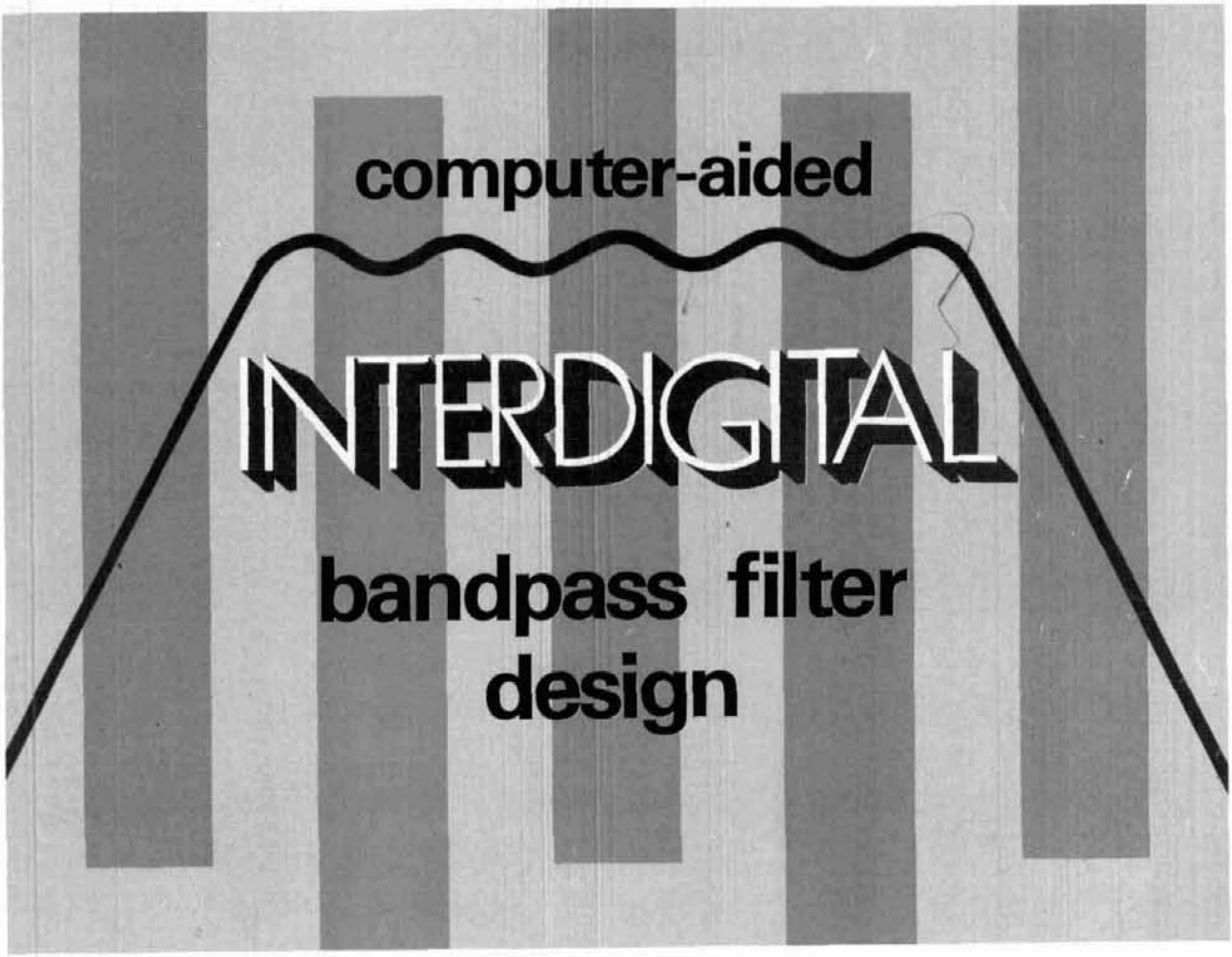
JANUARY 1985 / $2.50

ham radio magazine



simplified gamma matching • EME'ers: find the moon • sensitive field strength meter • IC low-pass filters • full-performance Delta loop • high power RF switching with pin diodes • W1JR on VHF/UHF high power amplifiers • plus W6SAI, KØRYW, and THE GUERRI REPORT

hr

focus on communications technology

ICOM HF Transceiver

IC-745

High Performance Maximum Flexibility

The IC-745 is a full featured, high performance HF base station transceiver with a 100dB dynamic range receiver. PLUS features usually found only in more expensive units.

Compare these exceptional Standard Features:

- 100KHz – 30MHz Receiver
- 100 Watt RF output / 100% Duty Cycle
- Passband Tuning AND IF Shift
- Adjustable Noise Blanker (width and level)
- Adjustable AGC
- Receiver Preamp
- 16 tunable Memories with lithium battery backup

IC-PS30 System Power Supply — IC-SM6 Base Mic

- Wide selection of filters and filter combinations (opt.)
- Continuously adjustable transmit power
- 10Hz/50Hz/1KHz Tuning rates with 1MHz band steps
- IC-HM12 Microphone with Up/Down Scan

Other Standard Features. Included as standard are many of the features most asked for by experienced ham radio operators: dual VFO's, RF speech compressor, tunable notch filter, program band scan, memory scan, all-mode squelch and VOX.

Options. IC-EX310 speech synthesizer, internal IC-PS35 power supply, external IC-PS15 or IC-PS30 system supply. IC-SM8 two-cable desk mic, EX241 marker, EX242 FM module, EX243 electronic keyer, IC-SM6 desk mic, and a variety of filters.

Filter	-6dB Width	Center Freq. MHz
FL45	500 Hz	9.000
FL53A	270 Hz	9.000
FL44A	2.1 KHz	0.455
FL52A	500 Hz	0.455
FL54	250 Hz	0.455

The IC-745 is the only transceiver today that has so much flexibility at a suprisingly low price...see it at your local ICOM dealer.

ICOM

First in Communications

ICOM America, Inc., 2380-116th Ave NE, Bellevue, WA 98004 / 3331 Towerwood Drive, Suite 307, Dallas, TX 75234

All stated specifications are approximate and subject to change without notice or obligation. All ICOM radios significantly exceed FCC regulations limiting spurious emissions. 745I184

What To Look For In A Phone Patch

The best way to decide what patch is right for you is to first decide what a patch should do. A patch should:

- Give complete control to the mobile, allowing full break in operation.
- Not interfere with the normal operation of your base station. It should not require you to connect and disconnect cables (or flip switches!) every time you wish to use your radio as a normal base station.
- Not depend on volume or squelch settings of your radio. It should work the same regardless of what you do with these controls.
- You should be able to hear your base station speaker with the patch installed. Remember, you have a base station because there are mobiles. ONE OF THEM MIGHT NEED HELP.
- The patch should have standard features at no extra cost. These should include programmable toll restrict (dip switches), tone or rotary dialing, programmable patch and activity timers, and front panel indicators of channel and patch status.

ONLY SMART PATCH HAS ALL OF THE ABOVE.

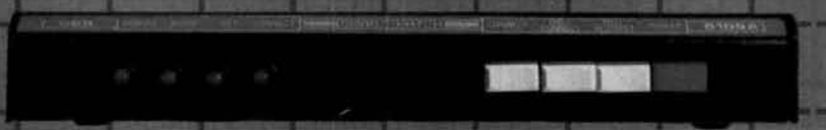
Now Mobile Operators Can Enjoy An Affordable Personal Phone Patch. . .

- Without an expensive repeater.
- Using any FM tranceiver as a base station.
- The secret is a SIMPLEX autopatch, The *SMART PATCH.*

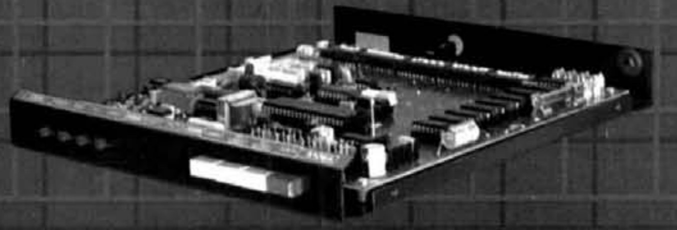
SMART PATCH Is Easy To Install

To install *SMART PATCH*, connect the multicolored computer style ribbon cable to mic audio, receiver discriminator, PTT, and power. A modular phone cord is provided for connection to your phone system. Sound simple? . . . IT IS!

With SMART PATCH You are in CONTROL

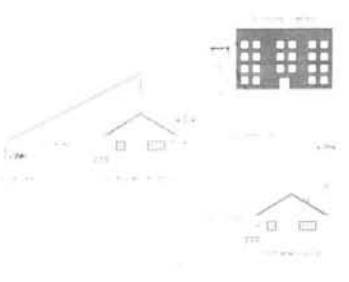


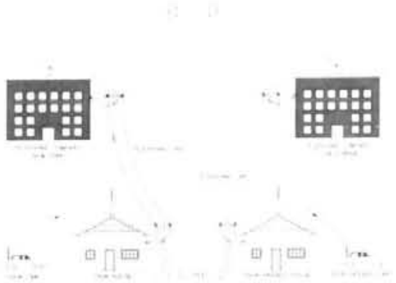
With CES 510SA Simplex Autopatch, there's no waiting for VOX circuits to drop. Simply key your transmitter to take control.



SMART PATCH is all you need to turn your base station into a personal autopatch. SMART PATCH uses the only operating system that gives the mobile complete control. Full break-in capability allows the mobile user to actually interrupt the telephone party. SMART PATCH does not interfere with the normal use of your base station. SMART PATCH works well with any FM transceiver and provides switch selectable tone or rotary dialing, toll restrict, programmable control codes, CW ID and much more.

To Take CONTROL with Smart Patch – Call 800-327-9956 Ext. 101 today.





How To Use *SMART PATCH*

Placing a call is simple. Send your access code from your mobile (example: *73). This brings up the Patch and you will hear dial tone transmitted from your base station. Since *SMART PATCH* is checking about once per second to see if you want to dial, all you have to do is key your transmitter, then dial the phone number. You will now hear the phone ring and someone answer. Since the enhanced control system of *SMART PATCH* is constantly checking to see if you wish to talk, you need to simply key your transmitter and then talk. That's right, you simply key your transmitter to interrupt the phone line. The base station automatically stops transmitting after you key your mic. *SMART PATCH* does not require any special tone equipment to control your base station. It samples very high frequency noise present at your receivers discriminator to determine if a mobile is present. No words or syllables are ever lost.

SMART PATCH Is All You Need To Automatically Patch Your Base Station To Your Phone Line.

Use *SMART PATCH* for:

- Mobile (or remote base) to phone line via Simplex base. (see fig 1.)
- Mobile to Mobile via interconnected base stations for extended range. (see fig. 2.)
- Telephone line to mobile (or remote base).
- *SMART PATCH* uses SIMPLEX BASE STATION EQUIPMENT. Use your ordinary base station. *SMART PATCH* does this without interfering with the normal use of your radio.

WARRANTY?

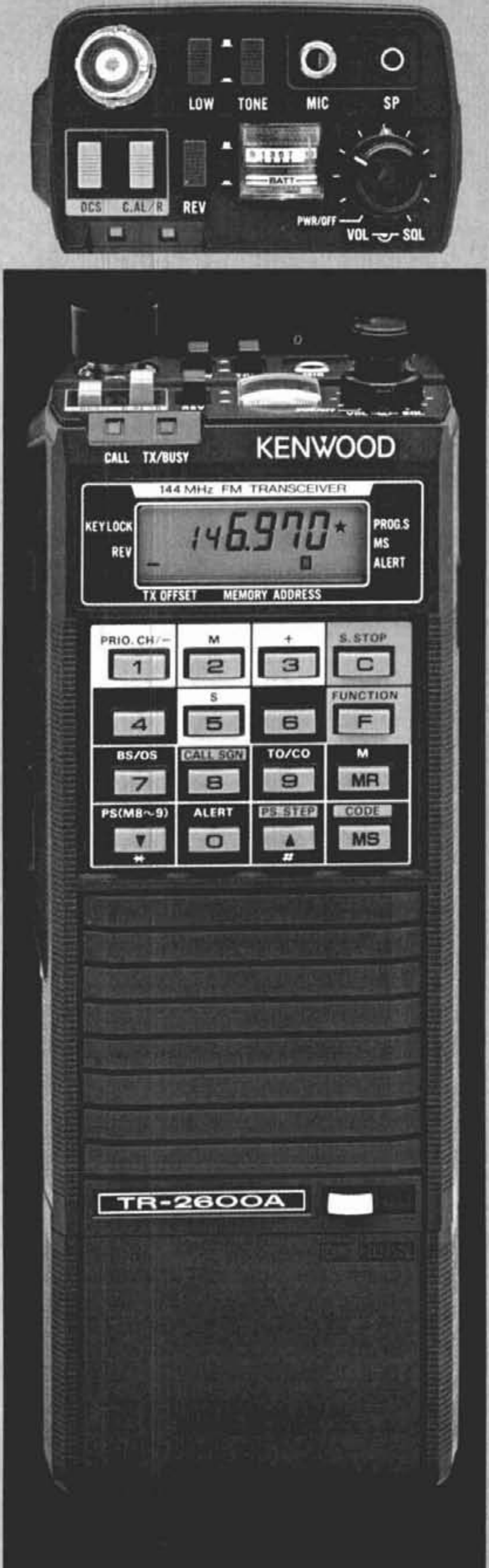
YES, 180 days of warranty protection. You simply can't go wrong.
An FCC type accepted coupler is available for *SMART PATCH.*

Communications Electronics Specialties, Inc.
P.O. Box 2930, Winter Park, Florida 32790
Telephone: (305) 645-0474 **Or call toll-free (800)327-9956**

KENWOOD

... pacesetter in amateur radio

Digital Code Squelch...
TR-2600A



Kenwood's TR-2600A introduces DCS (Digital Code Squelch) circuitry, a signaling concept developed by Kenwood. DCS allows each station to have its own "private call" code or to respond to a "group call" or "common call" code. There are 100,000 different 5-digit ASCII code combinations possible. You can program in call signs up to 6 digits in the ASCII code. When operating in the DCS mode, this information can then be automatically transmitted each time the transmit key is depressed. This revolutionary feature is only the beginning! The TR-2600A also sports a high impact plastic case, that is extra rugged and scuff-resistant. The molded-in color adds to the attractive appearance. The large L.C.D. display is easy to read in direct sunlight or in the dark with a convenient lamp switch. It displays transmit/receive frequencies, memory channels, and five arrow indicators for "F LOCK" frequency lock, "REV" repeater reverse, "PROG.S" programmed scan, "MS" memory scan, "ALERT.S" alert scan. A star indicates "MEMORY LOCK-OUT" is activated, and repeater offset indicated by "+, −, S and M." The TR-2600A has 10 memories, nine for simplex or transmit with frequency offset ±600 kHz and one (memory 0) for non-standard split frequencies. Memory scan and programmable band scan have the added convenience of "Time operated Resume" that stops on busy channel and holds for approximately 5 seconds, then resumes scanning, or "Carrier Operated Resume" that stops on busy channel and resumes when signal ceases.
Memory scan, scans only those memories in which data is stored, and memory lock-out allows you to skip selected memory channels without loss of data previously stored! Manual Scanning UP/DOWN in 5-kHz steps and programmable automatic band scan are also useful features. The TR-2600A has a built-in "S" meter on the top panel which also indicates battery level when in transmit mode. Extended frequency coverage, 142.000-148.995 MHz allows transmit capability in 5-kHz steps for simplex or repeater operation on most MARS and CAP frequencies. Receive frequency coverage includes 140.000-159.995 MHz.

