field Effect Transistor) technology.

This would appear to be an appropriate time for a short historical review of the rapid changes in the state-of-the art of low-noise amplifiers. In this column, I'll also provide an overview of state-of-the-art design and present examples of some typical circuits being used to obtain extremely low noise figures. I'll also try to look a bit into the future. Because of space limitations, I'll concentrate on 2 meters through 70 cm (432 MHz), presenting straightforward designs. If you tell me you're interested, I'll cover high-performance circuit techniques applicable to use on the higher frequencies in a future column.

history

Only 25 years ago vacuum tubes such as Western Electric 417A and the gold plated 416B triodes with associated high voltage supplies and blower cooling were the state-of-the-art on the VHF bands.[1,2] Nuvistors like the 6CW4, although less expensive and easier to use, came later, but they had higher noise figures.

Early in the 1960s, paramps (parametric amplifiers) made the first two-way EME contacts possible on 23 cm and later on 70 cm.[3,4] However, these amplifiers were quite complex and generally required ferrite circulators, special diodes in a complex matching structure, and, of course, a very high frequency (typically 5 to 10 times the operational frequency of the preamp) pump oscillator. At first a klystron pump was used, but it required a regulated high voltage and was very temperature sensitive. Later (about 1970), Gunn diode oscillators were used, but they too were not for the neophyte. When the ultimate in low-noise figure was required, paramps were mounted right at the antenna, but drift and temperature changes, not to mention inconveniences in tuning, made this a last resort!

table 1. Typical low-noise GaAs FET DC and RF parameters:

DC parameters	
BV_{ds}	5 volts
BV_{gs}	−5 volts
V_p	−1 to −6 volts
I_{dss}	20-125 mA
I_d	10-50 mA
power dissipation	250-500 milliwatts
g_m	20-100 millimhos
RF parameters	
F_{max}	50-100 GHz
MAG*	15-20 dB
minimum noise figure*	0.5-2.0 dB

*At specified test frequency.

Low-noise JFETs as well as low-noise bipolar transistors became available in the late 1960s.[5,6] The really reliable low-noise UHF transistor amplifiers, however, weren't readily available until the 1970s when the FMT4575 and NE64535 arrived.[7,8] Many of these devices required special filtering or alignment techniques and were quite expensive ($30 to $55 each).

Semiconductor experts soon realized that the state-of-the-art was rapidly closing in on the maximum usable frequency where silicon bipolar transistors could operate even when using specialized techniques such as arsenic emitters. Thus when GaAs MESFETs (Metal Semiconductor FET — the full formal name for a GaAs FET) using a narrow Schottky barrier gate were first developed,[9,10] they were received with great interest.

The first commercial GaAs FET was the Fairchild Microwave FMT-900 introduced in 1972, at $500 each. It was good news to Amateurs when affordable low-noise GaAs FETs became available in the 1970s;[11] they were soon in widespread use on 70 cm and later 2-meter EME. Now they're used almost exclusively on EME from 2 meters through 13 cm. Many VHF/UHFers are presently using inexpensive (under $6) single and dual-gate GaAs FETs, even for routine weak signal and tropo propagation work.

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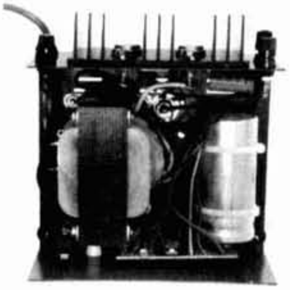

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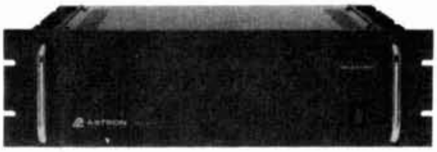

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RS-7B	5	7	4 × 7½ × 10¾	10
RS-10A	7.5	10	4 x 7½ x 10¾	11
RS-12A	9	12	4½ x 8 x 9	13
RS-20A	16	20	5 x 9 x 10½	18
RS-35A	25	35	5 x 11 x 11	27
RS-50A	37	50	6 x 13¾ x 11	46

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RS-50M	37	50	6 x 13¾ x 11	46

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RS-10L(For LTR)	7.5	10	4 × 9 × 13	13
RS-12S	9	12	4½ x 8 x 9	13
RS-20S	16	20	5 x 9 x 10½	18

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GaAs FET parameters

What are these wonderous GaAs FETs anyhow? In many ways the single-gate GaAs FET operates like a triode vacuum tube, but with extremely low voltage (typically less than 5 volts) and no filaments to heat! There are some similarities to ordinary FETs. The input is usually the gate, which has a high input impedance (1-10 kilohms typical at VHF/UHF). The common element is usually the source and the drain is the output, which has a typical impedance in the region of 200 to 1000 ohms. The width of the gate, typically 0.5 to 2 microns, has the greatest effect on gain and maximum frequency of operation, with the narrower widths required for the higher frequencies (more on this later — see **fig. 1**).

Table 1 shows that each GaAs FET device has certain important low frequency or DC parameters such as: BV_{ds}, the maximum drain to source breakdown voltage; BV_{gs}, maximum gate to source breakdown voltage; V_p, pinchoff voltage, the minimum gate voltage required to decrease drain current to a specified low value; I_{dss}, maximum drain current with zero volts gate bias; I_d, typical operating drain current; maximum power dissipation, and g_m or transconductance (in millimhos).

Table 1 also shows the most important GaAs FET RF characteristics such as F_{max}, MAG, NF_{min}. F_{max} is the maximum frequency of oscillation. The MAG (maximum available gain) is usually specified by the manufacturer at only one or two frequencies: 10 to 25 percent of F_{max} at a specified drain current, typically at 25 to 50 percent of I_{dss}. Noise figure is usually specified at the same frequency or frequencies at a drain current of 10 to 20 percent of I_{dss}.

One of the important properties of the GaAs FET is that noise figure is primarily limited by thermal noise generated in the channel or gate of the device. Plots of maximum gain and lowest possible noise figure over the usable frequency spectrum are often provided on data sheets. As with bipolar and JFET devices, the gain increases with decreasing frequency while noise figure may decrease slightly below the test frequency. Typical input and output impedances over a wide frequency range are often shown on the GaAs FET data sheets as well as the optimum source impedance for minimum noise figure.

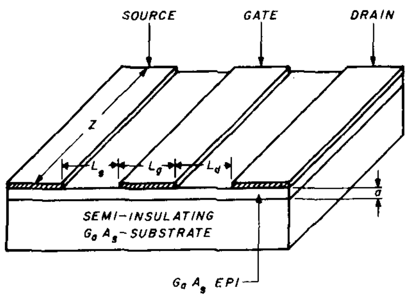

fig. 1. Cross-section of a GaAs FET.

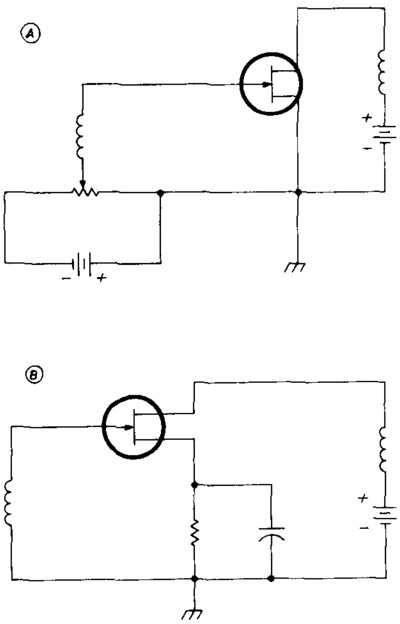
fig. 2. Typical GaAs FET biasing schemes: (*A*) battery bias; (*B*) source bias.

Recently, dual-gate GaAs FETs have been manufactured primarily for the low-cost home entertainment market. These devices are often used as amplifiers in the cascode configuration with optional AGC applied to the second gate. They've also been used as mixers, with the local oscillator applied to gate 2 in a manner similar to the dual-gate MOSFETs in the HF and VHF spectrum. SHF dual-gate GaAs FETs are also becoming available, but are still quite expensive — typically over $50.

So far I've described the so-called low-noise or small-signal GaAs FETs, primarily used in preamplifiers. High-power GaAs FETs are available with up to 8 watts of linear power; some of these also perform well in low-noise preamplifiers (more on this later). These devices also have potential for very high dynamic range. GaAs FETs are already appearing in commercial transmitter chains as well as in receivers!

A partial list of the most popular GaAs FETs typically used by Amateurs is shown in **table 2**. Typical noise figures and cost are also listed for comparison. Furthermore, new and improved devices are constantly becoming available as the technology matures. Low noise GaAs FETs usable to well beyond the 3 cm (10 GHz) band are now available, but are still quite expensive — over $50.

biasing GaAs FETs

GaAs FET biasing is quite different from bipolar transistors. In fact, the GaAs FET is more like the old triode vacuum tube. The commmon source configuration is presently the most popular topography. For proper operating parameters, the gate must be reverse biased with respect to the source.

When GaAs FETs were first introduced, battery bias was usually applied to the gate as shown in **fig. 2A**.[11] Many users, however, experienced problems such as burnouts due to battery failure or poor contacts on battery holders. Later, commercial amplifier manufacturers developed exotic bias supplies.[12] Source biasing with a single resistor (**fig. 2B**) became popular once it was determined that the source could be adequately RF bypassed without lowering gain or circuit performance.[13]

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input impedance matching

GaAs FETs have much higher input impedances than bipolar transistors. As a result, different input matching techniques are required. The most popular are the π-network, the broadband inductor match, and the tank circuit (**fig. 3**).

In the 1960s, π-network input matching became popular, with low-noise bipolar transistors used on 23 cm. Though Amateur GaAs FET preamplifiers on 23 and 13 cm often used this technique, shown in **fig. 3A**,[14] it never became very popular in Amateur preamplifiers for 2 meters through 70 cm.

Several years ago the NRAO (National Radio Astronomy Observatory) developed cooled GaAs FET preamplifiers for radio astronomy.[15] These used a single inductor to match to the gate and reactive lossless feedback in the source (**fig. 3B**). This technique had previously been used by Reisert in a 70-cm bipolar amplifier;[16] it has since been used by Angle and Sutherland in Amateur GaAs FET preamplifiers.[17,18] However, it's more popular on 33 cm (902 MHz) and above, since high-pass filtering (which can create additional losses) is required ahead of the preamplifier.[19]

The most popular input matching technique used by Amateurs on 2 meters through 70 cm is probably the tank circuit. The tapped inductor version is the least expensive and potentially has the lowest loss, since only one tuning capacitor is required (**fig. 3C**). The capacitance coupled tank circuit (**fig. 3D**), however, is even more commonly used because it doesn't require the tedious selection of the proper tap point.

The tank circuit in **fig. 3D** can match a GaAs FET over a very wide impedance range from optimum gain to optimum noise figure. In addition, it has built-in selectivity, eliminating the need for additional external filtering. However, the unloaded Q (more on this later) of all the components in the network must be very high if the inherent low-noise figure of the device is to be achieved.

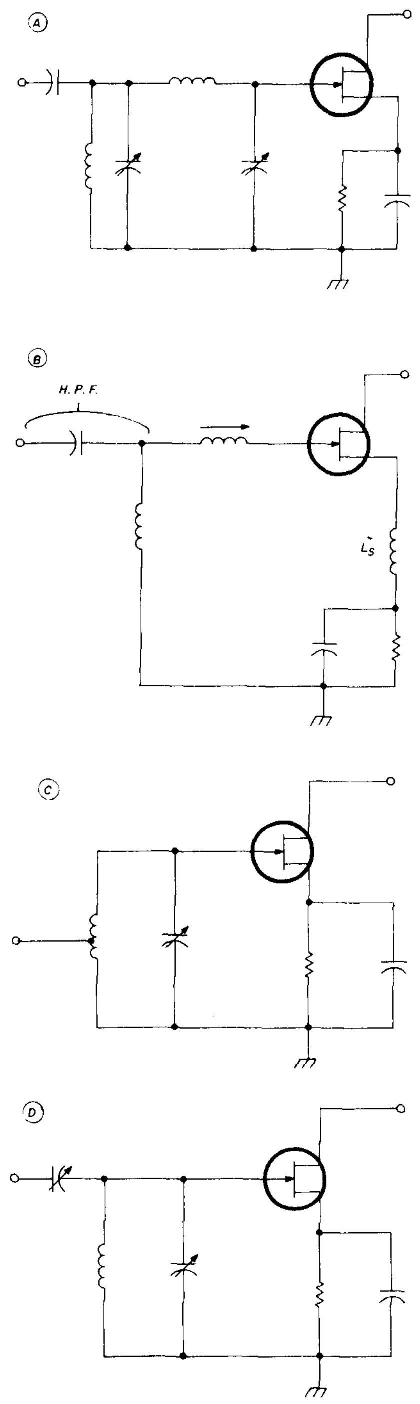


fig. 3. Typical input matching schemes for GaAs FET preamplifiers: *(A)* π-network; *(B)* broadband inductor; *(C)* inductively coupled tank circuit; *(D)* capacitance coupled tank circuit.

output impedance matching

Finally, the GaAs FET must be matched to the next stage. Here again, special techniques are required. The output impedance of most small-signal GaAs FETs is moderate (100-1000 ohms typically). Furthermore, high gain increases the chances of instability or oscillations. Therefore, the earlier Amateur GaAs FET circuits used simple loading techniques such as those shown in **fig. 4A**.[7,11] Sando used tuned tank circuits (**fig. 4B**)[20,21] and Sutherland advocated the use of a broadband 4:1 bifilar wound transformer (**fig. 4C**).[22]

All of these techniques have advantages and disadvantages. While the resistor loading approach is simple, improves stability, and is inexpensive, it lowers gain and output power by up to 3 dB. The tuned circuit delivers more gain but is more expensive and requires more input to output circuit isolation to prevent instability. The 4:1 transformer is easy to construct but has slightly less gain as well as poor to moderate output VSWR.

recommended circuit

Figure 5 shows a recommended GaAs FET circuit for operation on 2 meters through 70 cm. The input matching is the capacitance coupled tank circuit (**fig. 3D**) since it permits either conjugate matching for low input VSWR and maximum gain or optimum source impedance for minimum noise figure. The output circuit uses a 4:1 transformer as discussed earlier. One transformer covers the entire operating range from 2 meters to 70 cm. Notice that a ferrite bead is also placed on the drain lead of the GaAs FET. This reduces gain slightly 0.5-0.75 dB maximum, but improves stability, especially in the microwave region where oscillations often occur when using GaAs FETs.

Source biasing is used because it is inexpensive and self-protecting. Most single-gate GaAs FETs have two source leads and therefore lend themselves to using two source bypass capacitors and resistors. This also simplifies bias resistor selection and increases device protection in case one of the source leads should become loose. Ferrite beads are also slipped on the source resistors since they've been reported to prevent certain instabilities in particularly stubborn circuits.

Drain voltage is supplied by a three-terminal voltage regulator that acts like a zener diode but operates over a much wider range of device current. It is followed by a resistor and zener diode for over-voltage/current protection. Placing the resistor after the regulator protects the zener and decreases the drain voltage if too much drain current is present. For the ultimate in low-noise figure, this resistor value can be raised or lowered slightly for optimum drain voltage, which is typically not critical.

dual-gate GaAs FET circuit

A typical dual-gate GaAs FET circuit is shown in **fig. 6**. This is virtually the same circuit as the one just discussed but with an additional voltage source for gate 2. Most VHF/UHF dual-gate devices are very low in cost because they're primarily aimed at the TV and consumer market and have often equalled the performance of their single gate cousins, especially on 2 meters through 70 cm.[23] They typically run at higher I_d.

component selection

One of the first decisions you'll have to make is which GaAs FET to use. Some GaAs FETs (with typical specifications and prices) popular with Amateurs, are shown in **table 2**. The beauty of the circuit diagrams shown in **figs. 5** and **6** is that they'll work well through 500 MHz with just about any GaAs FET as long as the input inductor, L1, (see the parts list on **figs. 5** and **6**) is changed for the proper frequency range.

When striving for a very low-noise figure below 500 MHz, don't use a high-priced GaAs FET. The narrow gate (less than 1 micron) higher frequency units will rarely yield any better performance at lower frequencies than the less expensive units and will frequently be more susceptible to instabilities.

It has been speculated that 1/f noise, the property of a solid-state device to have increased noise figure when operated below a certain low frequency cutoff, can actually increase noise figure at the lower VHF frequencies if a device with too high an F_{max} is used. Furthermore, even the low-cost, lower frequency GaAs FETs have noise figures that equal the higher frequency higher cost units at *lower* frequencies. These points are illustrated in **fig. 7**.

Input circuit losses must be kept to an absolute minimum as discussed in reference 19. The input circuit shown in **figs. 5** and **6** was chosen because it has such a wide range of tuning, good out-of-band rejection, and a reasonably low loaded "*Q*." But it also uses two capacitors, which will increase the dissipation loss in the tank circuit. High-*Q*, low-inductance air variable capacitors such as the Johanson 5200 series or equivalent are a

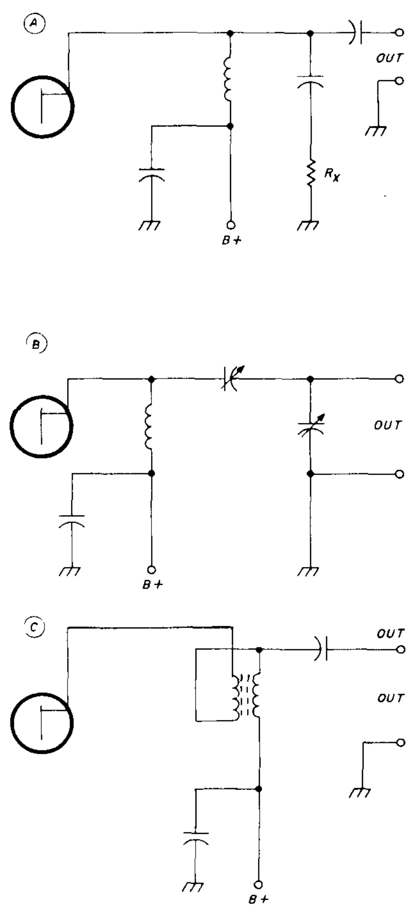
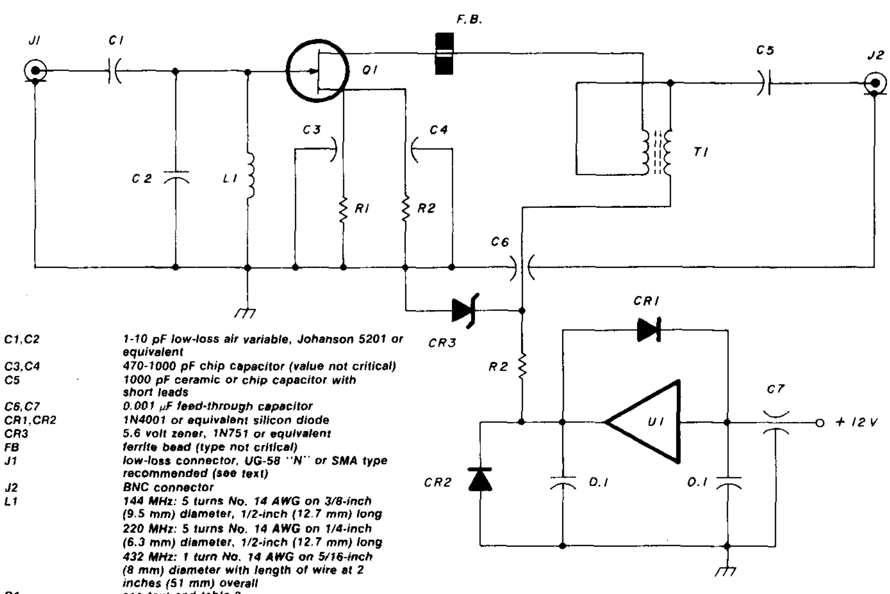
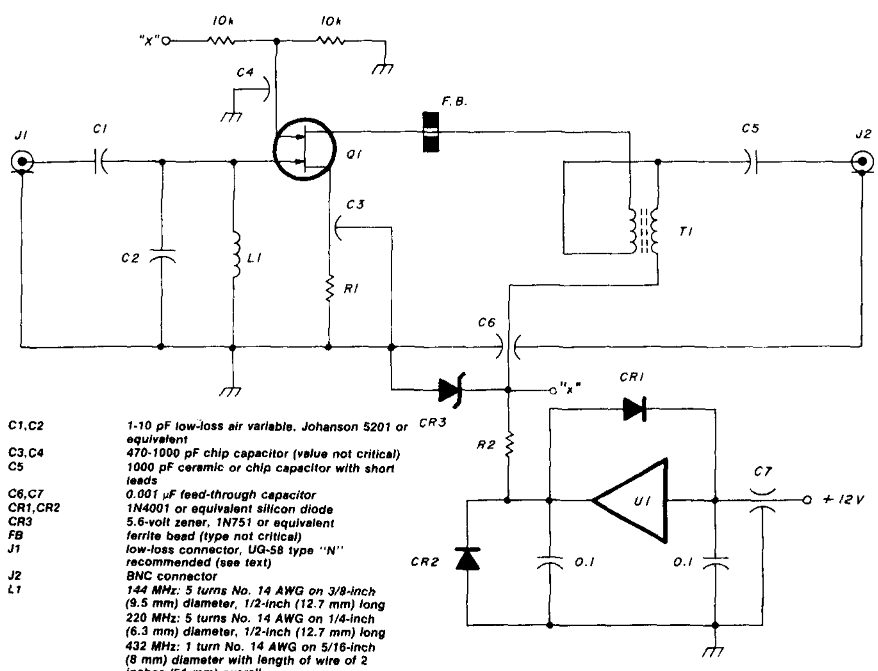


fig. 4. Typical output matching schemes for GaAs FET preamplifiers. (*A*) resistor loading; (*B*) tuned circuit; (*C*) broadband transformer.



C1,C2	1-10 pF low-loss air variable, Johanson 5201 or equivalent
C3,C4	470-1000 pF chip capacitor (value not critical)
C5	1000 pF ceramic or chip capacitor with short leads
C6,C7	0.001 μF feed-through capacitor
CR1,CR2	1N4001 or equivalent silicon diode
CR3	5.6 volt zener, 1N751 or equivalent
FB	ferrite bead (type not critical)
J1	low-loss connector, UG-58 "N" or SMA type recommended (see text)
J2	BNC connector
L1	144 MHz: 5 turns No. 14 AWG on 3/8-inch (9.5 mm) diameter, 1/2-inch (12.7 mm) long 220 MHz: 5 turns No. 14 AWG on 1/4-inch (6.3 mm) diameter, 1/2-inch (12.7 mm) long 432 MHz: 1 turn No. 14 AWG on 5/16-inch (8 mm) diameter with length of wire at 2 inches (51 mm) overall
Q1	see text and table 2
R1,R2	select for proper I_d, 300 ohms typical (see text)
R3	100 ohms typical (see text)
T1	3 turns No. 30 or No. 32 AWG bifilar wire wound on a ferrite bead
U1	5 volt, 3-terminal regulator, 78L05 or equivalent

fig. 5. Recommended GaAs FET preamplifier circuit for 144, 220, and 432 MHz operation.



C1,C2	1-10 pF low-loss air variable. Johanson 5201 or equivalent
C3,C4	470-1000 pF chip capacitor (value not critical)
C5	1000 pF ceramic or chip capacitor with short leads
C6,C7	0.001 μF feed-through capacitor
CR1,CR2	1N4001 or equivalent silicon diode
CR3	5.6-volt zener, 1N751 or equivalent
FB	ferrite bead (type not critical)
J1	low-loss connector, UG-58 type "N" recommended (see text)
J2	BNC connector
L1	144 MHz: 5 turns No. 14 AWG on 3/8-inch (9.5 mm) diameter, 1/2-inch (12.7 mm) long 220 MHz: 5 turns No. 14 AWG on 1/4-inch (6.3 mm) diameter, 1/2-inch (12.7 mm) long 432 MHz: 1 turn No. 14 AWG on 5/16-inch (8 mm) diameter with length of wire of 2 inches (51 mm) overall
Q1	see text and table 2
R1	select for proper I_d, 100 ohms typical (see text)
R2	100 ohms typical (see text)
T1	3 turns No. 30 or No. 32 AWG bifilar wound on a ferrite bead
U1	5-volt 3-terminal regulator, 78L05 or equivalent

fig. 6. Typical dual-gate GaAs FET circuit.

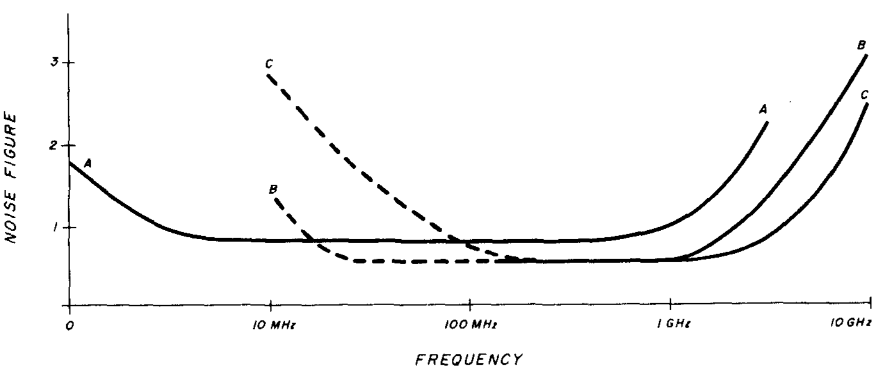


fig. 7. Typical noise figures versus frequency of several types of devices: (*A*) typical bipolar transistor; (*B*) typical GaAs FET; (*C*) narrow gate high frequency GaAs FET.

necessity if low-noise figure (operation) is required.