These features only tell part of the story. The TR-2600A also has keyboard frequency selection, built-in 16-key autopatch encoder, "TX STOP" switch, HI (2.5)/LOW (300 mw) power switch, REV switch, "SLIDE-LOC" battery pack, high efficiency speaker, BNC antenna terminal, and all of this in an extremely compact and lightweight package!

Kenwood's TR-2600A, with D.C.S., leads the way in high technology handheld transceivers!

Optional accessories:

- TU-35B built-in programmable sub-tone encoder
- ST-2 Base Stand
- MS-1 Mobile Stand
- PB-26 Ni-Cd Battery
- DC-26 DC-DC Converter
- HMC-1 Headset with VOX
- SMC-30 Speaker Microphone
- LH-3 Deluxe Leather Case
- SC-9 Soft Case
- BT-3 AA Manganese/Alkaline Battery Case
- EB-3 External C Manganese/Alkaline Battery Case
- RA-3, 5. Telescoping Antenna
- CD-10 Call Sign Display

More information on the TR-2600A is available from authorized dealers of Trio-Kenwood Communications, 1111 West Walnut Street, Compton, CA 90220.

Specifications and prices are subject to change without notice or obligation.

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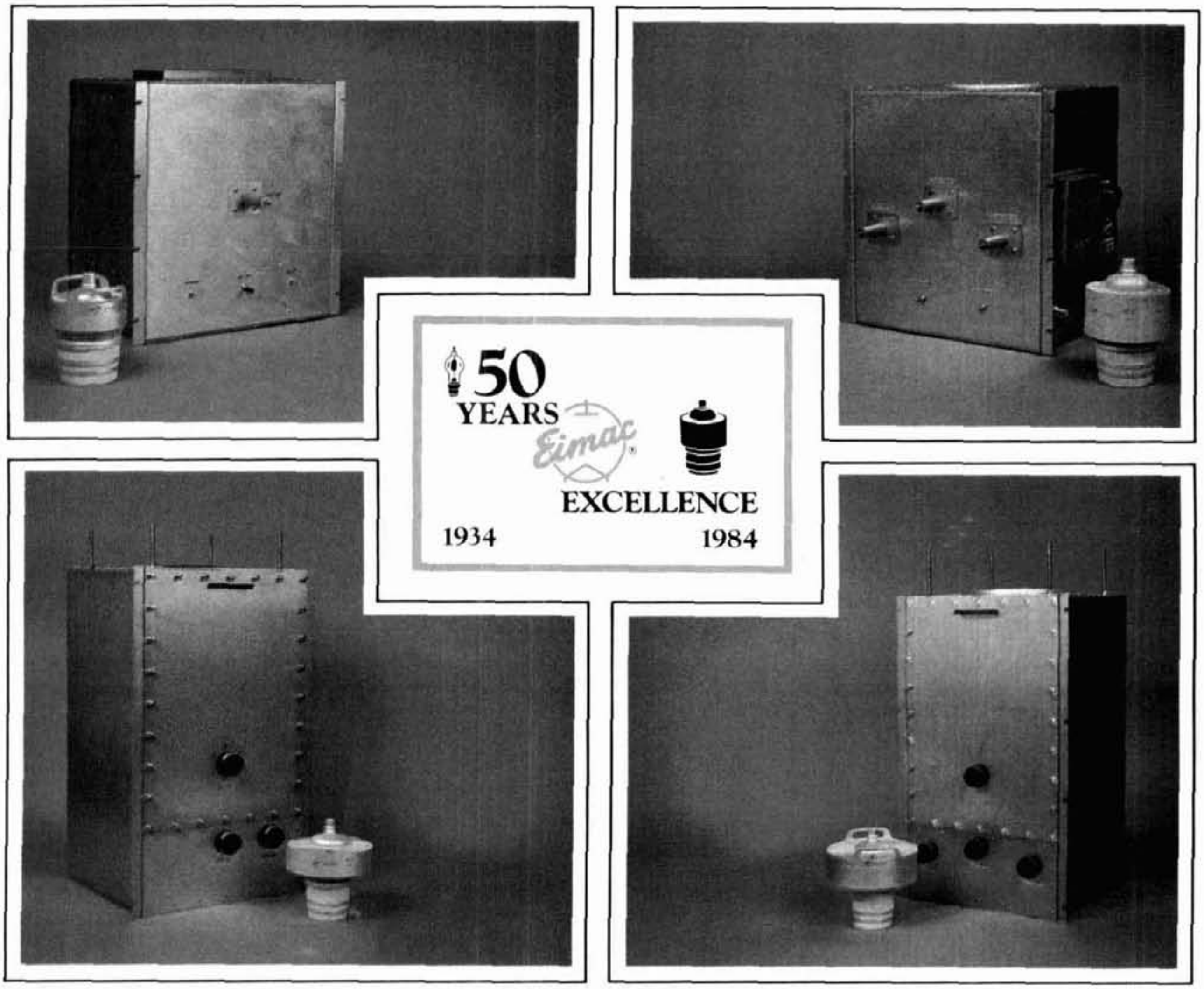

EIMAC celebrates its 50th Anniversary with an extensive line of FM Broadcast Cavity Amplifiers.

Varian EIMAC celebrates 50 years of service to the broadcast industry with a spectrum of FM from a powerful 60 kW to a mini power 150 W solid state IPA.

The cost-effective path to a modern FM transmitter.

No one knows more about broadcast tubes and cavities than EIMAC. Our strong cavity development capability reduces RF engineering problems. EIMAC cavities are inexpensive and simple to use.

For more information call or write Varian EIMAC or contact any Varian Electron Device Group sales office worldwide.

EIMAC FM BROADCAST CAVITY PRODUCT LINE

POWER	CAVITY	EIMAC TUBE
60 kW	CV-2230	4CX30,000G
30 kW	CV-2202	4CX20,000C
25 kW	CV-2200	4CX20,000A
15 kW	CV-2210	4CX12,000A
10 kW	CV-2228	4CX7500A
5 kW	CV-2225	4CX3500A
1.5 kW	CV-2220	3CX1500A7
150 W	AM-2215A	Solid State

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REFLECTIONS

160 meters: spectrum under fire

Once again, Amateur Radio faces a major threat. We recently lost 80 MHz of the 2300-MHz band to aeronautical telemetry. We lost the bottom 10 MHz of 450 MHz within 75 miles of the Canadian border. We lost 25 MHz of the 1200 MHz band to a NAVSAT (navigation satellite) service. We may *yet* lose the 220-MHz band.* Now the 160-meter band is threatened. There's a very real possibility that a proposal to allocate the top half of 160 (1.900 — 2.000 MHz, Amateur exclusive) to the offshore navigation service will be adopted by the FCC.

If you haven't been on 160 during the past few years, you might to surprised by what you would hear now. On any given evening you'll find plenty of stations conducting both CW and SSB ragchews. Later, the DX'ers appear.

Maybe you think 160 is not very important because you never operate there. But consider this: for every *two* stations on 160, there's that much more space available on one of the other bands for you to use. Ten years ago, 160 was considered the AMers' band and there was little activity on it. Not much commercial equipment covered the band. But now, try 160 during any major contest weekend; you'll find plenty of stations on and plenty of DX to be worked.

The FCC's proposal to move the offshore navigation service from 1.6-1.8 MHz to 1.9-2 MHz is predicated on the WARC decision to expand the AM broadcast band up to 1.705 MHz. Industry pundits have suggested that this expansion may end up being more of a boondoggle than a benefit to the consumer and the broadcast industry. There are millions of AM radios around — few cover the entire 1.6-1.750 MHz band!

Some serious questions need to be answered. Can the broadcast industry and the general business community financially support additional AM stations? Competition is tough enough now. Will other users of the spectrum be adversely affected in any way? It's been said that the proposed expansion is the result of some political debt being repaid. Whatever the motive, the proposal seems to be a foolhardy and unnecessary exercise.

Now what about the offshore navigation interests? Serious questions have been raised about this service. Why is it necessary to have an MF radiolocation service when there are other, more precise methods of radiolocation such as LORAN-C and NAVSAT? How about using technology similar to that of the new microwave landing system currently being integrated into airports by the FAA? The bottom line is that offshore navigation has few, if any, credible reasons to be moved into the 1.9-2 MHz slot. If the broadcast band is to be expanded only as far as 1.705 MHz, can anyone explain why, with selective receivers and stable transmitters, the offshore navigation interests cannot be accommodated between 1.705 and 1.8 MHz? It's almost the same size as the allocation proposed, and offshore navigation already operates there. Furthermore, if offshore navigation interests are allowed to get their way, what's to stop other services from claiming Amateur frequencies on the basis of equally flimsy justification? The ultimate result would surely be a major disruption in Amateur Radio as we know it. Don't kid yourself. This could be just the beginning!

If you haven't already filed your comments with the FCC on the proposed 160-meter reallocation, now is the time to do so. Your comments must be received by the FCC before January 24, 1985. Be reasonable and concise. Give solid technical and operational reasons as to why this proposal should not be accepted. Write "Docket 84-874" at the top of each page of your comments, include five additional copies (eleven if you wish each commissioner to have one), and send them to THE SECRETARY, FCC, Washington, D. C. 20554.

It is a unique privilege that we have to be able to be included in the decision-making process proposals such as this. If you don't sit down right now, write up your comments, and send them in, then when the FCC accepts the offshore navigation proposal because of a lack of adequate opposition from the Amateur community, all that can be said is "Quityergripen!"

J. Craig Clark, N1ACH
Assistant Publisher

*See "Reflections," *ham radio*, October, 1984 (page 6) and this month's *Presstop* (page 8).

THE MOST AFFORDABLE REPEATER

ALSO HAS THE MOST IMPRESSIVE PERFORMANCE FEATURES

(AND GIVES THEM TO YOU AS STANDARD EQUIPMENT!)

JUST LOOK AT THESE PRICES!

Band	Kit	Wired/Tested
10M,6M,2M,220	$680	$880
440	$780	$980

Both kit and wired units are complete with all parts, modules, hardware, and crystals.

CALL OR WRITE FOR COMPLETE DETAILS.

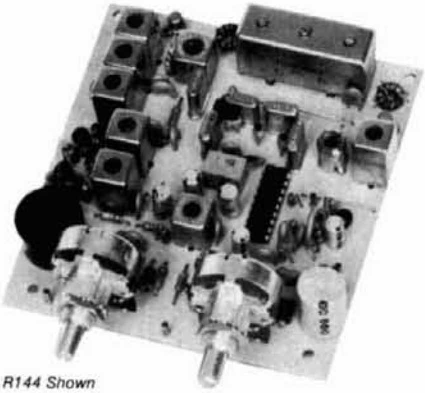
Also available for remote site linking, crossband, and remote base.

FEATURES:

- SENSITIVITY SECOND TO NONE; TYPICALLY 0.15 uV ON VHF, 0.3 uV ON UHF.
- SELECTIVITY THAT CAN'T BE BEAT! BOTH 8 POLE CRYSTAL FILTER & CERAMIC FILTER FOR GREATER THAN 100 dB AT ± 12KHZ. HELICAL RESONATOR FRONT ENDS. SEE R144, R220, AND R451 SPECS IN RECEIVER AD BELOW.
- OTHER GREAT RECEIVER FEATURES: FLUTTER-PROOF SQUELCH, AFC TO COMPENSATE FOR OFF-FREQ TRANSMITTERS, SEPARATE LOCAL SPEAKER AMPLIFIER & CONTROL.
- CLEAN, EASY TUNE TRANSMITTER; UP TO 20 WATTS OUT (UP TO 50W WITH OPTIONAL PA).

HIGH QUALITY MODULES FOR REPEATERS, LINKS, TELEMETRY, ETC.

HIGH-PERFORMANCE RECEIVER MODULES



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.
- **R76 FM RCVR** for 10M, 6M, 2M, 220, or commercial bands. As above, but w/o AFC or hel. res. Kits only $118. Also avail w/4 pole filter, only $98/kit.
- **R110 VHF AM RECEIVER** kit for VHF aircraft band or ham bands. Only $98.
- **R110-259 SPACE SHUTTLE RECEIVER,** kit only $98.

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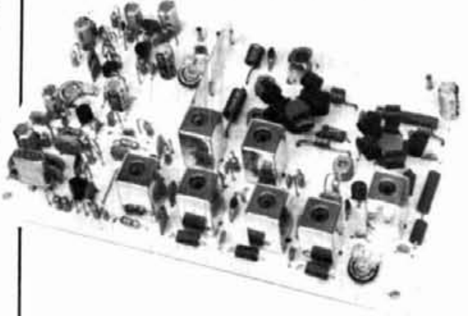
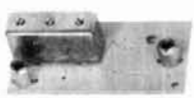
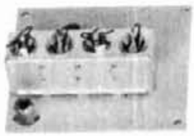
TRANSMITTERS



- **T51 VHF FM EXCITER** for 10M, 6M, 2M, 220 MHz or adjacent bands. 2 Watts continuous, up to 2½ W intermittent. $68/kit.

- **T451 UHF FM EXCITER** 2 to 3 Watts on 450 ham band or adjacent freq. Kit only $78.
- **VHF & UHF LINEAR AMPLIFIERS.** Use on either FM or SSB. Power levels from 10 to 45 Watts to go with exciters & xmtg converters. Several models. Kits from $78.
- **A16 RF TIGHT BOX** Deep drawn alum. case with tight cover and no seams. 7 x 8 x 2 inches. Designed especially for repeaters. $20.