From laboratory measurements I've made, the best unloaded *Q* for properly designed discrete inductors is about 400-500. **Table 3** shows that the typical loaded *Q* for 2-meter through 70-cm GaAs FET preamplifiers tuned for minimum noise figure is between 8.6 and 16. Hence the calculated tank circuit losses can easily be from 0.15 to 0.35 dB with properly adjusted preamplifiers.[19] Consequently the tank circuit losses ahead of the typical GaAs FETs presently used by Amateurs are undoubtedly the main contributor to the overall noise figure and deserve lots of attention by those striving for the ultimate low-noise preamplifier for EME. The type of input connector chosen is very important, especially on 70 cm and higher. A type "N" is highly recommended. SMA connectors, although more expensive, are also an excellent choice, especially if size is a constraint.

In laboratory measurements I've made, BNC connectors often caused the noise figure to increase or become erratic, especially when you press against the side of the connector. Hence they're not recommended for use on the input to a low-noise preamplifer, since losses and VSWR translate directly to increased noise figure. However, these connectors are acceptable on the output of a preamplifier because the gain ahead of the connector will mask any possible loss.

construction techniques

Needless to say, extremely low-noise circuits such as those just described require particular attention to lead length, layout, and quality of components if the specified performance is to be obtained. Therefore I've shown a recommended construction technique in **fig. 8**. This configuration is a combination of techniques I've seen used including those of the Michigan Microwave Group and Krauss in his 13 cm preamps.[24]

A medium-size cast box such as the Bud model CU-124 or equivalent is highly recommended. Smaller boxes such as the Bud CU-123 and the Pomona Electronics 2417 have sometimes caused oscillations. Smaller boxes may require placing the input inductors too close to the DC biasing component, causing excessive coupling between them and consequently lowering the unloaded *Q*. Furthermore, unloaded inductor *Q* is almost always higher when coils are physically large and the volume surrounding them is even larger. This is the technique used to design high-*Q* helical inductors.[19]

Finally, grounding is important. I use a double-clad PC board attached to the cover of the cast box by the input and output connectors. But because typi-

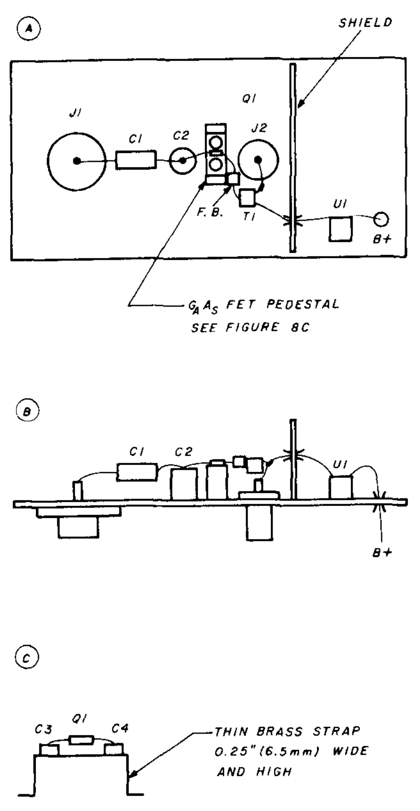


fig. 8. Recommended circuit layout for the preamplifiers shown in *figs. 5* and *6*.

cal air-variable capacitors will not go completely through the box cover and PC board, a hole large enough to pass the nut on the shunt capacitor, C2, must be drilled through the cast box. For proper grounding and lowest loss, drill two holes adjacent to this capacitor and tighten the PC board to the cover of the cast box with two short screws. A small hole can be drilled in the box cover between J1 and C2 to facilitate adjusting C1.

handling GaAs FETs

GaAs FETs with their low breakdown voltages and high impedances are far more likely to be destroyed by improper handling than their bipolar cousins. When installing a GaAs FET in a circuit, be sure your soldering iron has a grounded tip. If you're not absolutely certain the tip is grounded, place a wire with a pair of large alligator clips between the soldering iron tip and chassis ground. Touch the chassis with one hand and then pick up the GaAs FET container with your other hand. Then remove the device from its packing material while still holding on to the chassis. This will prevent static discharge when the device comes in contact with the circuit. *Caution: Don't wear woolen clothing or walk across a carpet just before working on the installation of a GaAs FET. Static electricity present can cause burnout!**

protecting GaAs FETs from RF and DC

If you use the biasing scheme recommended in **figs. 5** and **6**, there should be little chance of burnout due to DC parameters because the circuit is self limiting. Just make sure to observe the power dissipation rating when testing for I_{dss}. For ultimate DC reliability, the power supply should be of the regulated type despite the use of a three-terminal voltage regulator within the preamplifier. Diodes CR1 and CR2 in **figs. 5** and **6** provide additional protection from voltage spikes or reverse polarity. *Voltage should never be obtained from a power supply that is also used to power relays because voltage spikes may be induced by the relay coil inductance and hence cause solid-state devices to burn out. In addition, turning the power supply on and off between receive and transmit can actually increase the chances of burnout, since most solid-state devices are less likely to burn out when operating at their proper bias levels.*

The usual RF handling precautions apply. For maximum protection, the RF input level should never exceed 100 and preferably be no greater than 10 milliwatts. A dual relay protection scheme is highly recommended.[25] While GaAs FETs are not nearly as RF fragile as many persons would have you believe, they may, *when overstressed* with RF, tend to experience catastrophic failures, rendering them completely inoperative. Bipolars, on the other hand, may degrade slowly without you necessarily being aware of their progressive degradation.[26]

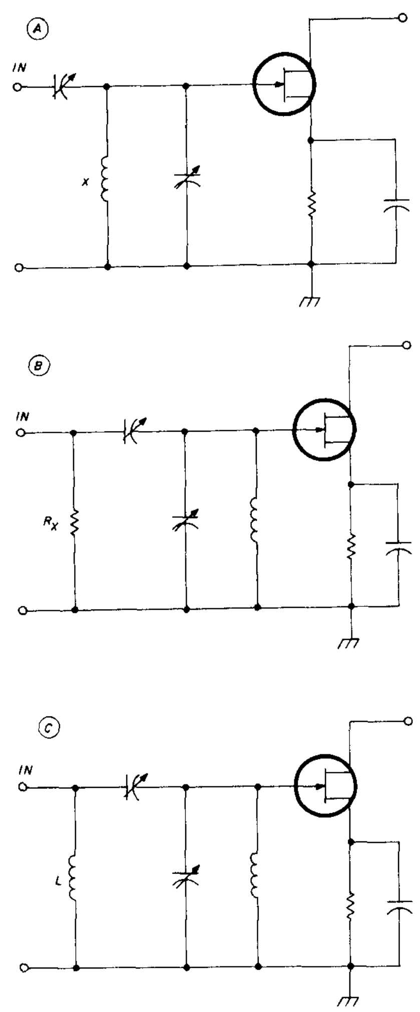


fig. 9. Some recommended protection techniques for decreasing the chances of lightning or static damage to GaAs FET preamplifiers: (*A*) capacitance coupled input tank; (*B*) resistor bleeder; (*C*) shunt inductor.

Most of the energy in static and lightning is concentrated in or below the HF spectrum. Therefore, use of the capacitance coupled tank just recommended will lessen chances of burnout since the input capacitor will act like a high-pass filter. Using a bleeder resistor or inductor across the input connector may also help, but can degrade noise figure slightly if not properly chosen (**fig. 9**). Back-to-back diodes across the input connector are not recommended because any presence

*See K4KEF's "Static Electricity and Modern Integrated Circuits," *ham radio*, March, 1984, page 33. — **Editor**

table 2. GaAs FETs presently in Amateur use.

type	NF (typically) and frequency	approximate price	notes
ALF 1023	1.3 at 4 GHz	$15.00	discontinued
D432	1.8 at 4 GHz	26.00	Dexcel
MGF 1100	2.5 at 4 GHz	7.35	dual-gate
MGF 1202	2.0 at 4 GHz	9.70	low cost MGF 1402
MGF 1402 (2SK274)	1.1 at 4 GHz	14.00	very popular
MGF 1404	0.65 at 4 GHz	66.60	expensive
MGF 1412 (2SK275)	1.0 at 4 GHz	21.00	popular above 70 cm
MRF 966	1.2 at 1 GHz	3.75	new dual-gate
NE41137 (3SK124)	1.3 at 900 MHz	5.00	dual-gate
NE72089	1.0 at 4 GHz	12.00	low-cost NE21889
3SK97	1.3 at 1 GHz	5.00	process problems
3SK121	1.5 at 800 MHz	3.50	dual-gate

of RF at lower frequencies may cause severe intermodulation products or spurious signals to appear.

testing and tuning the preamplifier

After the preamplifier is completely wired, inspect for any possible wiring errors. Then test for proper DC operation with a variable voltage power supply, first connecting the preamplifier ground to the power supply ground and then connecting the DC input and bringing the voltage up slowly to verify that the three-terminal voltage regulator is properly functioning. Now check for proper GaAs FET biasing. First connect a suitable 50 ohm termination on the preamplifier input and output connectors. Turn on the power supply and apply 12 volts to the preamplifier. In the GaAs FET, properly performing DC conditions are satisfied when the voltage drop across the 100-ohm resistor in series with the drain line is between 0.5 and 2.5 volts, indicating an I_d of 5-25 milliamperes, respectively.

The optimum value of I_d for low noise is not critical; 10-15 milliamperes is typical for low-noise GaAs FETs. But if you want to determine the optimum value, temporarily short the source lead to ground with a piece of wire and read the voltage across the 100-ohm resistor. Then calculate I_{dss}. Optimum I_d is typically 15 percent of I_{dss}. Therefore, to provide the optimum I_d for best noise figure, change the value of the source resistors until the correct I_d is obtained for each individual GaAs FET.

Next, tune the preamplifier for maximum gain, using a suitable signal source or by listening to a local station. 15-25 dB gain should be readily obtained, depending on the device and frequency range selected. Then, if you have a noise-figure generator or can obtain the use of one, optimize the input circuit for minimum noise figure. Obtaining the best noise figure will require the series input capacitance to be increased somewhat over the amount required for maximum gain. Note that the gain will drop somewhat when the circuit is optimized for best noise figure as indicated in **table 3**.

Finally, carefully measure the forward gain. If possible, also measure the reverse loss. For complete stability, the reverse loss should be at least 5 to 6 dB greater than the forward gain. For example, referring to the test data in **table 3**, the stability margin (the reverse loss minus the forward gain) at 2 meters is 9 dB for the low-noise unit but only 1.6 dB in the maximum gain unit, indicating that there are potential stability problems. A look at the 220 MHz test data will show an even worse example!

performance

Table 3 can be used as a guide for all GaAs FET preamplifier performance parameters except noise figure. As previously pointed out, the data was taken on a sample GaAs FET chosen not for lowest noise figure, but only for comparison of typical performance. Lower noise figure devices such as those shown in **table 2** should yield performance as indicated in the table.

When a GaAs FET is tuned for best noise figure, the gain will drop a few dB. The gain peak may move to a different frequency. However, this should still be more than sufficient gain even in the most stringent system. Because the suggested output circuit is a broadband device, the overall bandwidth measured for the circuit should be that of the input tank circuit. Hence, circuit losses can be easily calculated as shown in reference 19.

lower noise figures

The circuits shown should be more than adequate to yield state-of-the-art performance in the 2 meter through 70-cm spectrum. Furthermore, noise figures of less than 0.5 dB should be about all that can be used even on 70-cm EME, where the sky has a low noise temperature. Lower noise figures can sometimes be obtained by using the higher cost devices as discussed earlier. However, to lower noise figure further usually requires additional work on the components and structures used in the input tank circuit.

Q tests were performed on the microstrip circuit suggested by Sutherland.[22] However, tests typically yielded an unloaded *Q* of 400-500, no higher than the inductors suggested earlier. Although I have not tried it, adding a three-sided 1 inch (25.4 mm) square

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102	114	126	138	150	162	174	186	198	210	222	234	246	258	270	282	294	306	318	330	342
103	115	127	139	151	163	175	187	199	211	223	235	247	259	271	283	295	307	319	331	343
104	116	128	140	152	164	176	188	200	212	224	236	248	260	272	284	296	308	320	332	344
105	117	129	141	153	165	177	189	201	213	225	237	249	261	273	285	297	309	321	333	345
106	118	130	142	154	166	178	190	202	214	226	238	250	262	274	286	298	310	322	334	346
107	119	131	143	155	167	179	191	203	215	227	239	251	263	275	287	299	311	323	335	347
108	120	132	144	156	168	180	192	204	216	228	240	252	264	276	288	300	312	324	336	348
109	121	133	145	157	169	181	193	205	217	229	241	253	265	277	289	301	313	325	337	349
110	122	134	146	158	170	182	194	206	218	230	242	254	266	278	290	302	314	326	338	350
111	123	135	147	159	171	183	195	207	219	231	243	255	267	279	291	303	315	327	339	
112	124	136	148	160	172	184	196	208	220	232	244	256	268	280	292	304	316	328	340	

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table 3. Typical preamplifier performance: data taken with the same GaAs FET in the same circuit (fig. 5). Only the input inductor was changed for each frequency range. The GaAs FET used was not chosen for superior performance but for continuity of data. However, the data indicates the typical performance to be expected when optimizing for either gain or noise figure.

frequency (MHz)	144 (1)	144 (2)	220 (1)	220 (2)	432 (1)	432 (2)
forward gain in dB	25.4	21.00	25.20	20.10	19.20	17.70
reverse gain in dB	-27.0	-30.00	-24.00	-28.00	-25.50	-28.00
stability margin in dB	1.6	9.00	-1.20	7.90	6.30	10.30
input VSWR	1:1	4.8:1	1:1	4.2:1	1:1	3:1
output VSWR	12:1	2.6:1	16:1	1.8:1	2.2:1	1.7:1
output power at compression (dBm)	8.0	6.00	6.30	7.00	4.50	5.00
noise figure in dB	1.4	0.53	1.35	0.66	1.28	1.00
input impedance in kilohms	6.8	1.80	6.20	1.80	1.90	1.23
bandwidth in MHz	2.9	9.00	5.80	25.40	22.30	30.80
overall "Q" of amplifier	49.7	16.00	37.90	8.70	19.40	14.00

Note 1: Preamplifier gain matched.
Note 2: Preamplifier noise figure optimized.

copper shield to completely enclose the microstrip may decrease losses, since all the current surrounding the inductor would be within the shield and not circulating in the aluminum of the shielded box. Experiments were conducted using silver plated 1/4-wavelength cavities with an unloaded *Q* of over 1000. These are much larger and more expensive structures but they could lower input losses by at least 0.1 to 0.25 dB.

The air-variable capacitors used also contribute to loss. The ones suggested have a specified unloaded *Q* of 5000 at maximum capacitance at 100 MHz and commensurately lower *Q* at increasing frequency. Older models such as the Johanson 2950 series and its equivalent have an unloaded *Q* of 2500 respectively. Because the series matching capacitor seldom exceeds 3.0 pF in the 220 MHz and higher frequency preamplifiers, a physically smaller, higher *Q* capacitor such as the Johanson 5800 series with a 0.3-3 pF range and unloaded *Q* of 7500 at 100 MHz may be a way to decrease input losses further. The Johanson Giga-trim™ types are not recommended since they have lower *Q* than the air-variable types.

Finally, GaAs FETs can be cooled. Thermo-electric coolers, used by commercial manufacturers to cool GaAs FET type preamplifiers and thereby decrease noise figure, are now available to Amateurs. Liquid nitrogen has been used by some Amateurs, but is costly and requires special handling and mechanical arrangements.

looking ahead

GaAs FETs are becoming increasingly less expensive and easier to use. Almost monthly, a new lower noise figure and lower cost device seems to appear. This is particularly true as you go higher in frequency. Before long, receivers using only diode mixers will be a thing of the past, expecially at 3 cm (10 GHz), where devices with noise figures under 2 dB are now affordable. The increase in performance should be phenomenal, since most work on this band has been conducted with noise figures of 5 to 10 dB. The possibilities for SHF EME are considerable — if the industry also shifts to the production of inexpensive kilowatt transmitting devices!

The use of monolithic hybrid matched amplifiers will take hold. At the present state-of-the-art, the prices are acceptable ($5 to $20), but the silicon devices available often have noise figures of 3 to 6 dB and the lower noise devices seldom have adequate dynamic range. The present frequency limit is typically 3 GHz. However, monolithic GaAs FET hybrids are now commercially available. They hold great promise because while their price will be competitive with present silicon technology, they will go higher in frequency and have much lower noise figures and potentially higher dynamic range.

Although this is primarily a receiver preamplifier article, I should mention that power GaAs FETs are being developed for transmitter applications. They're presently available with up to 8 watts and are closing in on the 3 cm area. Soon the last bastions of tubes will be at stake.

This article was not intended to plow new ground, but instead to bring readers up-to-date on the state-of-the-art in low noise figure preamplifier design using GaAs FETs below 500 MHz. An attempt was made to take the fear out of using GaAs FETs even in your everyday circuits; if you follow the cautions mentioned above, you shouldn't have any problems.

While usually somewhat behind the commercial manufacturers, Amateurs are constantly striving to stay abreast of technology. Right now we're running neck-and-neck with industry on VHF/UHF noise figures. We have one great advantage in that we frequently only need narrow-band circuits that don't require economical reproduction. We can therefore take advantage of all the "tricks of the trade" and optimize to our hearts' content. The typical noise figures measured at VHF/UHF conferences on typical Amateur-built low-noise preamplifiers attest to the fact that we are very close to noise figure *perfection* on the VHF and lower UHF frequencies!

acknowledgements

I'd like to particularly thank Bob Sutherland, W6PO, and Shigeru Sando, JH1BRY, who encouraged me to try a GaAs FET preamp on my 70-cm EME station. I've been a firm believer in GaAs FETs ever since!

postscript

It seems hard to believe that a year has already passed since my first column in this series. I've tried to cover a lot of ground and still have plans for a few more articles dealing with important background material. When this material is all in place, I expect to build on this material and move forward. Your ideas for future columns are always welcome!

I sincerely want to thank in particular the *ham radio* staff — K2RR, KA1LBO, and N1ACH, who have all helped me through this year. I'd also like to thank all of you who have encouraged me with your notes and calls. Believe me, it takes many hours and a big slice out of your operating time to write this quantity of material month after month. A kind word here or there does push me along.

Unfortunately, I haven't always been able to answer all the letters received, expecially those without SASEs or those that hit me with multiple questions and new design requirements. If I were to answer all the letters as received, there'd be little time left to write the column!

Don't let this discourage you from writing. If you need an answer to a question, or if a problem occurs with something mentioned in one of my columns, please send a letter with an SASE. Keep your question brief so that I can respond quickly and easily. I'd rather answer several letters from one individual with one question per letter than multiple questions in a single letter. I hope you understand. Thank you, in advance, for your understanding and encouragement! It is truly appreciated.

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coming VHF/UHF events:

December 13, 0048 UTC: *predicted peak of Geminids meteor shower*

December 18: *EME perigee*

December 21, 1600 UTC: *predicted peak of Ursids meteor shower*

December 22, (±2 weeks): *annual secondary peak of sporadic "E" propagation*

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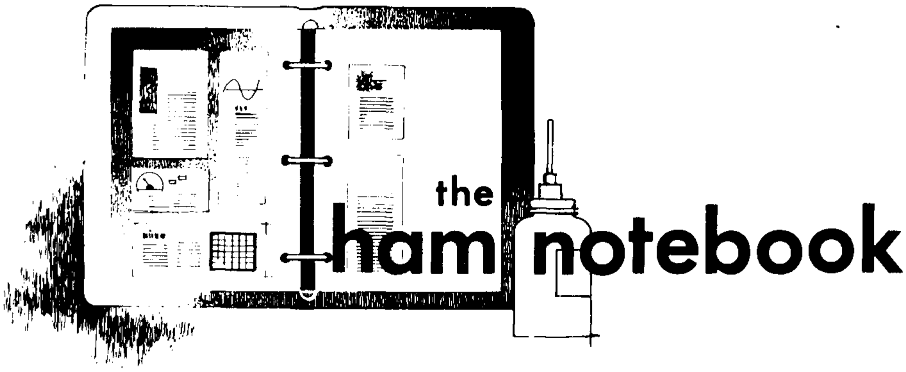

a different approach to the inductance equation

Although most Amateurs are familiar with the well-known Wheeler equation for inductance,[1] and with its expression in another form,[2] neither is readily solvable when you want to find the number of turns required when you already know the inductance desired. Actually, in almost all Amateur applications, you first determine the inductance you need and then, with whatever wire and coil form you have available, you solve for the required number of turns. The following equation, derived from the Wheeler equation, allows you to do just that:

$$n = \frac{\frac{20L}{t} + \sqrt{\frac{400\,L^2}{t^2} + 18\,d^3L}}{d^2}$$

where n = number of turns
L = desired inductance in microhenries
d = coil diameter in inches
t = number of turns per inch (close or space wound)
t = 1/wire diameter, for close wound coils

example

d = 2.5 inches
t = 19 turns per inch
L = 145 microhenries

The number of turns will come out to be 65.

Although at first glance the equation looks formidable, it really isn't. Using a small calculator, it can be solved in a few minutes. And as the equation shown is derived from the Wheeler equation, it naturally has the same constraints as to accuracy.*

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William Vissers, K4KI

*The familiar Wheeler equation that Bill referred to, that appears in reference 2 is reproduced below:

$$L_{uh} = \frac{a^2n^2}{9a + 10b}$$

where L = inductance in microhenries
a = coil radius in inches
b = coil length in inches

The practical difficulty in applying the more commonly used equation is that wire size (diameter) has to be carefully considered. If it is not, contradictory values of b will be obtained. Bill's developed equation eliminates that problem. And as stated, his equation is correct for both closely and widely spaced wound coils. — **Ed.**

ham radio

simple filter tester

I build all of my projects using a minimum of simple homebrew test gear. Generally, all that's necessary is a grid-dip oscillator, signal generator, VTVM, and an inductance bridge.

Because my VTVM with an RF probe isn't sensitive enough to pick up weak signals for aligning IFs or checking crystal filter frequencies, I built a filter tester to provide this capability (see **fig. 1**).

To use, simply feed in a modulated signal (tuned to the desired frequency) from a signal generator (start with 2 volts output) to the IF strip or filter, then listen for the maximum signal output on the tester.

Ed Marriner, W6XM

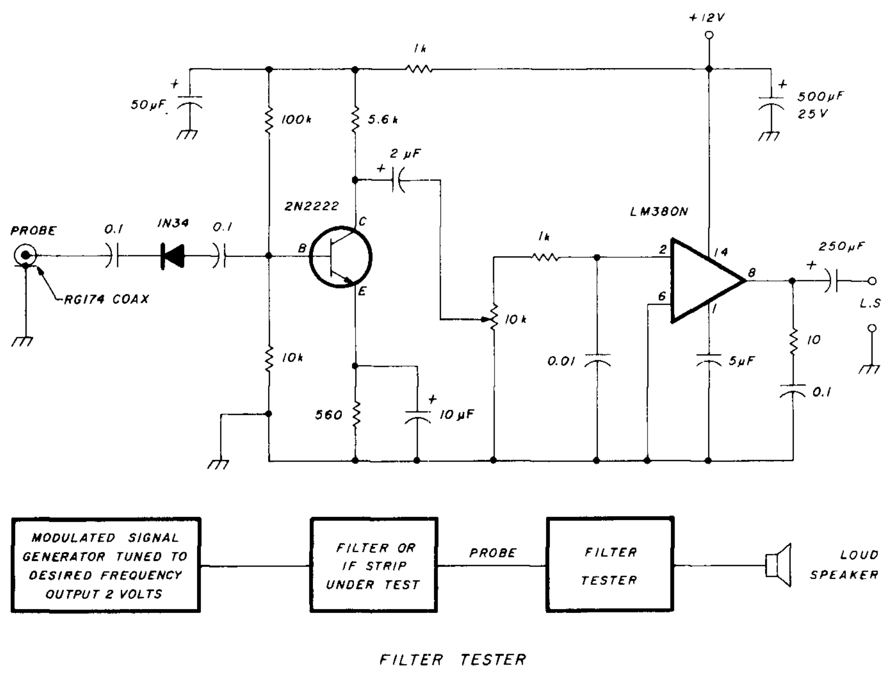

fig. 1. Filter tester.