ACCESSORIES

- **HELICAL RESONATOR FILTERS** available separately on pcb w/connectors.
  HRF-144 for 143-150 MHz $38
  HRF-220 for 213-233 MHz $38
  HRF-432 for 420-450 MHz $48
- **COR-2 KIT** With audio mixer, local speaker amplifier, tail & time-out timers. Only $38.
- **COR-3 KIT** as above, but with "courtesy beep". Only $58.
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- **DTMF DECODER/CONTROLLER KITS.** Control 2 separate on/off functions with touchtones®, e.g., repeater and autopatch. Use with main or aux. receiver or with Autopatch. Only $90
- **AUTOPATCH KITS.** Provide repeater autopatch, reverse patch, phone line remote control of repeater, secondary control via repeater receiver. Many other features. Only $90. Requires DTMF Module.
- **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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*

No Need to Pay $80 to $125 for a GaAs FET Preamp.

FEATURES:

- Very Low Noise: 0.7 dB VHF, 0.8 dB UHF
- High Gain: 18 to 28 dB, Depending on Freq.
- Wide Dynamic Range for Overload Resistance
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- Rugged, Diode-protected Transistors
- Easy to Tune
- Operates on Standard 12 to 14 Vdc Supply
- Can be Tower Mounted

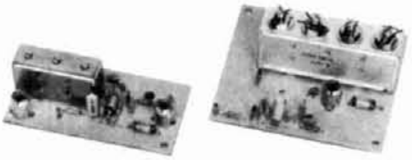
MODEL	TUNES RANGE	PRICE
LNG-28	26-30 MHz	$49
LNG-50	46-56 MHz	$49
LNG-144	137-150 MHz	$49
LNG-220	210-230 MHz	$49
LNG-432	400-470 MHz	$49
LNG-40	30-46 MHz	$64
LNG-160	150-172 MHz	$64

ECONOMY PREAMPS

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.

- P30K, VHF Kit less case $18
- P30W, VHF Wired/Tested $33
- P432K, UHF Kit less case $21
- P432W, UHF Wired/Tested $36

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
	50-52	28-30
Kit with Case $49	50-54	144-148
Less Case $39	144-146	28-30
Wired $69	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
	435-437	28-30
Kit with Case $59	432-436	144-148
Less Case $49	432-436	50-54
Wired $75	439.25	61.25

SCANNER CONVERTERS Copy 72-76, 135-144, 240-270, 400-420, or 806-894 MHz bands on any scanner. Wired/tested Only $88.

SAVE A BUNDLE ON VHF FM TRANSCEIVERS!

FM-5 PC Board Kit – **ONLY $178**
complete with controls, heatsink, etc.
10 Watts, 5 Channels, for 2M or 220 MHz.

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

REPEAT OF A SELLOUT!

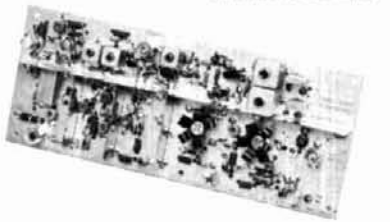
While supply lasts, get $60 cabinet kit free when you buy an FM-5 Transceiver kit. Where else can you get a complete transceiver for only $178

TRANSMIT CONVERTERS

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.

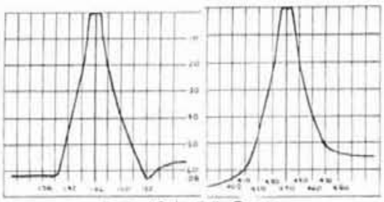
	Exciter Input Range	Antenna Output
For VHF,	28-30	144-146
Model XV2	28-29	145-146
Kit $79	28-30	50-52
Wired $149	27-27.4	144-144.4
(Specify band)	28-30	220-222*
	50-54	220-224
	144-146	50-52
	50-54	144-148
	144-146	28-30
For UHF,	28-30	432-434
Model XV4	28-30	435-437
Kit $99	50-54	432-436
Wired $169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

LOOK AT THESE ATTRACTIVE CURVES!



Typical Selectivity Curves of Receivers and Helical Resonators.

IMPORTANT REASONS WHY YOU SHOULD BUY FROM THE VALUE LEADER:

1. *Largest selection of vhf and uhf kits in the world.*
2. *Exceptional quality and low prices due to large volume.*
3. *Fast delivery; most kits shipped same day.*
4. *Complete, professional instruction manuals.*
5. *Prompt factory service available and free phone consultation.*
6. *In business 21 years.*
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COMMERCIAL PRESSURE ON THE 220-MHZ BAND HAS BEEN RELIEVED by the FCC's assignment of the Land Mobile reserve frequencies between 806 and 947 MHz to that and other services November 21. In its far-reaching action, the Commission rejected both Mura's proposal for a new Personal Radio Service for 8 MHz and a commercial proposal to allocate 896-898 and 941-943 MHz to an air-ground telephone service; GE's proposal for a Personal Radio Service in the 900 MHz region was dropped just a few weeks earlier.

12 MHz For Land Mobile And 12 MHz For Cellular Were Proposed in the November 21 meeting, while 6 MHz was reallocated to government and non-government fixed services and frequencies in the 944-947 MHz band were reallocated for broadcast links and relays. As the "220 Grab" was already in trouble from a variety of directions, this action is likely to effectively remove 220 from Land Mobile's sights. However, it's also just as likely to create a whole new set of user pressures on 220 MHz as well as other Amateur bands.

The Failure To Establish Any Form Of A "Personal" Radio Service in the Land Mobile reserve frequencies is regarded as a severe defeat by would-be users and suppliers alike. The concept had been strongly supported by the GMRS* community. The next logical targets for a Personal Radio Service could now be 902-928 MHz, and (repeating history) 220 MHz! The air-to-ground telephone proponents could, on the other hand, decide to go after 420-430 MHz for their new service since that band has already gone to Land Mobile in Canada and is being protected for Canadian benefit along the Canadian border.

Whether Or Not These Fears Materialize, pressure on Amateur frequencies is expected to continue. The recent loss of the Amateur secondary allocation of 2310-2390 MHz to flight telemetry is a case in point. The current stagnation of Amateur growth, coupled with apparent FCC coolness toward the Amateur Service, could spell trouble ahead.

TWO LONG-TERM ARRL DIRECTORS WERE UNSEATED in the League elections. In the New England Division Tom Frenaye, K1KI, soundly beat John Sullivan, W1HHR. In a closer race Linda Ferdinand, N2YL, beat out incumbent George Diehl, W2IHA, in the Hudson Division. Incumbents won in all other races in which they ran; however two new Vice Directors, Rush Drake, W7RM, (Northwestern Division) and Wayne Overbeck, N6NB, (Southwestern Division) were both also elected.

EXTENSIVE AMATEUR OPERATION DURING A 1985 SPACE SHUTTLE mission now seems certain, according to word from NASA. It's to be on mission 51-F, now scheduled for next April, though delays in the Shuttle schedule could push that back to July or even later. Two Amateurs are scheduled for mission 51-F, and it appears likely there'll be operation on a variety of modes and frequencies with some sort of repeater also possible.

ARRL'S LEAVING NEWINGTON AND A MAJOR PUSH FOR AMATEUR GROWTH were just two of the many significant items covered by the League's directors at their fall meeting. As for moving, the Management and Finance Committee was directed to conduct a study of possible sites, costs, timing, and impact on the membership and staff. Though not specified in the board minutes, Washington, D.C. is believed to be what's in mind. A growth of 50,000 new Amateurs a year for the next five years is the goal of a new plan devised by the League's General Manager, for implementation in 1985. (The number of individual U.S. licensed Amateurs has held almost steady at just over 400,000 for the past several years.) He's also been instructed to develop a parallel program to increase ARRL membership by 25,000 in 1985, and by 20% per year thereafter. Other items included an apparent endorsement of simplex autopatches, a membership survey of further phone expansion on 7 MHz, and a Plans and Programs Committee study on the League becoming involved with maintenance of Amateur licensing records and "especially in the administration of special call sign requests...."

ARRL'S REQUEST FOR FCC PREEMPTION OF ANTENNA REGULATIONS, PRB-1, has been drawing a lot more favorable—and effective—Amateur comment. November 9 the FCC extended the comment cutoff date to December 24, and indications are that the flow of comments from Amateurs describing the problems they've had with local restrictions is continuing at a good rate.

What The Outcome Will Be, However, Remains To Be Seen. There have also been some well presented arguments from the other side of the fence, and the issue is one which does have two valid sides. Which side, if either, the Commissioners will take cannot be predicted.

FCC'S PROPOSAL TO REALLOCATE 1900-2000 KHZ to radio direction finding is still open for comment. Just after last month's Presstop went to press, the Commission granted an ARRL request for an extension of the Comment due date to January 24, 1985, with Reply Comments due March 11. An original and five copies sent to the Secretary, FCC, are needed for a formal filing; refer to PR Docket 84-874. (See "Reflections," page 5.)

440-450 MHZ IS FULL IN SOUTHERN CALIFORNIA, according to the Southern California Repeater and Remote Base Association (SCRRBA), with all repeater channels filled. Unless SCRRBA's review of activity turns up some inactive systems, would-be repeater operators will have to go to 23 cm since 2 meters and 220 have long been filled in that area.

*General Mobile Radio Service. Formerly "Class A" CB, GMRS has developed into a sophisticated Personal FM Radio Service with frequencies in the 460-MHz band.

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computer-aided interdigital bandpass filter design

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One of the more challenging difficulties facing the Amateur experimenter in UHF and microwave communications is to construct good bandpass filters at these high frequencies. Good filters are especially important in those bands where the Amateur frequency allocation is shared with or is near high power services, such as radar, which can cause ruinous interference. And although a number of articles have appeared in Amateur publications describing specific bandpass filter types and specific construction techniques, none has provided a simple means of designing filters for different requirements.

This article describes a flexible computer program that makes it possible for Amateurs to design and build their own bandpass filters, custom-tailored to their specific needs. The program performs the design tasks for interdigital filters, a common bandpass type. The article also shows, step by step, how to proceed from the computer printout to the building and testing of a working filter. In addition, two different examples are used to illustrate two different construction methods.

filter design

The design of narrowband bandpass filters — *that* is, of filters that have passbands of approximately 10 percent of the center frequency or less — is often based on the well-known work of Matthaei, Young, and Jones,[1] which provides a wealth of analytical and practical design methods for many types of RF and microwave filters. Among the most common types are *filters that use comb or interdigitated coupled resonators.* These two types are mechanically similar, but this article discusses only the interdigital form.

Interdigitated bandpass filters have been widely used in the microwave electronics industry for many years. These filters are commonly used because they provide reasonably good passband characteristics, moderate loss, and fairly high attenuation in the stop-bands. Furthermore, they are simple to build and tune.

A typical interdigitated bandpass filter, such as shown in **fig. 1**, consists of a number of resonator elements or rods, each approximately a quarter wavelength long at the center frequency of the filter, which are electrically coupled together between two conducting ground planes. Each rod is shorted to ground at one end and open-circuited at the other end. The rods *alternate, with one rod's shorted end opposing the* next rod's open end. It is this alternating structure, which looks somewhat like interlaced fingers, that gives the interdigital filter its name. At the ends of the filter some form of impedance matching, either a transmission line transformer or a tap on the end rods, is used to couple energy into and out of the filter.

Interdigitated filters are most useful in the low microwave frequency range of about 0.5 to 5 GHz. In this region, lumped element filters are difficult to build, and waveguide filters are mechanically large. The inter-

By Jerry Hinshaw, N6JH, and Shahrokh Monemzadeh, 4558 Margery Drive, Fremont, California 94538

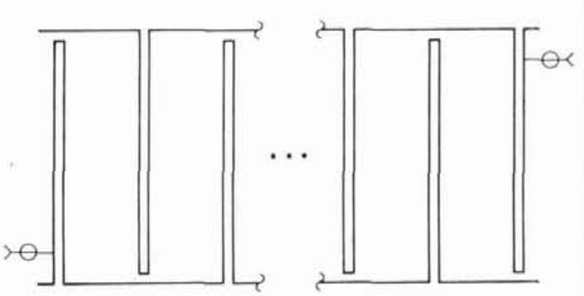

fig. 1. Typical interdigital bandpass filter structure. This filter consists of a number of rods, each open circuited at one end and grounded at the other, which are electrically coupled together between ground planes.

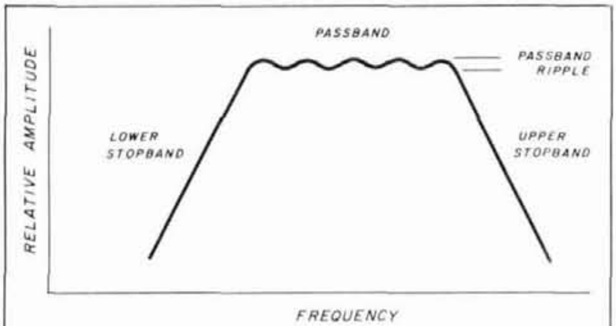

fig. 2. A generalized bandpass filter response. The ripple in the passband is exaggerated for clarity.

digital filter handily fills this gap between low frequency "coil and capacitor" filters, and microwave waveguide "plumbing." Thus, interdigital filters are of interest for frequencies up to at least several GHz, and down at least as low as the 420 MHz ham band.

The traditional design of interdigitated filters described by Matthaei, Young, and Jones calls for both the spacing and sizes of the rectangular resonant elements to be variables. While there is no problem with this theoretically, in practice it is often simpler to use round rods of equal diameter in place of rectangular ones of various sizes. In the 1960's, Dishal[2] described a method of designing narrow bandwidth filters using equal diameter round rods. His method provides a simple and accurate design guide which results in bandpass filters of straightforward mechanical construction and good electrical performance.

In addition to using uniform round rods in place of rectangular elements of various sizes, Dishal questioned the common practice of using additional elements, one at each end of the filter structure, whose only purpose was to match the filter to the desired input and output connections when a simple tap on the first and last element would serve as well. Taking it

defining terms

A *bandpass filter* is a device that permits the passage of signals within a certain range of frequencies only, but blocks out signals from above or below that range. The range of frequencies where signals can pass through the filter with low loss is called the *passband*. The frequencies outside the passband are variously referred to as the *rejection bands*, *stop bands* or *"skirts"* of the filter. (**Figure 2** shows a general filter response.)

An ideal bandpass filter would have zero loss in the passband and infinite rejection of undesired signals in the stopbands. In real life, however, the situation is not as clear-cut: the passband of a real filter has a certain finite loss, even though it may be quite low. Likewise, a real filter's stopbands do not absolutely reject signals, but merely reduce their amplitudes. The attenuation depends mainly upon the separation, in frequency, of the undesired signal from the passband.