TUBES

TYPE	PRICE	TYPE	PRICE	TYPE	PRICE
2C39/7289	$ 34.00	1182/4600A	$500.00	ML7815AL	$ 60.00
2E26	7.95	4600A	500.00	7843	107.00
2K28	200.00	4624	310.00	7854	130.00
3-500Z	102.00	4657	84.00	ML7855KAL	125.00
3-1000Z/8164	400.00	4662	100.00	7984	14.95
3B28/866A	9.50	4665	500.00	8072	84.00
3CX400U7/8961	255.00	4687	P.O.R.	8106	5.00
3CX1000A7/8283	526.00	5675	42.00	8117A	225.00
3CX3000F1/8239	567.00	5721	250.00	8121	110.00
3CW30000H7	1700.00	5768	125.00	8122	110.00
3X2500A3	473.00	5819	119.00	8134	470.00
3X3000F1	567.00	5836	232.50	8156	12.00
4-65A/8165	69.00	5837	232.50	8233	60.00
4-125A/4D21	79.00	5861	140.00	8236	35.00
4-250A/5D22	98.00	5867A	185.00	8295/PL172	500.00
4-400A/8438	98.00	5868/AX9902	270.00	8458	35.00
4-400B/7527	110.00	5876/A	42.00	8462	130.00
4-400C/6775	110.00	5881/6L6	8.00	8505A	95.00
4-1000A/8166	444.00	5893	60.00	8533W	136.00
4CX250B/7203	54.00	5894/A	54.00	8560/A	75.00
4CX250FG/8621	75.00	5894B/8737	54.00	8560AS	100.00
4CX250K/8245	125.00	5946	395.00	8608	38.00
4CX250R/7580W	90.00	6083/AZ9909	95.00	8624	100.00
4CX300A/8167	170.00	6146/6146A	8.50	8637	70.00
4CX350A/8321	110.00	6146B/8298	10.50	8643	83.00
4CX350F/8322	115.00	6146W/7212	17.95	8647	168.00
4CX350FJ/8904	140.00	6156	110.00	8683	95.00
4CX600J/8809	835.00	6159	13.85	8877	465.00
4CX1000A/8168	242.50*	6159B	23.50	8908	13.00
4CX1000A/8168	485.00	6161	325.00	8950	13.00
4CX1500B/8660	555.00	6280	42.50	8930	137.00
4CX5000A/8170	1100.00	6291	180.00	6L6 Metal	25.00
4CX10000D/8171	1255.00	6293	24.00	6L6GC	5.03
4CX15000A/8281	1500.00	6326	P.O.R.	6CA7/EL34	5.38
4CW800F	710.00	6360/A	5.75	6CL6	3.50
4D32	240.00	6399	540.00	6DJ8	2.50
4E27A/5-125B	240.00	6550A	10.00	6DQ5	6.58
4PR60A	200.00	6883B/8032A/8552	10.00	6GF5	5.85
4PR60B	345.00	6897	160.00	6GJ5A	6.20
4PR65A/8187	175.00	6907	79.00	6GK6	6.00
4PR1000A/8189	590.00	6922/6DJ8	5.00	6HB5	6.00
4X150A/7034	60.00	6939	22.00	6HF5	8.73
4X150D/7609	95.00	7094	250.00	6JG6A	6.28
4X250B	45.00	7117	38.50	6JM6	6.00
4X250F	45.00	7203	P.O.R.	6JN6	6.00
4X500A	412.00	7211	100.00	6JS6C	7.25
5CX1500A	660.00	7213	300.00*	6KN6	5.05
KT88	27.50	7214	300.00*	6KD6	8.25
416B	45.00	7271	135.00	6LF6	7.00
416C	62.50	7289/2C39	34.00	6LQ6 G.E.	7.00
572B/T160L	49.95	7325	P.O.R.	6LQ6/6MJ6 Sylvania	9.00
592/3-200A3	211.00	7360	13.50	6ME6	8.90
807	8.50	7377	85.00	12AT7	3.50
811A	15.00	7408	2.50	12AX7	3.00
812A	29.00	7609	95.00	12BY7	5.00
813	50.00	7735	36.00	12JB6A	6.50

NOTE * = USED TUBE

NOTE P.O.R. = PRICE ON REQUEST

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5.595-.500/4, 5.595-.500/4/CW
4 pole 500 cycles wide CW. Impedance 800ohms 15pf In/800ohms 0pf out. 19.99

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Lower sideband. (ZL) 19.99

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SDK	SCH-113A	11.2735MHz	10.00
TAMA	TF-31H250	CF 3179.3KHz	19.99
TYCO/CD	001019880	10.7MHz 2pole 15KHz bandwidth	5.00
MOTOROLA	4884863B01	11.7MHz 2pole 15KHz bandwidth	5.00
PTI	5350C	12MHz 2pole 15KHz bandwidth	5.00
PTI	5426C	21.4MHz 2pole 15KHz bandwidth	5.00
PTI	1479	10.7MHz 8pole bandwidth 7.5KHz at 3dB, 5KHz at 6dB	20.00
COMTECH	A10300	45MHz 2pole 15KHz bandwidth	6.00
FRC	ERXF-15700	20.6MHz 36KHz wide	10.00
FILTECH	2131	CF 7.825MHz	10.00

****

CERAMIC FILTERS

AXEL	4F449	12.6KC Bandpass Filter 3dB bandwidth 1.6KHz from 11.8-13.4KHz	10.00
CLEVITE	TO-01A	455KHz+-2KHz bandwidth 4-7% at 3dB	5.00
	TCF4-12D36A	455KHz+-1KHz bandwidth 6dB min 12KHz, 60dB max 36KHz	10.00
MURATA	BFB455B	455KHz	2.50
	BFB455L	455KHz	3.50
	CFM455E	455KHz +-5.5KHz at 3dB , +-8KHz at 6dB , +-16KHz at 50dB	6.65
	CFM455D	455KHz +-7KHz at 3dB , +-10KHz at 6dB , +-20KHz at 50dB	6.65
	CFR455E	455KHz +-5.5KHz at 3dB , +-8KHz at 6dB , +-16KHz at 60dB	8.00
	CFU455B	455KHz +-2KHz bandwidth +-15KHz at 6dB, +-30KHz at 40dB	2.90
	CFU455C	455KHz +-2KHz bandwidth +-12.5KHz at 6dB , +-24KHz at 40dB	2.90
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	CFU455I	455KHz +-1KHz bandwidth +-2KHz at 6dB , +-6KHz at 40dB	2.90
	CFW455D	455KHz +-10KHz at 6dB , +-20KHz at 40dB	2.90
	CFW455H	455KHz +-3KHz at 6dB , +-9KHz at 40dB	2.90
	SFB455D	455KHz	2.50
	SFD455D	455KHz +-2KHz , 3dB bandwidth 4.5KHz +-1KHz	5.00
	SFE10.7MA	10.7MHz 280KHz +-50KHz at 3dB , 650KHz at 20dB	2.50
	SFE10.7MS	10,7MHz 230KHz +-50KHz at 3dB , 570KHz at 20dB	2.50
	SFG10.7MA	10.7MHz	10.00
NIPPON	LF-B4/CFU455I	455KHz +-1KHz	2.90
	LF-B6/CFU455H	455KHz +-1KHz	2.90
	LF-B8	455KHz	2.90
	LF-C18	455KHz	10.00
TOKIN	CF455A/BFU455K	455KHz +-2KHz	5.00
MATSUSHIRA	EFC-L455K	455KHz	7.00

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2N1561	$25.00	2N5920	$ 70.00	40608 RCA	$ 2.48	BFY90	$ 1.50
2N1562	25.00	2N5921	80.00	40673 RCA	2.50	BLW60C5	15.00
2N1692	25.00	2N5922	10.00	40894 RCA	1.00	BLX67	12.25
2N2857	1.55	2N5923	25.00	60247 RCA	25.00	BLX67C3	12.25
2N2857JAN	4.10	2N5941	23.00	61206 RCA	100.00	BLX93C3	22.21
2N2857JANTX	4.50	2N5942	40.00	62800A RCA	60.00	BLY87A	7.50
2N2876	13.50	2N5944	10.35	62803 RCA	100.00	BLY88C3	13.08
2N2947	18.35	2N5945	10.00	430414/3990RCA	50.00	BLY89C	13.00
2N2948	13.00	2N5946	12.00	3457159 RCA	20.00	BLY90	45.00
2N2949	15.50	2N5947	9.20	3729685-2 RCA	75.00	BLY92	13.30
2N3118	5.00	2N6080	6.00	3729701-2 RCA	50.00	BLY94C	45.00
2N3119	4.00	2N6081	7.00	3753883 RCA	50.00	BLY351	10.00
2N3134	1.15	2N6082	9.00	615467-902	25.00	BLY568C/CF	30.00
2N3287	4.90	2N6083	9.50	615467-903	40.00	C2M70-28R	92.70
2N3288	4.40	2N6084	12.00	2SC568	2.50	C25-28	57.00
2N3309	4.85	2N6094	11.00	2SC703	36.00	C4005	2.50
2N3375	17.10	2N6095	12.00	2SC756A	7.50	CD1659	20.00
2N3478	2.13	2N6096	16.10	2SC781	2.80	CD1899	20.00
2N3553	1.55	2N6097	20.70	2SC1018	1.00	CD1920	10.00
2N3553JAN	2.90	2N6105	21.00	2SC1042	24.00	CD2188	18.00
2N3632	15.50	2N6136	21.85	2SC1070	2.50	CD2545	24.00
2N3733	11.00	2N6166	40.24	2SC1216	2.50	CD2664A	16.00
2N3818	5.00	2N6267	142.00	2SC1239	2.50	CD3167	92.70
2N3866	1.30	2N6304	1.50	2SC1251	24.00	CD3353	95.00
2N3866JAN	2.20	2N6368	30.00	2SC1306	2.90	CD3435	26.30
2N3866JANTX	3.80	2N6439	55.31	2SC1307	5.50	CD3900	152.95
2N3866JANTXV	4.70	2N6459	18.00	2SC1424	2.80	CM25-12	20.00
2N3866AJANTXV	5.30	2N6567	10.06	2SC1600	5.00	CM40-12	27.90
2N3924	3.35	2N6603	13.50	2SC1678	2.00	CM40-28	56.90
2N3926	16.10	2N6604	13.50	2SC1729	32.40	CME50-12	30.00
2N3927	17.25	2N6679	44.00	2SC1760	1.50	CTC2001	42.00
2N3948	1.75	2N6680	80.00	2SC1909	4.00	CTC2005	55.00
2N3950	25.00	021-1	15.00	2SC1945	10.00	CTC3005	70.00
2N3959	3.85	01-80703T4	65.00	2SC1946	40.00	CTC3460	20.00
2N4012	11.00	35C05	15.00	2SC1947	10.00	DV2820S	25.00
2N4037	2.00	102-1	28.00	2SC1970	2.50	DXL1003P70	22.00
2N4041	14.00	103-1	28.00	2SC1974	4.00	DXL2001P70	19.00
2N4072	1.80	103-2	28.00	2SC2166	5.50	DXL2002P70	14.00
2N4080	4.53	104P1	18.00	2SC2237	32.00	DXL3501AP100F	47.00
2N4127	21.00	163P1	10.00	2SC2695	47.00	EFJ4015	12.00
2N4416	2.25	181-3	15.00	A2X1698	POR	EFJ4017	24.00
2N4427	1.25	210-2	10.00	A3-12	14.45	EFJ4021	24.00
2N4428	1.85	269-1	18.00	A50-12	24.00	EFJ4026	35.00
2N4430	11.80	281-1	15.00	A209	10.00	EN15745	20.00
2N4927	3.90	282-1	30.00	A283	6.00	FJ9540	16.00
2N4957	3.45	482	7.50	A283B	6.00	FSX52WF	58.00
2N4959	2.30	564-1	25.00	A1610	19.00	G65739	25.00
2N5016	18.40	698-3	15.00	AF102	2.50	G65386	25.00
2N5026	15.00	703-1	15.00	AFY12	2.50	GM0290A	2.50
2N5070	18.40	704	4.00	AR7115	20.00	HEP76	4.95
2N5090	13.80	709-2	11.00	AT41435-5	6.35	HEPS3002	11.40
2N5108	3.45	711	4.00	B2-8Z	10.70	HEPS3003	30.00
2N5109	1.70	733-2	15.00	B3-12	10.85	HEPS3005	10.00
2N5160	3.45	798-2	25.00	B12-12	15.70	HEPS3006	19.90
2N5177	21.62	3421	28.00	BAL0204125	152.95	HEPS3007	25.00
2N5179	1.04	3683P1	15.00	BF25-35	56.25	HEPS3010	11.34
2N5216	56.00	3992	25.00	B40-12	19.25	HF8003	10.00
2N5470	75.00	4164P1	15.00	B70-12	55.00	HFET2204	112.00
2N5583	3.45	4243P1	28.00	BF272A	2.50	HP35821	38.00
2N5589	9.77	4340P3	18.00	BFQ85	2.50	HP35826B	32.00
2N5590	10.92	4387P1	27.50	BFR21	2.50	HP35826E	32.00
2N5591	13.80	7104-1	28.00	BFR90	1.00	HP35831E	30.00
2N5596	99.00	7249-2	10.50	BFR91	1.65	HP35832E	50.00
2N5636	12.00	7283-1	37.50	BFR99	2.50	HP35833E	50.00
2N5637	15.50	7536-1	30.00	BFT12	2.50	HP35859E	75.00
2N5641	12.42	7794-1	10.50	BFW16A	2.50	HP35866E	44.00
2N5642	14.03	7795	15.00	BFW17	2.50	HXTR2101	44.00
2N5643	25.50	7795-1	15.00	BFW92	1.50	HXTR3101	7.00
2N5645	13.80	7796-1	24.00	BFX44	2.50	HXTR5101	31.00
2N5646	20.70	7797-1	36.00	BFX48	2.50	HXTR6104	68.00
2N5651	11.05	40081 RCA	5.00	BFX65	2.50	HXTR6105	31.00
2N5691	18.00	40279 RCA	10.00	BFX84	2.50	HXTR6106	33.00
2N5764	27.00	40280 RCA	4.62	BFX85	2.50	J310	1.00
2N5836	3.45	40281 RCA	10.00	BFX86	2.50	J02000	10.00
2N5842	8.45	40282 RCA	20.00	BFX89	1.00	J02001	25.00
2N5847	19.90	40290 RCA	2.80	BFY11	2.50	J04045	24.00
2N5849	20.00	40292 RCA	13.05	BFY18	2.50	KD5522	25.00
2N5913	3.25	40294 RCA	2.50	BFY19	2.50	KJ5522	25.00
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M1131	5.15	MRF464	25.30	NE021350	5.30	SD1009-2	15.00
M1132	7.25	MRF466	18.97	NE13783	61.00	SD1012	10.00
M1134	13.40	MRF472	1.50	NE21889	43.00	SD1012-3	10.00
M9116	29.10	MRF475	3.10	NE57835	5.70	SD1012-5	10.00
M9579	6.00	MRF476	3.16	NE64360ER-A	100.00	SD1013	10.00
M9580	7.95	MRF477	20.00	NE64480 (B)	94.00	SD1013-3	10.00
M9587	7.00	MRF479	8.05	NE73436	2.50	SD1013-7	10.00
M9588	5.20	MRF492	23.00	NE77362ER	100.00	SD1016	15.00
M9622	5.95	MRF502	1.04	NE98260ER	100.00	SD1016-5	15.00
M9623	7.95	MRF503	6.00	PRT8637	25.00	SD1018-4	13.00
M9624	9.95	MRF504	7.00	PT3127A	5.00	SD1018-6	13.00
M9625	15.95	MRF509	5.00	PT3127B	5.00	SD1018-7	13.00
M9630	14.00	MRF511	10.69	PT3127C	20.00	SD1018-15	13.00
M9740	27.90	MRF515	2.00	PT3127D	20.00	SD1020-5	10.00
M9741	27.90	MRF517	2.00	PT3127E	20.00	SD1028	15.00
M9755	16.00	MRF525	3.45	PT3190	20.00	SD1030	12.00
M9780	5.50	MRF559	1.76	PT3194	20.00	SD1030-2	12.00
M9827	11.00	MRF587	11.00	PT3195	20.00	SD1040	5.00
M9848	35.00	MRF605	20.00	PT3537	7.80	SD1040-2	20.00
M9850	13.50	MRF618	25.00	PT4166E	20.00	SD1040-4	10.00
M9851	20.00	MRF626	12.00	PT4176D	25.00	SD1040-6	5.00
M9860	8.25	MRF628	8.65	PT4186B	5.00	SD1043	12.00
M9887	2.80	MRF629	3.45	PT4209	25.00	SD1043-1	10.00
M9908	6.95	MRF641	25.30	PT4209C/5645	25.00	SD1045	3.75
M9965	12.00	MRF644	27.60	PT4556	24.60	SD1049-1	2.00
MM1500	25.00	MRF646	29.90	PT4570	7.50	SD1053	4.00
MM1550	10.00	MRF648	33.35	PT4577	20.00	SD1057	10.00
MM1552	50.00	MRF816	15.00	PT4590	5.00	SD1065	4.75
MM1553	50.00	MRF823	20.00	PT4612	20.00	SD1068	15.00
MM1607	8.45	MRF846	44.85	PT4628	20.00	SD1074-2	18.00
MM1614	10.00	MRF892	35.50	PT4640	20.00	SD1074-4	28.00
MM1810	15.00	MRF894	46.00	PT4642	20.00	SD1074-5	28.00
MM1810	15.00	MRF901 3 Lead	1.00	PT5632	4.70	SD1076	18.50
MM1943	1.80	MRF901 4 Lead	2.00	PT5749	25.00	SD1077	4.00
MM2608	5.00	MRF902/2N6603JAN	15.00	PT6612	25.00	SD1077-4	4.00
MM3375A	17.10	MRF902B	18.40	PT6619	20.00	SD1077-6	4.00
MM4429	10.00	MRF904	2.30	PT6708	25.00	SD1078-6	24.00
MM8000	1.15	MRF905	2.55	PT6709	25.00	SD1080-7	7.50
MM8006	2.30	MRF911	2.50	PT6720	25.00	SD1080-8	6.00
MM8011	25.00	MRF965	2.55	PT8510	15.00	SD1080-9	3.00
MPSU31	1.01	MRF966	3.55	PT8524	25.00	SD1084	8.00
MRA2023-1.5	42.50	MRF1000MA	32.77	PT8609	25.00	SD1087	15.00
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MRF136	16.00	MRF2001	41.74	PT8639	25.00	SD1088-8	22.00
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MRF208	11.50	MRF5176	24.00	PT8679	25.00	SD1090	15.00
MRF212	16.10	MRF8004	2.10	PT8708	20.00	SD1094	15.00
MRF221	10.00	MSC1720-12	225.00	PT8709	20.00	SD1095	15.00
MRF223	13.00	MSC1821-3	125.00	PT8727	29.00	SD1098-1	30.00
MRF224	13.50	MSC1821-10	225.00	PT8731	25.00	SD1100	5.00
MRF227	3.45	MSC2001	30.00	PT8742	19.10	SD1109	18.00
MRF230	2.00	MSC2010	93.00	PT8787	25.00	SD1115-2	7.50
MRF231	10.00	MSC2223-10	245.00	PT8828	25.00	SD1115-3	7.50
MRF232	12.07	MSC2302	POR	PT9700	25.00	SD1115-7	2.10
MRF237	3.15	MSC3000	35.00	PT9702	25.00	SD1116	5.00
MRF238	13.80	MSC3001	38.00	PT9783	16.50	SD1118	22.00
MRF239	17.25	MSC72002	POR	PT9784	32.70	SD1119	5.00
MRF245	35.65	MSC73001	POR	PT9790	56.00	SD1124	50.00
MRF247	31.00	MSC80064	35.00	PT31083	20.00	SD1132-1	15.00
MRF304	36.00	MSC80091	10.00	PT31962	20.00	SD1132-4	12.00
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MRF313	11.15	MSC80593	POR	RE3754	25.00	SD1133-1	10.00
MRF314	29.21	MSC80758	POR	RE3789	25.00	SD1134-1	2.50
MRF315	28.86	MSC82001	33.00	RF35	16.00	SD1134-4	12.00
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SD1219-5	15.00	SD1347-1	1.00	SD1484	1.50	SRF2265	100.00	SD1278	13.75	SD1429-5	15.00	SD1575	6.95	TA8561	15.00
SD1219-8	15.00	SD1365-1	2.50	SD1484-5	1.50	SRF2281	5.00	SD1278-1	13.75	SD1430	12.00	SF4557	25.00	TA8562	15.00
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SD1225	18.00	SD1380-7	1.00	SD1499-1	36.00	SRF2597	25.00	SD1283-3	10.00	SD1442	15.00	SRF750	36.00	TP1028	15.00
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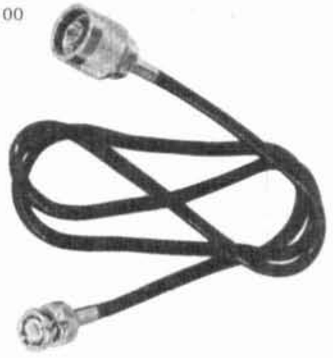

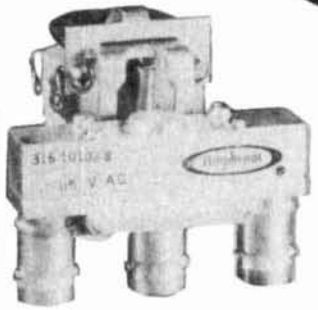

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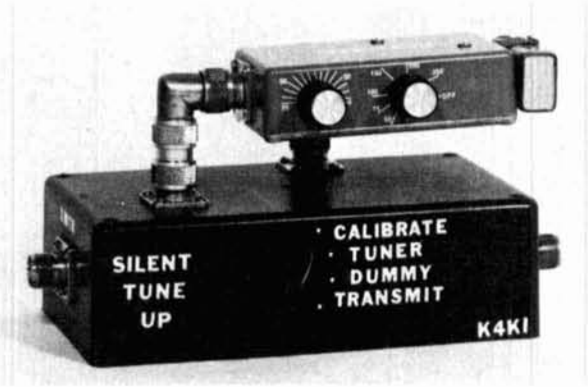
the
weekender

a safe, silent tuneup



A safe, silent, tuneup — without any radiation of power, and with your SWR at 1:1 at the same time, is readily possible with the unit described in this article.

One of the most useful pieces of equipment is an accurate RF bridge; with the advent of small, inexpensive noise bridges, a definite need has been fulfilled. In an effort to obtain maximum utility from such a bridge, the following circuit was developed and built. (The incorporated Palomar noise bridge was loaned to me by my buddy, K4YS, whose comments and advice are a part of this article.)

The noise bridge is mounted "piggy back" on a small chassis box using ordinary coaxial fittings. This mounting provides rigidity, eliminates unnecessary connection cables, and results in a single integrated unit. The circuit is best understood by referring to the schematic of **fig. 1**. S1 (A, B, C, D) is an old CRL, PA-1013, 2-gang, four-circuit rotary switch from my junk box. However, similar switches are available from various manufacturers in sizes appropriate for the power you plan to use. My own Yaesu FT-101-B worked very well with the unit built and described. A defeat circuit section of the switch, S1D, prevents you from accidentally turning on your transmitter when you are in the Calibrate and Tuner position, and possibly damaging your noise bridge.

With the rotary switch, S1B, in the A (Calibrate) position, a 50-ohm 1/2-watt carbon resistor is connected to the UNKNOWN jack of the noise bridge. The RECEIVER jack of the noise bridge is switched through S1A to the receiver input. The bridge is balanced by means of the reactance and resistance knobs of the noise bridge for a noise null. Now, when the rotary switch is thrown to the B (Tuner) position, and without touching the already nulled noise bridge control knobs, the antenna tuner is adjusted for a noise null. The tuner is now properly set for a 50-ohm input.

With the rotary switch set to the C (Dummy Load) position, your transmitter can now be properly tuned and loaded for 50 ohms, the nominal value of resistance of your dummy load. And now with the rotary switch thrown in the D (Transmit) position, you are ready to go on the air. As in all RF power switching functions, it is dangerous to engage the rotary switch with the RF power on because of possible transients. However, the transmitter key-up provides an easy way to control transmitter power, and the series defeat circuit of switch S1D provides an additional safety circuit as previously mentioned.

Although a 50 ohm, 1/2-watt carbon resistance is shown for R1, I found that a 47 or 51 ohms value would work as well.

To avoid cutting into the Palomar noise bridge, battery leads to connect to power switch S1C, a couple of battery connectors matching the ones on the noise bridge were connected in order to snap into the circuit as shown. This battery turn-off circuit may at first glance seem to be somewhat redundant, as the resistance knob of the noise bridge has a turn-off switch when the knob is rotated fully counter clockwise. However, if you were to do this, you would unbalance your noise bridge, and if you later switched the bridge back on, the unbalanced condition would make a loud and uncomfortable noise. The bridge could be rebalanced to eliminate this noise, but the power turn-off switch makes operation easier.

My own experience has been that after you're properly loaded up, and your antenna tuner is properly set, your SWR is 1:1, or so close to it that you do not have to do any of the "tweaking" sometimes found necessary with other systems. Safety is quite important; if you keep your transmitter SWR down to 1:1, you'll avoid the arc over problems that sometimes occur when your tune-up SWR gets out of control.

No problems were encountered in the building or testing of the unit. The jacks J1 through J5 were internally connected together with a piece of No. 12 copper antenna wire to provide a common ground at the shells. As a precaution, the leads from switches S1C and S1D were twisted to prevent RF pickup, although none was evident at any time.

I'll be pleased to answer all questions — just send along an SASE to me at 1245 S. Orlando Avenue, Cocoa Beach, Florida 32931.

By William Vissers, K4KI, 1245 S. Orlando Avenue, Cocoa Beach, Florida 32931

item	description
J1-5	SO-239 UHF panel receptacle
R1	50-ohm non-inductive resistor
S1	4-section, 4 position rotary switch
chassis	7-3/4 × 4-1/4 × 2-1/2 inches

Two PL-259 coax adapters, double male; two 646 or 83-1AP coax right angle adapters (used to mount Palomar noise bridge to J4 and J5 as shown in photograph).