The loss in the passband may be nearly constant, or flat, as in a Butterworth design, or it may vary with regular undulations or "ripple" in the attenuation curve. In a Chebyshev type filter, this passband ripple is related to an increase in stopband rejection. For our purposes, it is enough to consider a Butterworth filter as a variation of the Chebyshev design, but with zero passband ripple. It is important to realize that we can obtain better stopband rejection if we're willing to trade off passband flatness and accept higher ripple and VSWR.

Since the transition from passband to stopband in a real (rather than ideal) filter is not abrupt, we must define its location. For a Chebyshev filter, the "edges" of the passband are defined as those points where the attenuation exceeds the maximum ripple amplitude. Thus the passband of a Chebyshev filter is often known as the *ripple bandwidth*. The bandwidth of a Butterworth design, which has no ripple, is traditionally defined at the 3 dB points. This measure is called the *3 dB bandwidth*.

There are many good texts and articles dealing with filter topics, and the interested reader should refer to them if a better background is desired. A few of these sources are listed in the references. However, the terms and concepts introduced above are sufficient for a basic understanding of the design and construction of the bandpass filters we describe.

table 1. Interactive program calculates expected electrical performance and computes mechanical dimensions of filter.

```
10 REM *********************************************
20 REM   DESIGNS INTERDIGITAL BPFs
30 REM *********************************************
40  DEF FNRJ(TA,B,C,D)=(B*C-TA*D)/(C*C+D*D)
50 REM using equal diameter rods
60 REM g values based on ripple bw . q/coup on 3-bd
70 DIM G(200) , C(200) , RK(200) , AK(200) , FR(40) , ALOSS(40)
80  DIM A(200), B(200)
90 PI=3.14159265#
100 INPUT"# OF ELEMENT $ P-P RIPPLE IN PASSBAND (DB)";N,RIP
110 REM
120 INPUT"INPUT FILTER CENTER FREQ.(GHZ),BW(MHZ)&LOAD IMPEDENCE
     ZO";FZGC,BWMC,R
130 REM
140 PRINT"INPUT GROUND PLANE SPACING , ROD DIAMETER"
150 INPUT"& DISTANCE TO CENTER OF FIRST AND LAST ROD";H,D,E
160 REM
170 REM
180 INPUT"NO. OF FREQ. REJECTION PTS AND STEP SIZE (MHZ)";NFR,STP
190 FOR IP=-NFR/2 TO NFR/2
200      CONTER=CONTER+1
210      FR(CONTER)=FZGC+(STP*.001*IP)
220 NEXT IP
230 IDAT=1
240 GOTO 250
250 F1=FZGC-.0005*BWMC
260 F2=FZGC+.0005*BWMC
270 IF RIP>0 THEN GOTO 330
280 BW3GC=F2-F1
290 BWRGC=0
300 BW3=1
310 GOSUB 1960
320 GOTO 390
330 B=1/SQR(10^(.1*RIP)-1)
340 CA=LOG(B+SQR(B*B-1))/(N)
350 BW3=(EXP(CA)+EXP(-CA))/2
360 GOSUB 1740
370 BWRGC=F2-F1
380 BW3GC=BWRGC*BW3
390 REM
400 W=2*(F2-F1)/(F2+F1)
410 QF=FZGC/BW3GC
420 NFM=N-1
430 QWVL=11.8028/(4*FZGC)
440 FOR K=1 TO NFM
450      AK(K)=1/(BW3*SQR(G(K)*G(K+1)))
460      RK(K)=AK(K)/QF
470 NEXT K
480 AKO=G(1)*BW3
490 AK(N)=AKO
500 AK(N+1)=0
510 QS=G(1)*BW3*QF
520 CANH=(EXP(2*PI*E/H)-1)/(EXP(2*PI*E/H)+1)
530 ZM=59.9585*LOG(4*H/(PI*D))
540 ZE=59.9585*LOG(CANH*H*4/(PI*D))
550 RKM=RK(1)*SQR(ZM/ZE)
560 Z=PI*D/(2*H)
570 COTH=(EXP(Z)+1)/(EXP(Z)-1)
580 Y=PI*RKM/4
590 T=COTH^Y
600 C(1)=(H/PI)*LOG((T+1)/(T-1))
610 MFL=N-2
620 REM IF N-3<0 THEN AG=1 ELSE IF N-3=0 THEN AG=2 ELSE AG=3
630 ON (2+SGN(N)*1) GOTO 690 , 690 , 640
640 FOR K=2 TO MFL
650      Y=PI*RK(K)/4
660      T=COTH^Y
670      C(K)=(H/PI)*LOG((T+1)/(T-1))
680 NEXT K
690 C(N-1)=C(1)
700 X=SQR(PI*R/(4*ZE*QS))
710 AQ=2*QWVL*ATN(X/SQR(1-X*X))/PI
720 QU=2200*H*SQR(FZGC)
730 SUMG=0
740 FOR J1=1 TO N
750      SUMG=SUMG+G(J1)
760 NEXT J1
770 BLOSS=4.34*FZGC*SUMG/(QU*(F2-F1))
780 DELAY=SUMG/(2*PI*(F2-F1))
790 IF RIP > 0 THEN GOTO 820
800 PRINT"DESIGN DATA FOR ";N;" POLE INTERDIGITAL FILTER.
     BUTTERWORTH RESPONSE"
810 GOTO 830
820 PRINT "DESIGN DATA FOR";N;"POLE INTERDIGITAL FILTER .BAND
     PASS RIPPLE";RIP;"DB"
830 PRINT"CENTER FREQ. ";FZGC ;"GHZ"
840 PRINT"CUTOFF FREQ. ";F1;" (GHZ) AND ";F2;"  GHZ"
850 PRINT"RIPPLE BW.  ";BWRGC;"GHZ"
860 PRINT"3 DB BW.    ";BW3GC;"GHZ"
870 PRINT"FRACTIONAL BW.";W
880 PRINT"FILTER Q  ";QF
890 PRINT"EST QU     ";QU
900 PRINT"LOSS BASED ON THIS QU  ";BLOSS;"  DB"
910 PRINT"DELAY AT BAND CENTER ";DELAY;"NANOSECONDS"
920 FOR JK=1 TO NFR
930      IF JK=1 THEN PRINT "FREQUENCY REJECTION INFORMATION "
940      NFN=ABS(2*(FR(JK)-FZGC)/(W*FZGC))
950      IF RIP >0 THEN GOTO 980
960      ALOSS(JK)=10*LOG(1+NFN^(2*N))/LOG(10)
970      GOTO 1040
980      IF NFN<1 THEN NFN=1
990      ANG=N*LOG(NFN+SQR(NFN*NFN-1))
1000      YAK=.5*(EXP(ANG)+EXP(-ANG))
1010      ALOSS(JK)=10*LOG(1+(10^(.1*RIP)-1)*YAK*YAK)/LOG(10)
1020      IF ALOSS(JK)>65 THEN ALOSS = 65 ELSE ALOSS = ALOSS(JK)
1030      FR=INT(FR(JK)*10000)/10000 : ALOS=INT(ALOSS(JK))
1040      PRINT TAB(INT(ALOSS))"*";TAB(66)FR;TAB(73)ALOS
1050 NEXT JK
1060 WO=2*PI*FZGC*1E+09
1070 F=D/H
1080  CF=(-.0000422+.0857397*F+.0067853*F*F-9.092165E-
     02*F^3+.169088*F^4)*PI*H*2.54
1090 REM
1100 WW=WO*1E-12
1110 B2=PI*AQ/(2*QWVL)
1120 GG=1/R
1130 BB=-COS(B2)/(ZE*SIN(B2))
1140 EL1=.8*QWVL
1150 ANG=EL1*PI/(2*QWVL)
1160 B1=ANG-B2
1170 YL=-COS(ANG)/(ZM*SIN(ANG))
1180 CP=WW*(CF+.17655*D*D/(QWVL-EL1))
1190 Y1=CP+YL
1200 EL2=.87*QWVL
1210 ANG=EL2*PI/(2*QWVL)
1220 B4=ANG-B2
1230 YL=-COS(ANG)/(ZM*SIN(ANG))
1240 CD=WW*(CF+.17655*D*D/(QWVL-EL2))
1250 Y2=CD+YL
1260 EL3=.95*QWVL
1270 ANG=EL3*PI/(2*QWVL)
1280 B5=ANG-B2
1290 YL=-COS(ANG)/(ZM*SIN(ANG))
1300 CQ=WW*(CF+.17655*D*D/(QWVL-EL3))
1310 Y3=CQ+YL
1320  ELEM=Y3*Y2*EL1/((Y1-Y2)*(Y1-Y3))+Y1*Y3*EL2/((Y2-Y1)*(Y2-
     Y3))+Y1*Y2*EL3/((Y3-Y1)*(Y3-Y2))
1330 TANN=SIN(B1)/COS(B1)
1340 YL=FNRJ(GG,BB+TANN/ZE,1-ZE*BB*TANN,ZE*GG*TANN)
1350 Y1=CP+YL
1360 TANN=SIN(B4)/COS(B4)
1370 YL=FNRJ(GG,BB+TANN/ZE,1-ZE*BB*TANN,ZE*GG*TANN)
1380 Y2=CD+YL
1390 TANN=SIN(B5)/COS(B5)
1400 YL=FNRJ(GG,BB+TANN/ZE,1-ZE*BB*TANN,ZE*GG*TANN)
1410 Y3=CQ+YL
1420  ELEQ=Y3*Y2*EL1/((Y1-Y2)*(Y1-Y3))+Y1*Y3*EL2/((Y2-Y1)*(Y2-
     Y3))+Y1*Y2*EL3/((Y3-Y1)*(Y3-Y2))
1430 REM
1440 PRINT"QUARTER WAVELENGTH =";QWVL ;"INCHES"
1450 PRINT"THE LENGTH OF INTERIOR ELEMENTS =";ELEM;" INCHES"
1460 PRINT"LENGTH OF END ELEMENTS =";ELEQ;" INCHES"
1470 PRINT"GROUND-PLANE SPACE =";H," INCHES "
1480 PRINT"ROD DIAMETER   =";D;" INCHES"
1490 PRINT"END PLATES";E;" INCHES FROM C/L OF END ROD "
1500 PRINT"TAP EXTERNAL LINES UP ";AQ;" INCHES FROM SHORTED END "
1510 PRINT"LINE IMPEDANCES: END ROD";ZE;" ,OTHER ";ZM;" , EXT.
     LINES ";R;"OHM"
1520 PRINT"DIMENSIONS"
1530 PRINT"EL. NO.      END TO C      C TO C     G(K)        Q/COUP"
1540 DOM=E
1550 GOO=1
1560 PRINT "0";TAB(41)GOO;TAB(55)AKO
1570 PRINT "1";TAB(16)E;TAB(41)G(1);TAB(55)AK(1)
1580 FOR K=1 TO NFM
1590      L=K+1
1600      PRINT TAB(28)C(K)
1610      DOM=DOM+C(K)
1620      PRINT L;TAB(16)DOM;TAB(41)G(L);TAB(55)AK(L)
1630 NEXT K
1640 LQ=N+1
1650 PRINT LQ;TAB(41)G(LQ)
1660 DOM=DOM+E
1670 PRINT TAB(16)DOM
1680 IF IDAT =1 THEN GOTO 2070
1690 REM
1700 REM
1710 REM DEFINE FUNCTION
1720 DEF FNRJ(TA,B,C,D)=(B*C-TA*D)/(C*C+D*D)
1730 END
1740 REM SUB CHEB
1750 REM
1760 C=2*RIP/17.37
1770 BETA=LOG((EXP(C)+1)/(EXP(C)-1))
1780 GAMMA=.5*(EXP(BETA/(2*N))-EXP(-BETA/(2*N)))
1790 FOR K=1 TO N
1800      A(K)=SIN(.5*(2*K-1)*PI/N)
1810      B(K)=GAMMA^2+SIN(K*PI/N)^2
1820 NEXT K
1830 G(1)=2*A(1)/GAMMA
1840 FOR K=2 TO N
1850      G(K)=4*A(K-1)*A(K)/(B(K-1)*G(K-1))
1860 NEXT K
1870 NN=N/2
1880 NNN=(N+1)/2
1890 REM IF NNN-NN<0 THEN AG1=1 ELSE IF NNN-NN=0 THEN AG1=2 ELSE AG1=3
1900 ON (2+SGN(NNN-NN)*1) GOTO 1910,1910,1930
1910 G(N+1)=((EXP(BETA/2)+1)/(EXP(BETA/2)-1))^2
1920 RETURN
1930 G(N+1)=1
1940 RETURN
1950 END
1960 REM SUB FOR BUTT
1970 REM
1980 REM
1990 REM
2000 REM
2010 POV2=1.57079633#
2020 FOR K=1 TO N
2030      G(K)=2*SIN(POV2*(2*K-1)/N)
2040 NEXT K
2050 G(N+1)=1
2060 RETURN
2070 END
```

number of elements	ripple dB	bandwidth MHz	loss at 440 dB	loss at 445 dB
2	0.25	2	1.2	21.6
2	0.25	3	0.8	14.4
3	0.25	2	2.5	41.4
3	0.25	3	1.7	30.5
3	0.25	4	1.3	22.6
4	0.50	1	7.9	88.8
4	0.50	3	2.7	50.0
4	0.25	3	2.4	46.8
6	0.25	3	4.2	79.4
6	0.25	4	3.2	63.4

fig. 3. Program tests various configurations for the 440-MHz bandpass filter. For comparison, all are for 0.38 inch diameter rods between groundplanes 1 inch apart, and all have a center frequency of 440 MHz.

a step further, he showed how to determine these tap locations. A computer program, described below, is developed around this design approach.

program description

The BASIC listing is given in **table 1**. This program follows in loose form a program originally written in Fortran IV by Rook and Taylor.[3] We translated into BASIC, modified it for use on a personal computer, and added additional plotting output. The program uses an interactive approach to request design information from the user, and then calculates the expected electrical performance and computes the mechanical dimensions of the filter. It is written in BASIC in its IBM PC™ version, but it is structured so that conversion into other versions of BASIC for different computers should not be difficult.

The first portion of the program sets up the required variable dimensions and types. Next, an interactive question sequence collects the input data for the design. Once the required data are available to the program, it computes the expected electrical performance and gives details of the filter's mechanical construction. Finally, the program prints a graph of the passband and rejection skirts of the filter.

Two different examples of filters for two different ham bands are given. Each example gives general mechanical details of construction techniques which have been proven to give good results. The examples are just that, examples, and serve as guides to help the reader design and build filters which are optimized for his own particular application. The explanations of these techniques give enough information so that the examples themselves can be duplicated without too much difficulty.

440 MHz bandpass filter

The first example is a front-end filter for a 440 MHz FM receiver in a linear translator system. A linear translator, like a repeater, retransmits what it receives on a frequency 5 MHz (the 440 MHz spacing) away from the input channel. In order to prevent receiver overload or excessive intermodulation distortion in the presence of the strong transmitted signal, the rejection of the near-by repeater transmitter at 445 MHz must be approximately 50 dB. At the same time, the filter's passband loss should be moderate, less than about 3 dB, or the receiver sensitivity will be degraded excessively. These two considerations dictate the choice of filter type and design.

An interdigital bandpass filter turns out to be a good choice for this application. It can be simply and inexpensively built, and has relatively low passband losses together with good out-of-band rejection. And while it is true that a cavity diplexer filter of the sort often seen in amateur repeaters can give still lower losses and greater rejection, its good electrical performance can be accomplished only at the expense of increased mechanical complexity, larger size, and higher cost. Here, the required performance does not absolutely dictate the use of a multiple cavity diplexer, so we should definitely consider using the much simpler interdigital filter. If still less skirt rejection and somewhat greater loss can be tolerated, a helical resonator filter might be a good choice, for it would provide less rejection and probably would have higher passband losses, but it would also be much smaller mechanically than an equivalent interdigital bandpass filter. This size reduction is possible because the inductances in a helical filter are coils and because the inter-element couplings are not dependent only on the physical spacing of elements.

The first step in the choice of filter parameters is to determine the required passband loss limit, passband ripple, and out-of-band rejection that can be tolerated. An interactive computer-aided design program makes these tradeoffs considerably simpler to evaluate. The designer begins by entering a first estimate, or guess, of the approximate number of elements and passband ripple. From these inputs, the computer program quickly determines the approximate loss and plots the pass and reject bands. In the course of a few minutes' work, several different configurations can be tried out. From these it is easy to select the optimum design.

As an example of this interactive optimization, **fig. 3** lists a number of different 440 MHz filter configurations tested by the program. These designs differ mainly in the number of elements and in their passband widths. Filters with from 2 to 6 elements and ripple bandwidths of from 1 to 4 MHz are compared. In each case, the loss at the center frequency, 440 MHz, and at the transmitter frequency to be rejected, 445 MHz, are listed for comparison.

The results of this scan of various possible designs

```
# OF ELEMENT $ P-P RIPPLE IN PASSBAND (DB)? 4,.25
INPUT FILTER CENTER FREQ.(GHZ),BW(MHZ)&LOAD IMPEDENCE ZO? .440,3,50
INPUT GROUND PLANE SPACING , ROD DIAMETER
& DISTANCE TO CENTER OF FIRST AND LAST ROD? 1,.38,.5
NO. OF FREQ. REJECTION PTS AND STEP SIZE (MHZ)? 38,.5
DESIGN DATA FOR 4 POLE INTERDIGITAL FILTER .BAND PASS RIPPLE .25 DB
CENTER FREQ.  .44 GHZ
CUTOFF FREQ.  .4385  (GHZ) AND  .4415   GHZ
RIPPLE BW.   3.000021E-03 GHZ
3 DB BW.     3.419328E-03 GHZ
FRACTIONAL BW. 6.81823E-03
FILTER Q   128.6803
EST QU     1459.315
LOSS BASED ON THIS QU   2.422713   DB
DELAY AT BAND CENTER  294.6652 NANOSECONDS
FREQUENCY REJECTION INFORMATION
                                                                * .4305  69
                                                                * .431   67
                                                                * .4315  65
                                                              *   .432   63
                                                            *     .4325  61
                                                         *        .433   58
                                                       *          .4335  56
                                                    *             .434   53
                                                 *                .4345  50
                                             *                    .435   46
                                         *                        .4355  42
                                     *                            .436   38
                                *                                 .4365  33
                          *                                       .437   27
                  *                                               .4375  19
        *                                                         .438   9
*                                                                 .4385  0
*                                                                 .439   0
*                                                                 .4395  0
*                                                                 .44    0
*                                                                 .4405  0
*                                                                 .441   0
*                                                                 .4415  0
        *                                                         .442   9
                  *                                               .4425  19
                          *                                       .443   27
                                *                                 .4435  33
                                     *                            .444   38
                                         *                        .4445  42
                                             *                    .445   46
                                                 *                .4455  50
                                                    *             .446   53
                                                       *          .4465  56
                                                         *        .447   58
                                                            *     .4475  61
                                                              *   .448   63
                                                                * .4485  65
                                                                * .449   67
QUARTER WAVELENGTH = 6.706136 INCHES
THE LENGTH OF INTERIOR ELEMENTS = 6.418164  INCHES
LENGTH OF END ELEMENTS = 6.438183  INCHES
GROUND-PLANE SPACE = 1       INCHES
ROD DIAMETER  = .38  INCHES
END PLATES .5  INCHES FROM C/L OF END ROD
TAP EXTERNAL LINES UP  .2294638  INCHES FROM SHORTED END
LINE IMPEDANCES: END ROD 67.31341  ,OTHER  72.49873  , EXT. LINES  50 OHM
DIMENSIONS
EL. NO.       END TO C      C TO C      G(K)         Q/COUP
0                                       1             1.570873
1               .5                      1.378239      .6633376
                            1.894495
 2              2.394495                1.269327      .5431323
                            1.969942
 3              4.364437                2.055808      .6633376
                            1.894495
 4              6.258931                .8509719      1.570873
 5                                      1
                6.758931
```

fig. 4. Computer printout for the 440-MHz filter.

IC-02AT

IC-2AT

IC-02AT

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quickly reveals some expected trends. For instance, it is clear that filters that are too narrow (i.e., filters that have passband widths well under 1 percent of their center frequency) have increasingly higher passband losses. Also, filters with wide passbands have lower loss at the center frequency but decreased out-of-band rejection. These are fundamental tradeoffs. Next, it is also apparent that increasing the number of elements increases both the passband loss and the desired rejection of out-of-band signals, and that a filter with a relatively wide passband and many elements will have low in-band losses and good rejection. However, the number of elements cannot be increased arbitrarily because experience has shown that filters with many elements are less easily built and tuned, and that for Amateur construction and tuning techniques it is best to avoid the use of more than about five or six elements.

With all this in mind, we selected a four-section filter with passband ripple of 0.25 dB (VSWR = 1.62) and a bandwidth of 3 MHz. For our application, these choices resulted in approximately the desired loss and rejections, namely 2.4 dB loss at 440 MHz and a rejection of about 46.8 dB at 445 MHz. Note that this rejection is relative to the passband loss, so that the filter's total loss at 445 MHz is estimated to be approximately 49.2 dB.

computer output

The computer program's output for this filter design is shown in **fig. 4**. The printout contains information on both the electrical performance estimates as well as mechanical information in detail sufficient to fully describe the filter.

After the "RUN" command is entered, the program asks for some electrical design information such as the number of sections, or elements, in the filter and the passband ripple in decibels. Enter these two numbers, separated by a comma, followed by the "return" key, and the computer will respond with the next questions. These are the center frequency, expressed in gigahertz, the ripple bandwidth of the filter in megahertz, and the desired load impedance in ohms (usually 50). As before, these three entries should be separated by commas and followed by a return.

Next, the program requests some mechanical information. The spacing between the top and bottom ground planes, the diameter of the resonant rods and the desired spacing between the end of the filter and the first rod are entered in response to the questions. All of these dimensions should be entered in decimal inches because the design formulas within the program contain constants in inches.

The third and last section of input data concerns the plotter output. The operator must specify how many plot points and the spacing in megahertz between each point. The maximum number of plot points is 40. It is usually convenient to specify a step size of somewhat smaller than the passband width to give a good picture. In this example, the filter is 3 MHz wide and the plot with a one-half MHz resolution shows the rejection skirts clearly.

After these last bits of information are entered, the computer program proceeds without further intervention. It first prints out some of the calculated parameters of the filter. Then the computer graphically plots the pass and reject bands on the screen and on a printer if one is selected.

The last block of computer printout gives the mechanical details of the filter. The quarter wavelength listed is the inside dimension of the filter cavity. The length of the interior elements is listed, followed by the length of the two end elements. All of the interior elements have the same length, but the two end elements may have a different length from the end rods. The tap point is the point on the end element at which the external connection is made, and it is measured from the "cold" or grounded end of the rods. A tabular summary of the filter's dimensions appears at the end of the printout. It lists the end to center dimensions, the element center to center dimensions, and two coupling coefficients. The mechanical dimensions are easily translated into a sketch of the filter. Naturally, the designer must have certain dimensions and construction materals in mind before the computer program can be run. The ground plane spacing and rod diameter must be entered as constants to permit the completion of the design. These variables depend on the materials used to build the filter, and on the construction technique.

construction

Now that the computer program has been used to select an optimized filter and it has printed the mechanical dimensions of the filter's structure, it is time to consider how to translate the filter design into a form that can be realized. The mechanical structure must be sufficiently sound to produce the expected performance.

The table of numbers at the end of the computer printout, **fig. 4**, gives fairly complete information on the dimensions of the filter structure. The first column gives the element number, with zero indicating the first edge of the filter and a number one greater than the total number of rods denoting the other edge of the structure. The second column gives dimensions in inches from the end of the housing to the center of the rod indicated. For example, in our design, the second element is to be located a distance of 2.39 inches from the end of the housing. The total length of the filter will be the dimension listed opposite the final entry which is, in this case, 6.759 inches (17.168 cm).

Although the printout gives information on how to size the filter, it is probably a good idea to make a drawing to fix in your mind just how the box is to be assembled. **Figure 5** is a mechanical sketch of the 440 MHz filter. It is clear that the dimensions on this drawing have been taken directly from the computer printout, but the sketch also shows how the walls of the box are to be joined together. The dimensions on the printout are for the electrical housing, which consists of the copper conductor inside the box, and not the outside, mechanical dimensions. Keep in mind that the external dimensions of the box are not the critical factors.

One of the more popular amateur construction materials is copper clad printed circuit board. It is widely available at low cost, is strong and stable, and it can be easily joined together by regular soldering techniques to make boxes and circuit housings of any given size. It can be cut with a sheet metal shear or tin snips to fairly close tolerances, and so it is a good choice as the basic "building block" material for a custom bandpass filter. Furthermore, copper is one of the best conductors for use at high frequencies as its electrical resistivity is quite low. Of the more common materials, only silver has better high-frequency conductivity, and it is considerably more expensive.

To construct the filter housing of this example, we used 1/16 inch double-clad fiberglass epoxy board throughout. The top, bottom, sides and end pieces of the filter's housing were cut to shape with a sheet metal shear, drilled to accept the resonator rods, and soldered together with "tack" joints, that is with small flows of solder at intervals along the edges of the pieces to be joined. The pieces need not be soldered with a continuous seam, but a soldered tack should be used about every inch. This dimension corresponds to only about 5 percent of a wavelength, and so it ensures good electrical interconnection at all points along the copper.

The most critical dimensions of the housing are the width of the filter, which is usually a quarter wavelength, and the inside height of the cavity. A small piece of wood or metal with perpendicular faces is a good "jig" to help solder the housing walls accurately. If you work carefully, you should be able to build a housing that's both nearly square and accurate enough to take advantage of the custom design made possible with the computerized design aid described in this article.

Because we want to hold all of the parts of this filter together with regular tin-lead solder, all of the parts obviously should be solderable. A good material for the filter rods is common copper tubing which is available in a number of sizes in hardware and plumbing supply stores. For the design of this 440 MHz filter we chose 3/8 inch tubing. The "3/8 inch" refers to the outside dimension of the tubing, so the filter rod diameter measures approximately 0.375 inches.

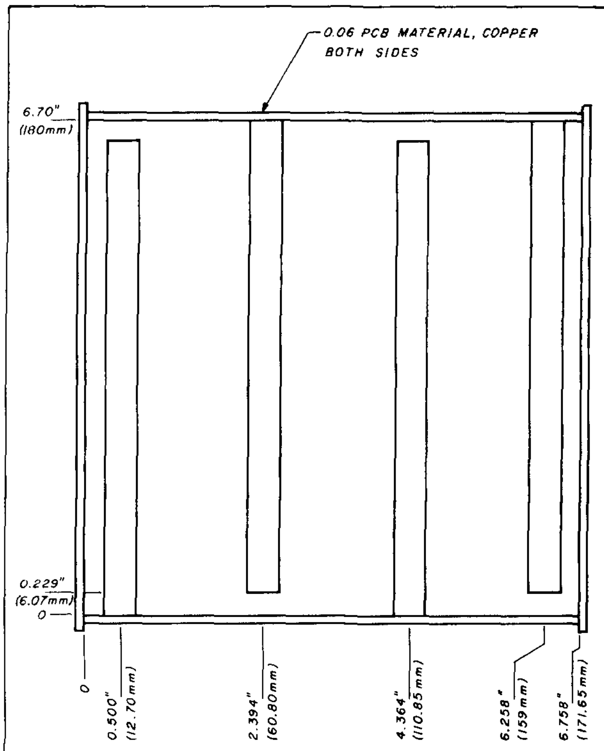

fig. 5. Sketch showing the top view of the 440-MHz filter. The end walls are deliberately cut longer so they can be soldered to both sides of the side walls.

The choice of rod diameter is not entirely arbitrary, although it is not very critical, either. As a rule of thumb, the rod diameter should be roughly 1/3 the housing height. In general, large diameter rods have lower losses than smaller rods. This is because skin effect losses predominate at radio frequencies. A larger diameter rod has greater surface area and, hence, less resistive loss than a smaller diameter rod. However, using a larger rod diameter can lead to mechanical difficulties. In this filter, for example, it was difficult to solder the large 3/8 inch diameter copper rods. These rods have fairly large masses and good thermal conductivity, so a regular soldering iron, intended for lighter duty circuit board work, just wouldn't provide enough heat. In the end, it took the greater heat of a propane torch to solder the rods to the copper walls.

The rods should be cut a bit longer than the correct length given by the printout. Then they are fit through holes drilled in the housing wall and soldered on both sides of the double clad circuit board, as shown in **fig. 6**. A good solder joint on each side increases the mechanical strength and improves grounding. The interior lengths of the rods and the center-to-center spacings between the rods are among the most critical dimensions affecting the frequency response of the filter, so measure them as carefully and precisely as possible. In spite of all the care you use in construction, however, it will almost certainly be necessary to peak tune the filter response.