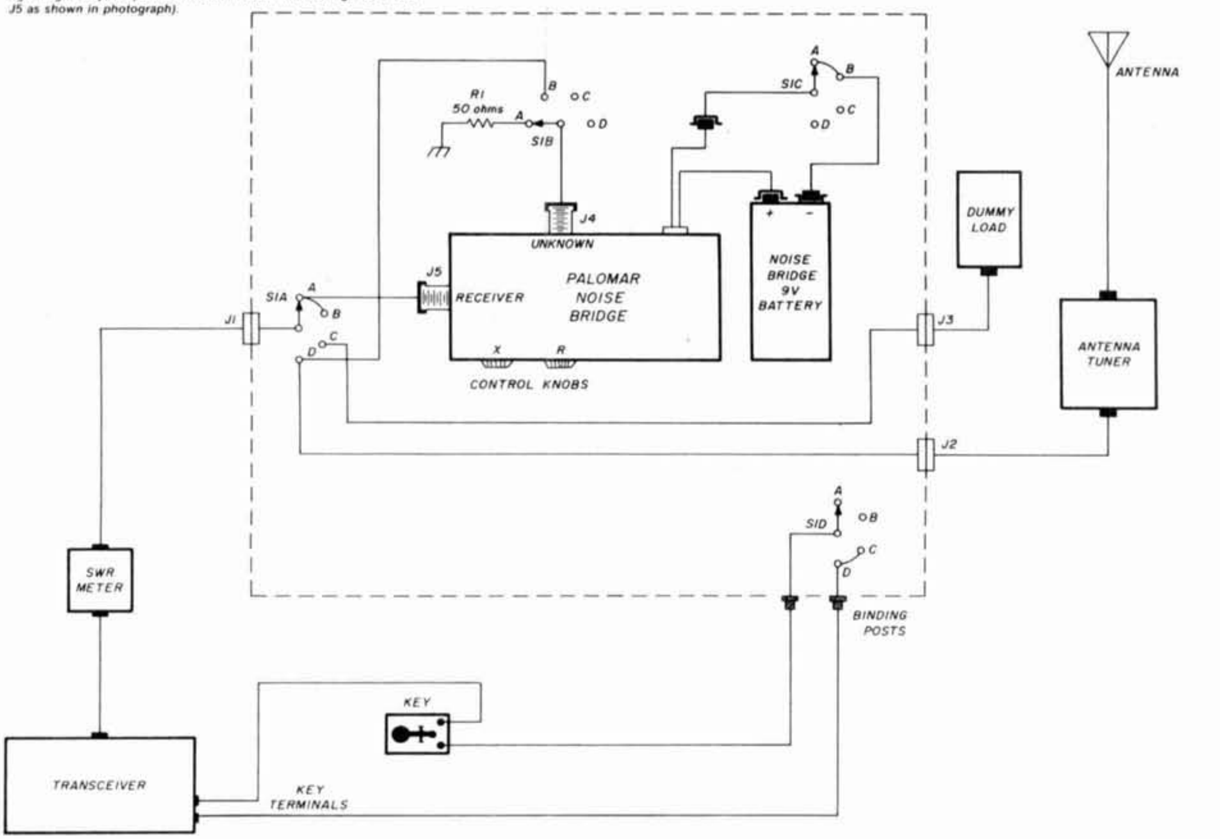


fig. 1. Silent tune-up accessory unit schematic and interconnection diagram.

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1980-1984

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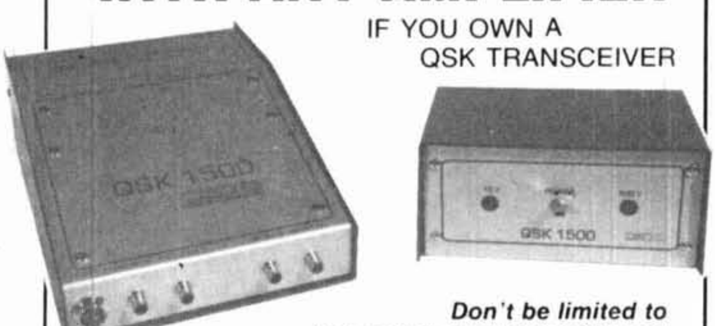

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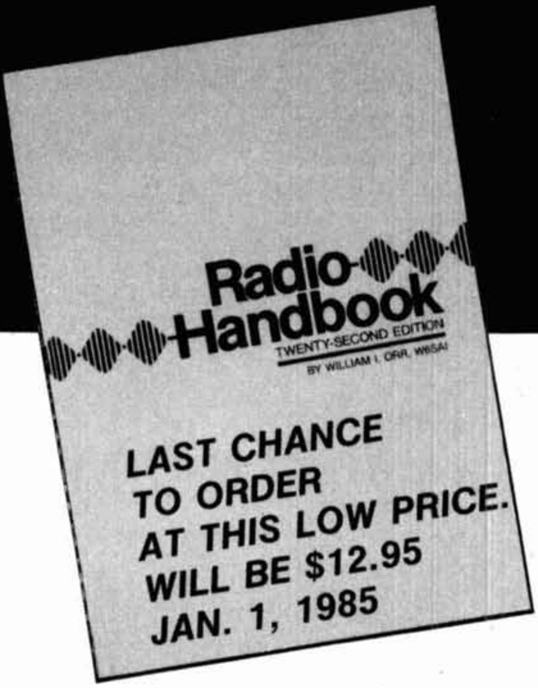

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NEW products

mobile encoder

The new Model SE-714 mobile encoder from Cetec Vega, when triggered, produces a 3-to-6-digit DTMF ANI (automatic number identification) message plus a 2-digit status message. The ANI message is field-programmable by means

of a diode matrix. The status message is operator-programmable by means of two panel-mounted back-lit thumbwheel switches, allowing up to 100 different status messages.

The SE-714 has two timing features. One feature inhibits the message for a timed period after message transmission, allowing short voice responses without the message. The other ("stuck-mic") feature automatically sends the message after a timed period of continuous transmission.

The SE-714 expands channel capacity by replacing mobile-to-base voice responses with a complete ANI plus status message that requires only a 3/4-second transmission time. The message may be initiated by PTT (or off-hook) operation or by a manual pushbutton on the front panel. The messages can be displayed on any of several types of DTMF displays available from Cetec Vega.

For additional information, contact Cetec Vega, 9900 Baldwin Place, El Monte, California 91731.

Circle #301 on Reader Service Card.

adhesive/sealant and dielectric compound

General Electric Silicone Products Division is introducing RTV-108 Adhesive/Sealant and G-635 Dielectric Compound for the protection of components from the corrosive effects of water, ozone, oxidation, and chemicals and to keep RFI leakage to a minimum.

RTV-108 is a paste-like sealant that cures at room temperature to a flexible silicone rubber. When poured over connectors, taps, and other components, RTX-108 renders them virtually immune to the elements. The product is preferable to heat shrink tubing, especially in underground systems where the equipment is constantly exposed to water.

Because of silicone's unique molecular structure, RTV-108 bonds easily to most clean surfaces without the aid of primers. The product can withstand temperatures from −70 to 400 degrees F without becoming brittle or melting.

G-635 Dielectric Compound, unlike RTV (room temperature vulcanizing) sealants, maintains its soft-to-medium consistency when exposed to the environment. The product is used as a lubricant on aluminum connector threads to prevent seize-up due to oxidation or the bonding of dissimilar metals. When applied as a grease to F fittings, taps, traps, ground blocks, feeder/connector gaps and house splitters, G-635 protects them from the attacks of airborne pollutants such as salt, moisture, and acid rain.

G-635 Dielectric Compound maintains its spreadability at −100 degrees F and shows minimal bleed up to 450 degrees F, making it an ideal lubricant under adverse conditions. A light coat of G-635 on amplifier gaskets keeps them from drying out and splitting, thereby insuring that RFI leakage is kept to a minimum. The product is also used on high-voltage industrial, automotive, and aircraft applications.

For more information about GE silicone rubbers and lubricants, contact Silicone Products Division, General Electric, Waterford, New York 12188.

Circle #302 on Reader Service Card.

legal limit amplifier

The Titan 425 Linear Amplifier from Ten-Tec delivers the full new legal power limit of 1500 watts PEP SSB output and 1500 watts of full break-in power for QSK, CW or AMTOR. This cool-running design consists of two sections, an amplifier and a power supply. Styled to match modern transceivers, and extremely compact for its ratings, the amplifier contains all operating controls and indicators. The power supply is housed in a utility type hide-away enclosure.

The amplifier utilizes two EIMAC 3CX800A7 triodes in a ducted forced air system, operating in a grounded grid configuration easily provide rated output power with up to 65 percent efficiency. Maximum input power of 3 kW (2 kW CCS) requires only 100 watts of drive power. A high-low plate voltage switch assures optimum efficiency at lower power (1 kW output) for tune-up or RTTY and SSTV. Three LED status indicators display Standby, Wait, or Operate mode and a fourth LED alerts you when the input is overdriven. Two panel meters provide full-time indication of plate current and switch-selected choice of plate voltage, grid current, forward power or reflected power. Peak power is indicated on a ten-element LED bar graph display. Band coverage of 1.8 through 23 MHz Amateur bands is standard. Ten-Tec's export model extends coverage to 29.7 MHz.

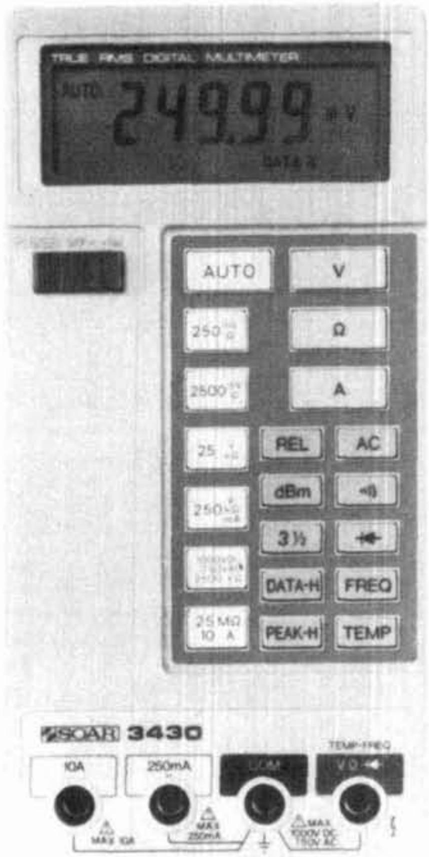
The power supply is conservatively designed for cool operation under full load; the circuit employs a tape-wound Hypersil transformer for minimum weight and size. Primary power of 220-250 volts at 20 amperes is standard. 115 volt operation is possible but not recommended when operating full power. Fuses are provided for primary circuit and plate current. Protective interlocks are included on the AC and high voltage lines.

The Amateur net price of the Titan 425 Linear Amplifier is $2,485.00.

For complete information, contact Ten-Tec, Inc., Sevierville, Tennessee 37862.

multi-function DMM

Designed and developed by North American SOAR Corp., the Model 3430 uses SOAR's new custom LSI chip plus two 4-bit CPU's to achieve 13 different functions and three operational



modes. Unlike the traditional 20,000 count units presently available, the Model 3430 is a 25,000 count instrument. This allows full 4-1/2 digit resolution to the second decimal place for readings such as 240.15 VAC and resistance to 25 megohms. The unit measures True RMS for both AC voltage and current. The unit is priced at $339.

A second DMM, Model 3450, similar to the 3430, is priced at $299.

For details, contact North American SOAR Corp., 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #303 on Reader Service Card.

NEW products

climber's belt

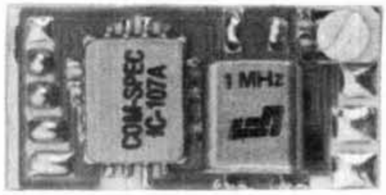
The North Shore Ham Services' Tower Climber's Belt is an all-nylon general purpose safety belt with a 3-inch wide nylon cushion pad and a 1-3/4 inch wide nylon buckle strap. Brass grommet reinforced buckle holes accommodate a single tongue buckle. A 1/2-inch diameter nylon rope lanyard is 3 feet long, making it well suited to use on ham radio towers. The price is $50.00 postpaid.

For additional information, contact North Shore Ham Services, P.O. Box 54, W. Lynn, Massachusetts 01905.

Circle #304 on Reader Service Card.

super-small CTCSS encoder

Communications Specialists of Orange, California, recently announced what it says is now the smallest CTCSS encoder available. The SS-32HB measures only 0.5 × 1.0 × 0.15 inches and will fit into any portable requiring send-only CTCSS. The unit may be programmed to any of the 32 standard sub-audible tone frequencies by bridging solder pads on the board. The price is $29.95.



For further information, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665.

Circle #305 on Reader Service Card.

RTTY, SSTV, and Morse programs

A hardware-free RTTY program for Radio Shack™ Model III and IV computers is available from Woodall & Associates. SOFFTY 1.2 provides the most commonly used configuration for RTTY on Amateur bands: 60 and 100 WPM with a 170 Hz tone shift. Keyboard selection of high or low receive tones, as well as normal and inverted code reception are also provided.

SOFTSCAN 1.1 is a no-hardware SSTV decoder and print program for high-resolution hard-copy printouts on the Epson RX series of printers, or on MX's with graphic options.

COMPCODE 1.1B, a hardware-less Morse code program, connects a communications receiver to a TRS-80™ Model I, III, or IV via the computer cassette interface and cables. Instructions for connecting the receiver to the computer are written into the program itself.

SOFFTY 1.2 sells for $34.95; SOFTSCAN 1.1 for $19.95; and COMPCODE 1.1B for $24.95. Shipping and handling charges are additional.

For complete information, contact Woodall & Associates, P.O. Box 284, Plainfield, Indiana 46168.

Circle #306 on Reader Service Card.

SWL catalog

A new SWL catalog has recently been released by Spectronics describing its complete line of name-brand receivers, antennas, accessories, and publications for the short-wave listener. Written in a non-technical, conversational manner, it is an ideal tool to introduce the beginner to short wave, yet diverse enough to appeal to the more serious listener.

For a free copy, contact Spectronics, Inc., 1009 Garfield Street, Oak Park, Illinois 60304-1890.

dielectric resonator oscillators

Microwave Research & Manufacturing has announced its new line of dielectric resonator oscillators covering 2 GHz to 20 GHz, with standard output powers of up to +23 dBm. Stability is better than 3 PPM per degree over a temperature range of −54 to +100 degrees Celsius. Phase locking ability is optional.