One simple way to tune the filter rod lengths is to load their open-circuited ends with variable capacitors. If the rods are just a bit shorter than the design calls for, then the small capacitance of a tuning screw at the open end will tune that element's resonant frequency. This tuning compensates for the minor inaccuracies which inevitably occur in construction.

These tuning screws need good grounds at the points where they penetrate the wall. It is important to realize that it is the inside grounded surface that matters most, because the inside copper cladding forms the conductive boundary that contains the filter's electric fields. For this reason the tuning screws are supported by brass nuts soldered to the inside surface of the box. These nuts are visible in the overall photograph of the filter and especially in the close-up view, **fig. 7**. The nut makes a simple threaded support for the screw and serves as the low impedance path from the screw body to the ground plane. On the outside of the filter housing a second nut holds the screw firmly in place. This nut is tightened after all of the filter tuning is completed, and it ensures that the screw is tightly bound to the soldered-down nut and prevents it from moving and thereby detuning the filter.

The computer program also lists the tap point distance, which is the position at which the external connectors pass through the walls and are coupled to the filter. The distance is given relative to the shorted end of the rods. This junction should be by a short length of wire or cable. The closeup view shown in **fig. 7** shows how the connector's center pin has been joined to the end rod with a short length of solid wire. The actual tap point distances may need to be adjusted for best performance, but if the design value is used it will serve for most cases without change.

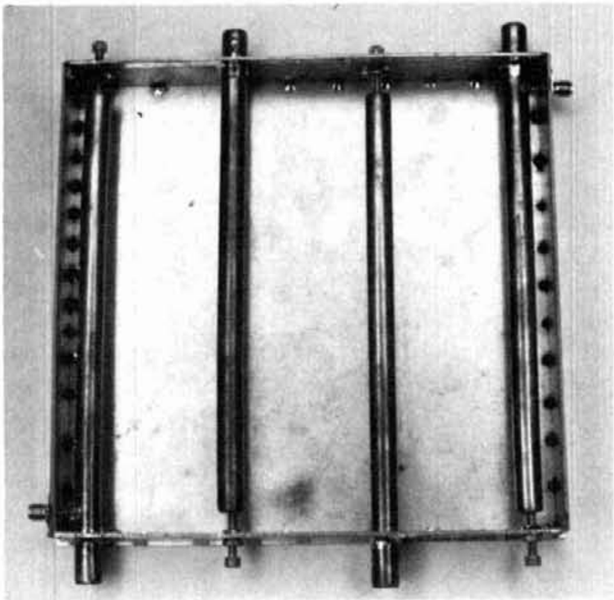

fig. 6. An overall view of the 4 section 440-MHz bandpass filter. The top cover has been removed to show the internal detail.

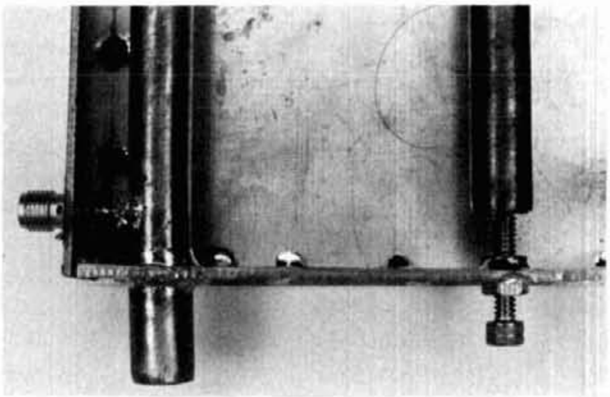

fig. 7. Close-up view of the 440-MHz filter showing details of the 50-ohm coaxial connector tap and of a tuning screw.

tuning

After the filter has been fully assembled, which means after all of the soldered joints are fully completed and the top is well secured electrically to the sides, it is time to test and tune.

Tuning microwave filters can be done in a number of ways. Several methods are described in reference 1, but the simple procedure of "sight tuning," or tuning by eye, is adequate for amateur filters, especially for filters with fewer than five or six sections. The basic principle involved is to tune for maximum signal at the center frequency, a process sometimes called synchronous tuning. The tuning is interactive, which means that one adjustment affects the tuning of adjacent elements, so it is necessary to return to each element once or twice to achieve peak performance.

A basic test set that can be used to peak the performance of any of the filters described here is shown in **fig. 8**. Inject a signal at the center frequency of the filter, detect or monitor the output power level, and tune for maximum. If the center frequency is within the ham bands, the input signal can be a transmitter or exciter. A signal generator can also be used. Harmonics of lower frequency sources or crystal oscillators can be useful as well. For example, the third harmonic of a 2-meter transmitter falls within the 420 to 450 MHz band and could be used to tune this filter. The signal detector can be a receiver with a signal level meter, a diode detector, a sensitive power meter or signal analyzer. Reference 6 describes many good low-cost UHF test methods in detail.

fig. 8. Simple test set block diagram. The signal source produces an output at the filter's center frequency. The filter is then tuned for maximum signal at the detector.

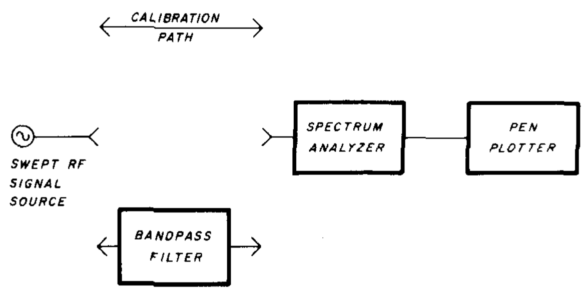

fig. 9. Test set used to produce the filter frequency response charts.

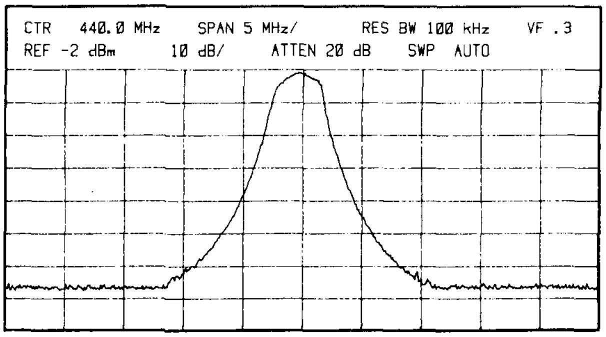

fig. 10. Swept frequency response of the four section 440-MHz filter. The horizontal scale is 5 MHz per division and the vertical scale is 10 dB per division. The spectrum analyzer settings were recorded automatically at the top of the plot.

Apply the test signal to one of the filter's connectors and connect the signal detector to the other and tune each of the screws to achieve the maximum output signal. The tuning range of this type of filter design is rather limited, which helps prevent tuning to the wrong harmonic, as can happen with broadly resonant circuits. This is a useful feature if the simple test set, with its potentially low spectral purity, is to be used.

If the filter is properly designed and carefully constructed, the slight tuning range afforded by the tuning screws should be sufficient. Each tuning screw should show a definite maximum. If it does not, this is an indication that the element which you are tuning is not properly resonant. In such a case, the rod length must be corrected before the filter will operate properly.

Once you've adjusted each tuning screw to yield the maximum output signal, go back and readjust each screw again slightly to peak the filter. If the filter response is not as calculated, and if the passband losses seem high even though each of the resonators gives a good peak tuning point, it may be necessary to adjust the tap points at the two end resonators. If no means of carefully measuring the losses is available, it is probably better to stay with the calculated tap dimensions. At this point the tuning is done.

A more sophisticated measuring system is useful for measuring the actual performance of the filter. **Figure 9** is a diagram of the test set which produced the swept frequency response of the filter examples. The input signal from a sweep generator scans across the frequency range in regular sweeps. A spectrum analyzer measures the output signal from the bandpass filter, and provides a graphic plot of the filter's output signal across the frequency range swept by the signal generator. Because the power from the generator to the filter input is constant, the spectrum analyzer's display is a direct representation of the filter's attenuation at various frequencies.

Figure 10 is the response of the 440-MHz filter. This graph is the plotter output from the test set described above. The filter was tuned using the simple single-frequency method of peaking all adjustments for maximum signal at 440 MHz. With the aid of such sophisticated test equipment it is possible to tune for improved passband flatness or for more precise centering if desired, but this clearly was not necessary in this case.

In summary, the 440 MHz bandpass filter example fully met its design goals of low cost, simple construction and easy adjustment, and it produced the desired electrical performance. The filter is physically small enough and sufficiently sturdy to be used in a fixed base system, although mechanical improvements would be needed for mobile service. This general construction technique using copper clad board and tubular copper or brass resonators has been applied to successfully build filters for the amateur bands from 440 to 2304 MHz.

1296 MHz filter

The second example of the use of this program is a bandpass filter centered at 1296 MHz with a desired ripple bandwidth of 20 MHz, a passband ripple of 0.25 dB and rejection points of 35 dB and 50 dB specified. Using these data, a few iterations with the program revealed that a four-section filter would again meet the requirements.

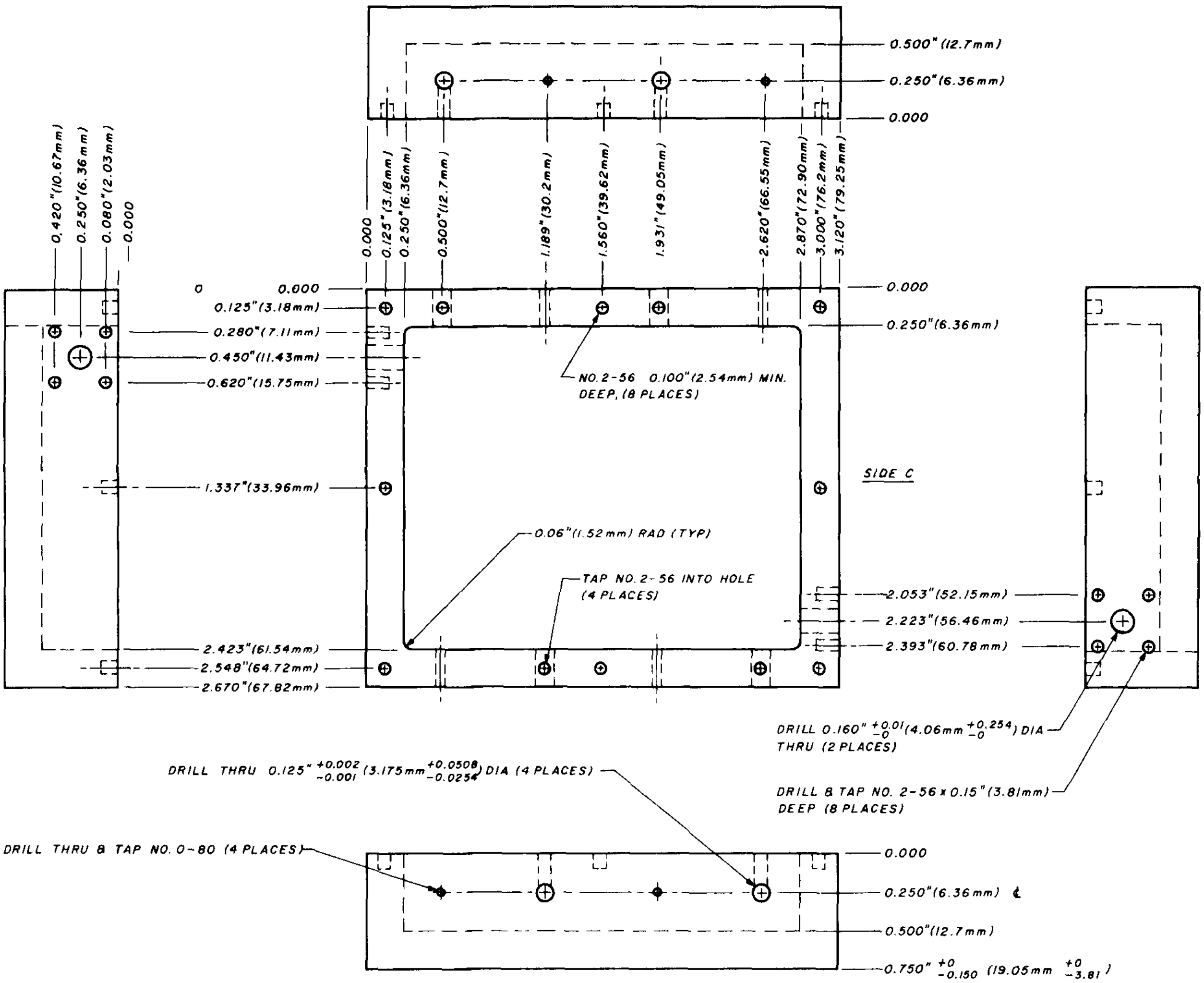

fig. 11. Mechanical details of the 1296-MHz filter housing. This housing is machined from aluminum stock. The top cover, not shown, is made from 0.06 sheet aluminum.

If the construction techniques used to build the 440-MHz filter seemed spartan, then this 1296 MHz filter is by contrast decidedly upscale. In order to prove that close tolerance construction could give good agreement with the computed data, a machined aluminum housing was used for this filter. Machining a housing using a metal milling machine gives precise control of the housing dimensions. This housing is considerably more accurate than the hand-made structure described in the first example.

The mechanical data supplied by the program were used to make a sketch, shown in **fig. 11**, of a housing that could be manufactured simply on a metal milling machine. The elements, machined from 0.125 inch brass rod stock (a standard size) were tightly "press fit" into holes in the housing walls. The rods are held tightly in position with a small setscrew once the exact interior lengths were determined. At the input and output ends of the filter SMA type connectors were installed so that their center pins contacted the rods at the tap point calculated by the computer program. Small diameter tuning screws were installed in the threaded holes in the walls opposite the open circuit end of each resonator rod. These screws, as in the 440-MHz filter, make it possible to fine-tune the filter passband. A top cover of 0.06-inch aluminum sheet was attached by eight screws that go into the threaded holes along the top edge of the housing. The photographs of this filter, **figs. 12** and **13**, show the construction details clearly. (Note: while 0-80 screws are specified in the construction drawings, 2-56 screws could be used as well.)

The housing was manufactured by machinists in a small shop, working from a simple sketch of the housing and cutting the aluminum stock with hand-operated machinery. In order to reduce costs, we

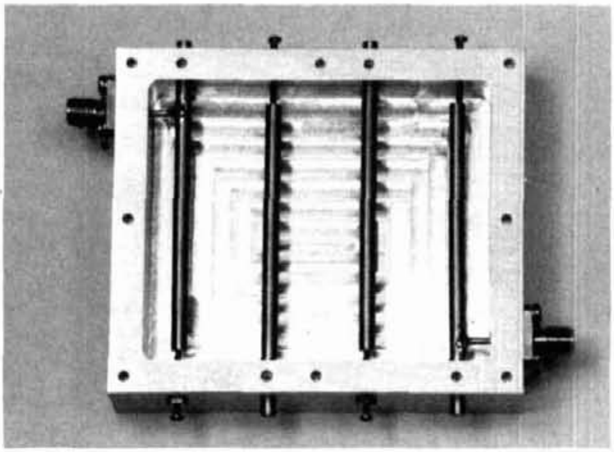

fig. 12. An overall view of the machined 1296-MHz filter. The sheet metal cover has been removed to show the internal cavity and the 4 resonator rods.

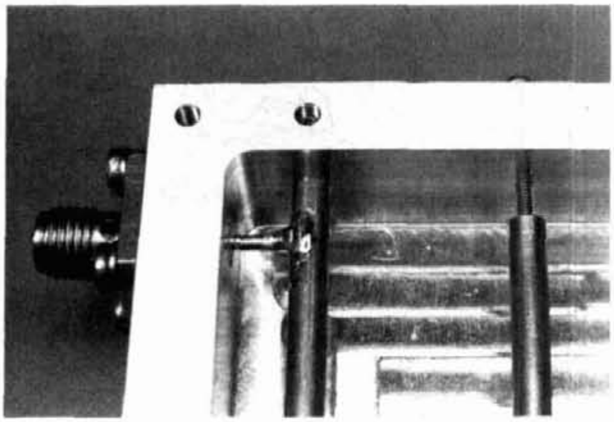

fig. 13. Close-up details of the 1296-MHz filter. The SMA connector center pin is clearly visible where it passes through the housing wall and contacts the end resonator rod. The small tuning screw at the end of the second resonator is also seen.

asked the shop to produce only a blank housing. They machined away the cavity to the precise 0.500 inch depth and drilled locating holes for the resonator rods, the tuning scews, the end connectors and the cover mounting screws, but they did not tap any of the holes. We were then able to finish the mechanical work by doing the time-consuming hand-tapping of all of the threaded holes. This reduced the cost of the housing by more than a third. Even so, the cost was in the $50.00 range, which may be justifiable only when precise results are essential.

However, the care and expense expended on the precise housing produced a filter that was nearly on frequency at first try, with precisely the initial resonator length settings that the computer predicted. Fine adjustments of the trimmer screws centered the passband precisely. **Figure 14** shows a plot of the swept RF response of the filter. Superimposed on the plot are circles that indicate the expected response calculated by the computer prediction. The calculated and actual values are in close agreement throughout the passband, and the rejection skirts are close to the computed values as well.

As before, the filter was peaked using the simple, single-frequency approach in order to illustrate the results obtainable with simple equipment. The test set used to make the swept frequency response plot was the same as that used in the 440 MHz filter tests.

This 1296 MHz filter, with its careful and precisely machined construction, shows the power and accuracy of the computer routine. The construction technique is a good one, and a simple metal milling machine, and perhaps even a drill press, can be used to make housings such as this one.

The use of 0.125 inch brass stock for the resonator rods was a bit of a compromise. It would have been better to use a somewhat larger diameter rod to reduce skin effect losses, but the 0.125-inch stock was on hand. Also, although brass rod is a good choice from a mechanical viewpoint, it is not a very good conductor of RF energy because its resistivity is about four times worse than copper's. Aluminum rods would be better than brass, because aluminum is both stronger and a better conductor, but with aluminum rods the tap point connections couldn't be easily soldered. Tradeoffs, as always, seemed to abound.

conclusion

This program is a powerful tool that greatly simplifies the selection and design of bandpass filters. The interdigital structure is useful from UHF to microwave frequencies, and provides good selectivity, low loss, small size, and an ease of construction that makes it suitable for many applications. The ease with which many different designs can be evaluated in software means that Amateurs can custom-design filters for specific applications and need not merely copy published designs that only approximate their re-

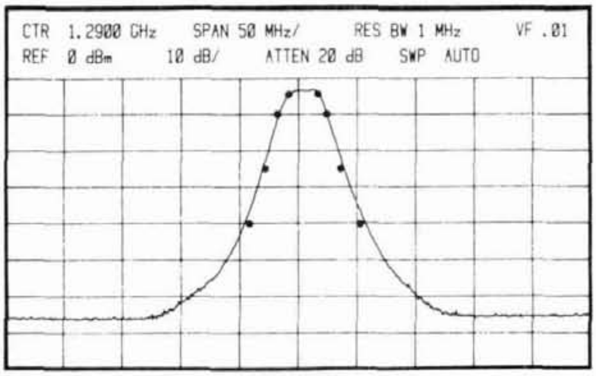

fig. 14. Frequency response curve of the 1296-MHz four section bandpass filter. The circles indicate the response computed by the program.

quirements. The widespread use of home computers together with software written specifically for Radio Amateurs should make possible a new generation of home-built equipment designs.

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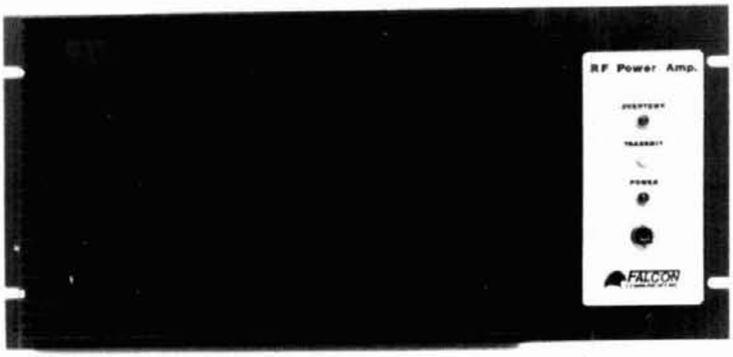
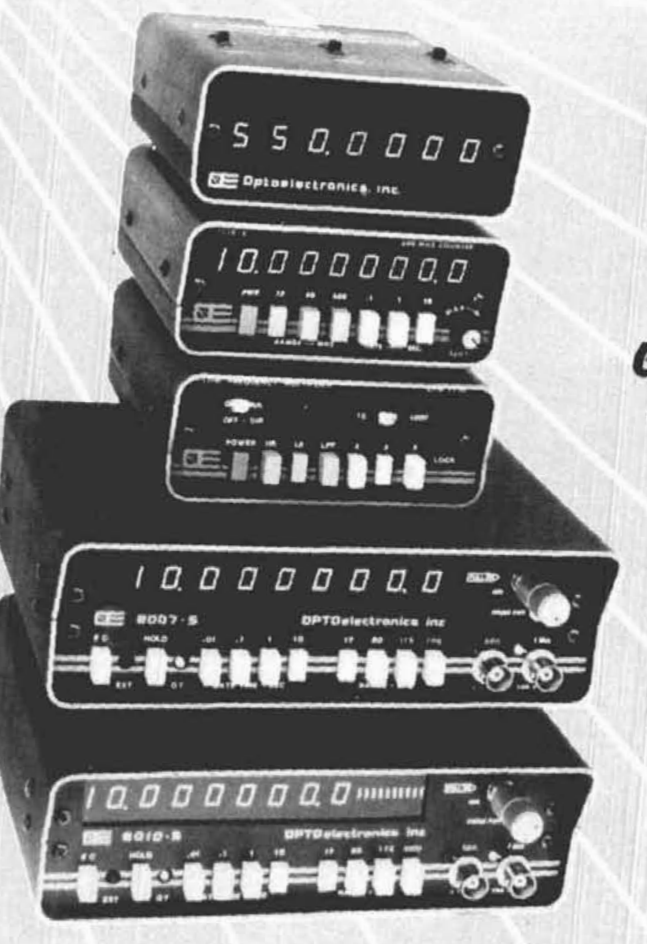
ham radio





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#7010-S/600 MHz	$235.	$25.	$75.	N/A
#7010-S/1 GHz	$310.	$25.	$75.	N/A
#8007-S	$350.	$60.	$75.	$125.
#8010-S	$435.	$60.	$75.	$125.
#8013-S	$499.	$60.	$75.	$125.

MODEL	RANGE (FROM 10 Hz)	TIME BASE		AVERAGE SENSITIVITY		GATE TIMES	MAX RESOLUTION					SENSITIVITY CONTROL	EXT CLOCK INPUT/OUTPUT	METAL CASE	PROBE POWER JACK
		FREQ	STAB-DESIGN	BELOW 500 MHz	ABOVE 500 MHz		12 MHz	17 MHz	60 MHz	175 MHz	MAX FREQ				
K-7000-AC	550 MHz	5.24288	±1 PPM-RTXO	15 mV −24 DBM	N/A	(2) .1, 1 SEC	10 Hz			100 Hz		No	No	Yes	No
7010-S	1 GHz 600 MHz	10.0 MHz	±1 PPM-TCXO *±0.1 PPM-TCXO	10 mV −27 DBM	20 mV −21 DBM	(3) .1, 1, 10 SEC	.1 Hz	1 Hz		10 Hz		Yes	No	Yes	No
8007-S	700 MHz														
8010-S	1 GHz	10.0 MHz	±1 PPM-TCXO *±0.1 PPM-TCXO *±0.05 PPM-OCXO	10 mV −27 DBM	20 mV −21 DBM	(4) .01, .1, 1, 10 SEC	.1 Hz		1 Hz		10 Hz	Yes	Yes	Yes	Yes
8013-S	1.3 GHz														

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basic gamma matching

Simplify antenna matching with this BASIC program and your microcomputer

Antenna homebrewers have found the gamma match to be an ideal choice for matching a coaxial feedline to an all-metal radiator. It is simple to build, adds little weight and wind loading, is very strong mechanically, and allows you to match an unbalanced transmission line to either an unbalanced or balanced antenna.

Unfortunately, it isn't always easy to obtain a good match, and many have aborted their attempts in frustration. The problem is that initial gamma match dimensions are generally chosen arbitrarily. Sometimes you may be lucky and choose a reasonable starting point, but just as often your initial dimensions won't even be close. In this case you may spend hours on the tower going in circles looking for a match.

Formulas that allow you to generate gamma match designs are available.[1] However, the math involved is tedious, especially if several iterations must be performed. A home computer can simplify these calculations, allowing a variety of gamma matching networks to be examined in just a few minutes. A program that will design a gamma match for practically any Yagi or vertical antenna is presented here. While designed for the Apple II +, the program will work equally well with any microcomputer with only a few modifications.

background

The design of the gamma match is represented in **fig. 1**, and the schematic of the equivalent electrical circuit[2] in **fig. 2**. The circuit consists of a gamma rod and a resonating capacitor. The gamma rod of diameter d and length L runs parallel to the driven element of diameter D, separated by the center to center spacing S. It provides the desired resistance transformation, but at the same time introduces inductive reactance at the feed point. The gamma capacitor compensates for the inductive reactance, leaving only a resistive component.

Any of several gamma capacitors may be used. An air variable with adequate plate spacing for the anticipated power level is the usual choice. It may be mounted in a small weather resistant enclosure and connected to the gamma rod by means of a feedthrough insulator. Another method is to construct a coaxial capacitor within the gamma rod. This technique has been successfully applied by many commercial manufacturers and basement engineers.[3]

The shorting bar that determines length *L* generally takes the form of a strap, bent to conform to the driven element and gamma rod and secured with screws. When very long gamma rods are required, as when shunt feeding towers on the lower frequencies, a wire may be used for the gamma rod. If the diameter

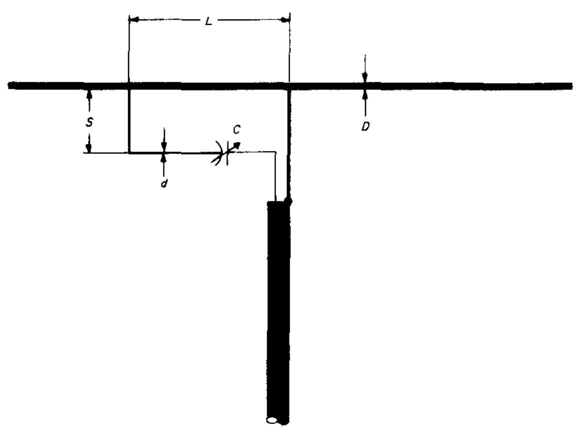

fig. 1. A representation of the mechanical design of the gamma match.

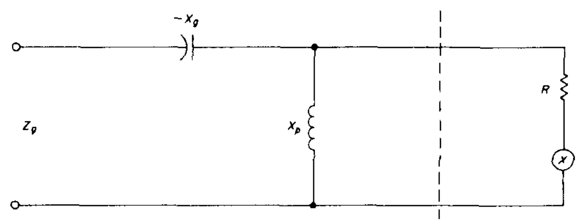

fig. 2. Schematic diagram of the gamma match. The antenna impedance R + jX is transformed to the feedline resistance Z_g by the gamma rod X_p and gamma capacitor $-X_g$.