Phase noise is typically −93 dBc at 10 kHz and frequency pulling is less than 0.3 MHz into an output VSWR of 1.5:1. The mechanical tuning range is 8 percent for most models with electrical tuning optional. Power requirements are 5 to 28 VDC and offer high efficiencies of typically 20 percent. Harmonic content is −40 dBc. The size of model MW-7 is extremely small, measuring only 1.42 x 0.76 x 0.600 inches.

Millimeter wave sources are also available using F.E.T./D.S.O. doubling techniques yielding highly stable millimeter wave sources of high efficiencies and small size.

Microwave Research & Manufacturing is a supplier of microwave and millimeter wave FET and Gunn diode dielectrically stabilized oscillators, VCOs and downconverters.

For additional information, contact D.C. Mitchell, Microwave Research & Manufacturing Company, 1 Cider Mill Lane, Upton, MA 01568.

Circle #307 on Reader Service Card.

two-digit DTMF decoder

A new two-digit sequential DTMF decoder is available from Palomar Engineers. The decoder, Model P-202, features 0.5 ampere 115 volt relay contacts, dual bandpass filters, quartz crystal frequency control, as well as operation over an ex-

NEW products

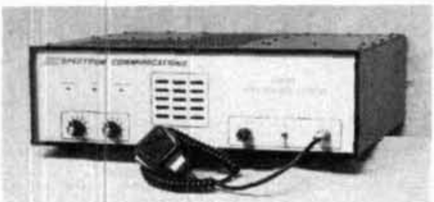
tended temperature range and over a wide range of signal levels. It is highly immune to false triggering. Operation is from 12 to 30 volts DC. DTMF coding is factory-set to order but easily changed by moving circuit board jumpers. The unit is priced at $125.

For more information, contact Palomar Engineers, 1924-F West Mission Road, Escondido, California 92025.

Circle #308 on Reader Service Card.

desktop/portable repeater

The SCR77D from Spectrum Communications is a new "Decktop/Portable" Repeater. The first of its kind on the Amateur market, the 77D was developed in response to customer demand over the past several years. Its compact, low-power configuration makes it suitable for "local" use (within a 0-20 mile radius, depending on antenna and terrain). It may be used at a fixed location, or Portable/Mobile.



Powered by a 12 volt battery pack, the SCR77D can be mounted in a vehicle parked on a hilltop for temporary coverage as needed. Autopatch and 'PL' are included. An AC power supply, as well as jacks for 12 VDC power, are built in. Full duplex base station applications, such as computer data links or export "rural telephone," are also ideal for the SCR77D. Standard models include 10 watt UHF unit with built-in duplexer and a 15-watt VHF unit with external duplexer.

For details, contact Spectrum Communications Corporation, 1055 W. Germantown Parkway, Norristown, Pennsylvania 19401-9616.

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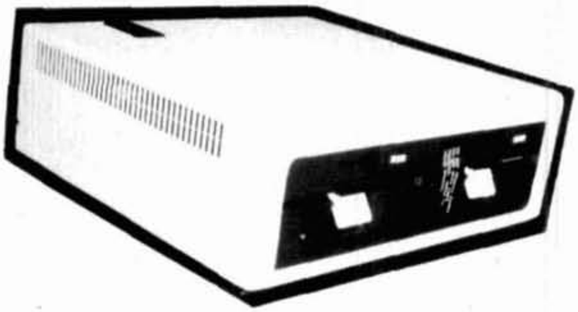

BRAND NEW, single sided, dual floppy disc drives made for Digital Equipment Corp. (DEC). This beautiful piece of computer hardware consists of 2 Shugart compatible TEAC 40 track, double density, 5¼" mini-floppy disc drives brand new in the case with their own regulated, switching power supply, cooling fan & on/off switch. Each unit also comes with a line cord & documentation. These were made for DEC, but are also compatible with other personal computers such as IBM, TRS 80 models I, II, & the Color Computer, and other Shugart compatible interfaces. Naturally, you supply the cables and disc controller card to suit your particular system. The RX-180 AB runs off of 115/230 VAC 50/60 Hz. w/out any modifications to the drives. Each system comes in the original factory box and are guaranteed functional. A blockbuster of a buy !!

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These compact, lightweight switching power supply boards were originally made for the Texas Instrument 99/4A micro computer. It measures only 4-3/4 x 4-1/8 x 1-1/4" and puts our +5 VDC @ 1.2 A, −5 VDC @ 0.120 A, and +12 VDC @ .350 A. Input is 14.4 VAC to 21.6 VAC. A simple transformer or plug in the wall adapter will do. Our SPL 461-33 supply listed below is perfect for this board. Each one is new, individually packaged with data. For short money, you can get a nice little power supply or spare parts for your TI 99/4A. Shpg. Wt. 1 Lb.

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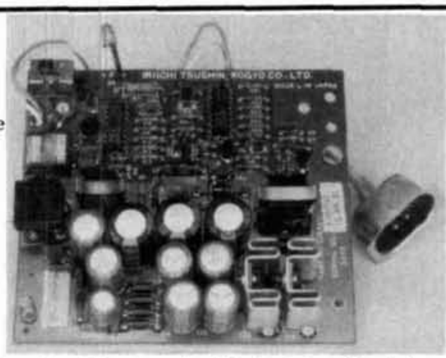

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Coming Events

ACTIVITIES

"Places to go..."

WISCONSIN: The "original" annual midwinter Swapfest, Saturday, January 12, 8 AM, Waukesha Co. Expo Center Forum, Milwaukee. Admission $2.00 advance, $3.00 door. Reserved 4' tables $3.00 prior to December 31. $4.00 at door. Sponsored by the West Allis Radio Amateur Club. For tickets: WARAC Swapfest, PO Box 1072, Milwaukee, WI 53201. SASE please.

MINNESOTA: The annual Handi-Ham Winter Hamfest, Saturday, December 1, Eagles Club in Faribault. Registration 9 AM. Handi-Ham equipment auction. Dinner at noon. Talk in on 19/79. For information: Don Franz, WØFIT, 1114 Frank Avenue, Albert Lea, MN 56007.

INDIANA: South Bend Hamfest Swap & Shop, January 6, first Sunday after New Year's Day at Century Center downtown on US 33 Oneway North between St. Joseph Bank Building *and river. Industrial history Museum in same building.* Carpeted half acre room. Open tables $1 per ft. Four lane highways to door from all directions. Talk-in freq: 52-52, 99-39, 93-33, 78½18, 69-09, 145.29.

VIRGINIA: The 8th annual Richmond Frostfest, sponsored by the Richmond Amateur Telecommunications Society, Sunday, January 13, Virginia State Fairgrounds. 8:30 AM to 3:30 PM. General admission $4.00. Flea market spaces $3.00 without table; $7.00 with 8' table. Entire show indoors. Deadline for booth, December 30; flea market, January 10. Building open Saturday afternoon for setup. Write Richmond Frostfest, PO Box 1070, Richmond, VA 23208 or call Bill Scruggs, N4DDM (804) 272-8206.

OPERATING EVENTS

"Things to do..."

THE SANDY RIVER ARC will operate the Chester N. Greenwood special event station in commemoration of the inventor of the Earmuff. 1500Z December 21 to 2100Z December 23. 10K up from the General split on 80, 40 and 20M. QSL with 40¢ postage to KA1CNG via Callbook for special 8 × 10 certificate.

A SPECIAL EVENT STATION commemorating the 63rd anniversary of the Tuscaloosa Jaycees on Saturday, January 12. KE4TN will operate from 13 to 23 Zulu and will offer an 8½ × 11 certificate to all contacts. Send QSL card to the Tuscaloosa Jaycees, PO Drawer L, Tuscaloosa, AL 35404 or Callbook address of KE4TN.

MERSEYSIDE SPECIAL EVENT GROUP will be operating during the month of December special event call sign block GBO 1, 2, 4, 6 and 8 BCL. *(Beatle City Liverpool) from 00.00* December 1 to 24.00 December 31. The group will operate all HF bands and 2m and 70cm, all modes. The special event calls are to celebrate the opening of the Beatle City Museum in Liverpool. QSL information direct only to: QSL Manager, G4VKV, c/o Beatle City, PO Box 12, Liverpool, England. Special QSL cards depicting the Beatle City Museum and special event status will be available for all verifiable QSO's and all listener reports.

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P144VDA	144-148	<1.0	15	0	DGFET	$37.95
P144VDG	144-148	<0.5	24	+12	GaAsFET	$79.95
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P220VDA	220-225	<1.2	15	0	DGFET	$37.95
P220VDG	220-225	<0.5	20	+12	GaAsFET	$79.95
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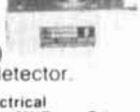

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THE GUERRI REPORT

Ernie Guerri W6MGI

digital signal processing is just around the corner

Applications and techniques for signal processing have become complicated and diverse. An increasing number and variety of integrated circuits whose specialized functions are related to signal processing tasks are now available.

The basic task of signal processing is enhancement of signal-to-noise ratio. The "noise" can be electronic, acoustic, optical, even mathematical. This last aspect has led to modern digital signal processing techniques.

Consider, for example, a television picture: in its analog form, a single frame will have an almost infinite number of shades of gray — every shade from white to black. Some portions of the picture signal may include undesired information such as ignition noise, snow, ghosts, or sync instability. If we were to sample this picture in some manner that would convert the analog signal into a numerical (digital) signal, it would then be possible to assign the undesired information a specific numeric value and address. If we had a way of knowing which pieces of information were intended to be part of the actual picture, then all other pieces could be assumed to be extraneous, and removed. This is the basic technique used in assembling pictures from information sent back from planetary probes, in removing noise and ghosts from network television transmissions, and in compressing the bandwidth needed for transmitting real-time video.

Recently I saw a demonstration in Japan of a TV ghost eliminator made by Toshiba. Multiple images created by tall buildings in an urban environment were completely eliminated using advanced signal processing techniques in the TV set. At the moment, the cost is high — but the technology is here. It won't be too much longer before we have the capability for transmitting acceptable quality real-time video in the space needed for only a few 2-meter FM signals.

higher frequencies

In examining frequency band utilization in decade steps (220 MHz, 2300 MHz, 25 GHz, and 250 GHz) it appears that the higher one goes, the less experimental activity one finds. Yet it's at 250 GHz that the state-of-the-art is actively being explored. The results of this activity are surprising. It takes every trick in the book to get a front-end working at these frequencies, but a team at the National Radio Astronomy Observatory has fabricated a fundamental frequency mixer with a noise figure of less than 9 dB. This device operates at 21 degrees K, using a Schottky diode with a connecting whisker 90 μm (1 μm = 10^{-6} meter) long. The RF bandwidth is over 70 GHz, and IF bandwidth can be up to 10 GHz. Each of these devices can therefore handle an information bandwidth equal to the entire commercial electromagnetic spectrum we now use!

At present, no corresponding devices exist for the generation and amplification of RF energy which would make regular operation at these frequencies possible. But the incentive is there: the spectrum around 260 GHz is a "window" in the atmosphere in which microwave attenuation is only about 5 dB/km.

"faceless" test equipment

We're accustomed to test equipment dedicated to specific measurement tasks. We bring an ailing unit *to* a location — usually a test bench — loaded with meters, generators, analyzers, and such. However, in many major technical companies this traditional approach is being replaced by distributed instrumentation, using microcomputers for display and data processing, with several different "front ends" at hand to accomplish the desired test. One company now offers a dual-channel 50-MHz digital storage "scope" for the Apple computer, and a major instrument supplier has announced a 128-channel waveform digitizer, featuring 0.1 percent accuracy, which multiplexes its data channels to an IBM PC.

This approach should be useful for modular instruments that replace those based on the plug-in concept. Users of very complex instruments such as spectrum analyzers will benefit from the processing power available in this new approach. Although cost will be a limitation at first, we can expect that portable and field test equipment will eventually make use of this approach. Think of the benefit to the service technician with a whole library of diagnostic data available in a single multifunction scopevoltalyzer!

Amateurs have long lacked access to accurate measurements in the frequency domain. The advent of affordable digital signal processing could put a quality spectrum analyzer into the typical Amateur installation within the forseeable future.

ham radio

Food for thought.

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KENWOOD

pacesetter in amateur radio

Pocket-size performers!

TH-21AT/41AT

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Both the 2 meter and 70 cm versions deliver one watt R.F. output on HI power and 150 mW low, for really extended battery life! Functional design includes three digit thumb-wheel switch for easy frequency selection along with a built-in 5 kHz UP-Shift switch and repeater offset switch. (±600 kHz or simplex, 2m version and ±5 MHz or simplex 70 cm version.)

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See your authorized Kenwood dealer and take home a pocket full of 2 m or 70 cm performance today!

Optional accessories:

- HMC-1 headset with VOX
- SMC-30 speaker microphone
- PB-21 Ni-Cd 180 mAH battery
- DC-21 DC power supply
- BT-2 battery case
- EB-2 external C manganese/ alkaline battery case
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- SC-8T soft case for TH-21AT/41AT
- TU-6 programmable sub-tone unit
- AJ-3 thread-loc to BNC female adapter

More information available from authorized dealers of Trio-Kenwood Communications, 1111 West Walnut Street, Compton, CA 90220.

TH-21A/41A

Standard versions.

TH-21AT/41AT Specifications and prices are subject to change without notice or obligation.