By Richard A. Nelson, WB0IKN, Analog Technology, P.O. Box 8964, Fort Collins, Colorado 80525

```
10  HOME : CLEAR
20  PRINT "GAMMA MATCH DESIGN"
30  PRINT "BY RICHARD A. NELSON - WB0IKN"
40  PRINT
50  DEF  FN CSH(X) =  LOG (X +  SQR (X * X - 1))
60  PRINT "ENTER <M> FOR MONOPOLE"
70  INPUT "ENTER <D> FOR DIPOLE  > ";DM$
80  IF DM$ = "D" OR DM$ = "M" THEN  GOTO 100
90  GOTO 60
100  INPUT "ENTER FREQ IN MHZ > ";F
110  INPUT "ENTER FEEDPOINT RESISTANCE > ";RA
120  IF DM$ = "D" THEN RA = RA / 2
130  INPUT "ENTER FEEDPOINT REACTANCE > ";XA
140  IF DM$ = "D" THEN XA = XA / 2
150  INPUT "ENTER FEEDLINE RESISTANCE > ";RO
160  PRINT : PRINT
170  PRINT "(THE FOLLOWING ARE IN INCHES)"
180  PRINT : INPUT "ENTER DRIVEN ELEMENT DIAMETER > ";DE
190  INPUT "ENTER GAMMA ROD DIAMETER > ";DG
200  INPUT "ENTER GAMMA ROD SPACING > ";S
210 HZ = (1 + (( FN CSH((4 * S * S - DE * DE + DG * DG) / (4 * S * DG)))
     / ( FN CSH((4 * S * S + DE * DE - DG * DG) / (4 * S * DE))))) ^ 2
220 ZO = 60 *  FN CSH((4 * S * S - DE * DE - DG * DG) / (2 * DE * DG))
230 T = HZ / ZO
240 A = ((RO * XA) / (HZ * RA - RO))
250 B = (RO * ((RA) ^ 2 + (XA) ^ 2)) / (HZ * RA - RO)
260 Q = A +  SQR (A * A + B)
270 XS = HZ * ((RO * XA +  SQR ((RO * XA) ^ 2 + RO * (HZ * RA - RO) * ((
     RA) ^ 2 + (XA) ^ 2))) / (HZ * RA - RO))
280 LDGA =  ATN (Q * T)
290 LDG = (LDGA * 360) / (2 * 3.14159)
300 E = (RO / RA) * (((RA) ^ 2 + (XA) ^ 2) / Q)
310 G = (RO / RA) * XA
320 CR = 1000000 / (2 * 3.14159 * (E + G) * F)
330  HOME
340  IF DM$ = "D" THEN RA = RA * 2: IF DM$ = "D" THEN XA = XA * 2
350  PRINT
360  IF DM$ = "D" THEN  PRINT "DIPOLE ANTENNA"
370  IF DM$ = "M" THEN  PRINT "MONOPOLE ANTENNA"
380  PRINT
390  PRINT "FREQUENCY (MHZ) = ";F
400  PRINT "DRIVEN ELEMENT DIAM = ";DE
410  PRINT "GAMMA ROD DIAM = ";DG
420  PRINT "GAMMA ROD SPACING = ";S
430  PRINT "DRIVEN ELEMENT RESISTANCE = ";RA
440  PRINT "DRIVEN ELEMENT REACTANCE = ";XA
450  PRINT "FEEDLINE RESISTANCE = ";RO
460  PRINT
470  PRINT "GAMMA LENGTH (DEGREES) > ";LDG
480 FT = (948 / F) * (LDG / 360): PRINT "GAMMA LENGTH (FEET) > ";FT
490 IN = FT * 12: PRINT "GAMMA LENGTH (IN) > ";IN
500 CM = IN * 2.54: PRINT "GAMMA LENGTH (CM) > ";CM
510  PRINT "GAMMA CAP IN PF > ";CR
520  PRINT : PRINT "TYPE ANY KEY TO CONTINUE >": GET T$: GOTO 10
```

fig. 3. A listing of the program for an Apple II+ computer. Simple modifications will allow the program to be used with practically any home computer.

```
GAMMA MATCH DESIGN
BY RICHARD A. NELSON - WBØIKN

ENTER <M> FOR MONOPOLE
ENTER <D> FOR DIPOLE  > D
ENTER FREQ IN MHZ > 14.2ØØ
ENTER FEEDPOINT RESISTANCE > 2Ø.Ø
ENTER FEEDPOINT REACTANCE > +7.5
ENTER FEEDLINE RESISTANCE > 5Ø


(THE FOLLOWING ARE IN INCHES)

ENTER DRIVEN ELEMENT DIAMETER > 1.5
ENTER GAMMA ROD DIAMETER > .25
ENTER GAMMA ROD SPACING > 3.Ø

DIPOLE ANTENNA

FREQUENCY (MHZ) = 14.2
DRIVEN ELEMENT DIAM = 1.5
GAMMA ROD DIAM = .25
GAMMA ROD SPACING = 3
DRIVEN ELEMENT RESISTANCE = 2Ø
DRIVEN ELEMENT REACTANCE = 7.5
FEEDLINE RESISTANCE = 5Ø

GAMMA LENGTH (DEGREES) > 28.7984588
GAMMA LENGTH (FEET) > 5.34Ø55927
GAMMA LENGTH (IN) > 64.Ø867112
GAMMA LENGTH (CM) > 162.78Ø246
GAMMA CAP IN PF > 188.92213

TYPE ANY KEY TO CONTINUE >
```

fig. 4. The screen display for a sample run showing the sequence of data entry and output.

```
DIPOLE ANTENNA

FREQUENCY (MHZ) = 14.2
DRIVEN ELEMENT DIAM = 1.5
GAMMA ROD DIAM = .25
GAMMA ROD SPACING = 6
DRIVEN ELEMENT RESISTANCE = 2Ø
DRIVEN ELEMENT REACTANCE = 7.5
FEEDLINE RESISTANCE = 5Ø

GAMMA LENGTH (DEGREES) > 25.3194Ø33
GAMMA LENGTH (FEET) > 4.69538231
GAMMA LENGTH (IN) > 56.3445877
GAMMA LENGTH (CM) > 143.115253
GAMMA CAP IN PF > 241.894184


DIPOLE ANTENNA

FREQUENCY (MHZ) = 14.2
DRIVEN ELEMENT DIAM = 1.5
GAMMA ROD DIAM = .5
GAMMA ROD SPACING = 3
DRIVEN ELEMENT RESISTANCE = 2Ø
DRIVEN ELEMENT REACTANCE = 7.5
FEEDLINE RESISTANCE = 5Ø

GAMMA LENGTH (DEGREES) > 38.34Ø6199
GAMMA LENGTH (FEET) > 7.11Ø11495
GAMMA LENGTH (IN) > 85.3213794
GAMMA LENGTH (CM) > 216.7163Ø4
GAMMA CAP IN PF > 262.1ØØ816
```

fig. 5. Calculated results for two additional runs on the antenna in *fig. 4*.

is too small to provide a match, a "cage" may be constructed, using several wires to effectively create a cylinder at the required radius. Variations in gamma match design are limited only by the mechanical and electrical integrity of the structure. This versatility greatly adds to the gamma's usefulness.

about the program

A listing of the program for the Apple II+ computer is shown in **fig. 3**. Although written specifically for the Apple, I tried to keep the number of commands unique to Applesoft to a minimum. It should be easy for owners of other brands of microcomputers to translate the program for use on their machine.

Lines 10 through **40** clear the memory and the screen, and print the heading. **Line 50** defines the inverse hyperbolic cosine function needed in the calculations. **Lines 60** through **150** input data regarding the antenna and feedline, and **lines 180** through **200** input the driven element diameter and anticipated gamma match constants (gamma rod diameter and spacing). **Lines 210** through **320** perform the calculations of gamma rod length and capacitance. Note that since a gamma match essentially loads into only one side of a dipole, the feedpoint impedance of a dipole antenna must be divided in half before performing calculations. This is done by **lines 120** and **140**. **Line 340** restores the original impedance value before outputting data. **Line 330** clears the screen and **lines 360** through **510** output the specified and calculated data. **Line 520** returns the program to the beginning.

Once you have the program running, you may wish to add custom features. For example, you could change **line 520** to allow you to either return to the beginning, or to **line 180** if you want to evaluate a different gamma match design without changing the antenna parameters. If you'd like hard copy of the data, you could add the appropriate "printer on" and "printer off" commands between **lines 330** and **340**, and **510** and **520** respectively. For an Apple II the commands would look like this:

```
335 PR#1
515 PR#Ø
```

Another possibility is a "WILL NOT MATCH" message in response to a divide-by-zero error (for Apples, use ONERR GOTO), indicating that a different design should be tried. I included all of these in my

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original program, but they were eliminated in the version presented here in order to simplify program entry and translation.

The hardest part of using this program will be determining the driven element impedance. Fortunately, the feedpoint characteristics of most common antennas are sufficiently documented to provide good data. The error introduced by "guesstimating" will in many cases be better than typical homebrew construction tolerances. If you have a noise bridge or impedance meter you should have no trouble determining the feedpoint impedance. Refer to any of the standard antenna texts if you are unsure about a particular antenna.

Once you have the impedance data — assuming that you know the diameter of your driven element and gamma rod — select an arbitrary gamma rod spacing. Start with an estimate based on mechanical and "eyeballed" electrical considerations. Three inches is a bit small for a shunt-fed tower, and a foot is obviously too big for a 2-meter Yagi.

Load and run the program. It will begin by asking whether you are evaluating a dipole or a monopole. Then it will request the values of a number of constants — feedpoint impedance, design frequency, etc. After all data has been entered, the computer will calculate and display the gamma rod length and the value of the gamma capacitor. If these values are not acceptable, run the program again, trying a different gamma rod spacing, and/or a different gamma rod diameter (if practical). You can vary the feedpoint impedance by slightly changing the length of the driven element. By examining a variety of alternative designs you can find the best combination, with reasonable mechanical and electrical parameters. If the antenna is not suitable for gamma matching, you can learn this quickly, without wasting hours in hands-on experimentation.

design examples

To demonstrate program operation, and to allow you to check operation once the program is keyed in, I will present several design examples.

Figure 4 shows the screen display on a run for a computer generated six-element, 20-meter Yagi design by Lawson.[4] The calculated feedpoint impedance of the antenna is 20 + j7.5 ohms. In this example, I've assumed 1.5-inch (38.1-mm) diameter element centers, a gamma rod diameter of 0.25 inch (6.35 mm), and a gamma rod spacing of 3 inches (76.2 mm). **Figure 5** shows the results of two additional gamma match designs for this antenna. In **fig. 5A** the gamma rod spacing has been increased to 6 inches (152.4 mm), and in **fig. 5B** the gamma rod diameter has been increased to 0.5 inch (12.7 mm).

Figure 6 shows the results for a monopole approximately 1/4-wavelength high. In this case the gamma rod is a length of No. 10 wire (approximate diameter = 0.1 inch or 2.54 mm). Run A is for a 60-foot (18.3-meter) tower used as vertical radiator on 3.8 MHz. Computer analysis shows its impedance to be approximately 33 + j1.3 ohms. Run B is for a 55-foot (16.8-meter) tower operated on the same frequency. The results show how a smaller gamma capacitor may be used if the radiator is made capacitively reactive by reducing its overall height.

```
MONOPOLE ANTENNA

FREQUENCY (MHZ) = 3.8
DRIVEN ELEMENT DIAM = 12
GAMMA ROD DIAM = .1
GAMMA ROD SPACING = 12
DRIVEN ELEMENT RESISTANCE = 33
DRIVEN ELEMENT REACTANCE = 1.3
FEEDLINE RESISTANCE = 50

GAMMA LENGTH (DEGREES) > 44.8174331
GAMMA LENGTH (FEET) > 31.0576949
GAMMA LENGTH (IN) > 372.692339
GAMMA LENGTH (CM) > 946.638541
GAMMA CAP IN PF > 122.616005


MONOPOLE ANTENNA

FREQUENCY (MHZ) = 3.8
DRIVEN ELEMENT DIAM = 12
GAMMA ROD DIAM = .1
GAMMA ROD SPACING = 12
DRIVEN ELEMENT RESISTANCE = 25
DRIVEN ELEMENT REACTANCE = -38
FEEDLINE RESISTANCE = 50

GAMMA LENGTH (DEGREES) > 53.7315208
GAMMA LENGTH (FEET) > 37.2350012
GAMMA LENGTH (IN) > 446.820015
GAMMA LENGTH (CM) > 1134.92284
GAMMA CAP IN PF > 76.9934529
```

fig. 6. Results of two runs on a 75-meter vertical monopole antenna.

I have used this program to design gamma matching networks for a 20-meter Yagi and a 2-meter cubical quad. In both cases the results were superb. No trimming of the gamma rod was necessary. Adjustment of the variable capacitors was all that was needed to achieve a 1:1 VSWR. The normal trial and error process was totally eliminated.

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ham radio

determining basic moon coordinates

Track the moon with this computer program

Serious EME (Earth-Moon-Earth) communication requires a way of accurately determining the position of the moon. Of the several ways this can be done, moon-tracking by computer is one of the more convenient.

The program described here runs on TRS-80 models I/III with a minimum of 16K RAM and either level II/III or disk BASIC; it should run on other computers as well with only minimal changes. With your QTH as the point of reference, it provides readings of azimuth, elevation, GHA, declination, and right-ascension of the moon as well as relative path loss, local time of day and local sidereal time (LST). The positional data are expressed in *topocentric* values — i.e., they've been corrected for your location on the surface of the earth. (Many other moon tracking programs provide *geocentric* data referenced to the center of the earth.)

To output the above information, you simply input your latitude and longitude and the day(s) and time(s) in GMT (Greenwich Mean Time). The computer displays the information on the CRT monitor, and/or sends it to the printer. **Figure 1** shows a sample output for W2WD (40 degrees 39 minutes North, 74 degrees 22.5 minutes West) on June 8, 1983. Although this output came from the printer, the CRT display was similar; the data appears on the printer at the same time it's displayed on the CRT.

keyboard input data

Figure 2 shows a screen printout of the inputs required to produce the output data discussed above. After loading, RUN "MOON" <ENTER> starts the program from either level II or disk BASIC. Once initiated, the logo appears and suggests a listing of lines 1000-1360 if you need a short description of the program. If you hit <ENTER> to go on, the computer asks, WHAT ARE THE STATION CALL LETTERS? (As shown, I entered W2WD.) Next, the computer asks WHAT IS YOUR LOCAL TIME (EST, EDT, PST, etc.)? (The reply was EST.) and then, HOW MANY HOURS/MINUTES DIFFERENCE FROM GMT? USE + IF EARLIER, − IF LATER (e.g., EST would be −0500). (The answer was −0500.) Next it asks, YOUR LATITUDE (DEGREES, MINUTES) +NORTH/ −SOUTH? (In my case this was 40,39 for 40 degrees and 39 minutes north latitude. If you are in the southern hemisphere, you must use a negative sign, e.g., −40,39.) Longitude is entered in a similar fashion. (My input was 74,22.5 for 74 degrees and 22.5 minutes west longitude.) Use a negative sign (−) for east longitude.

The accuracy of the final calculations depends on the precisely accurate determination of your geodetic location. Your town manager or city engineer may be able to help. If not, you can use maps sold by the U.S. Geological Survey, especially those drawn to 1:24,000 scale. Some automotible road maps may show sufficient detail to allow fairly close determination of your location.

The next information to be inputted is the DESIRED PRINTING INCREMENT IN MINUTES (1-60). I used 20 minutes between each line of output data, but this can be anything between 1 and 60 minutes. Positions can be determined for time intervals of less than 1 minute apart, but the displays and printouts, as the program is now written, will round the results to increments of full minutes — i.e., if you need data every 6 seconds (0.1 minutes), you can input 0.1 as the print-

By Warren Butler, W2WD, 2305 Morse Avenue, Scotch Plains, New Jersey 07076

POSITION OF THE MOON ON 6 / 8 / 1983 GMT FROM W2WD

GMT	GHA	DEC	LST	EST	AZ	EL
0800	337.2	+10.5	2006	0300	77.4	1.4
0820	342.0	+10.5	2026	0320	80.5	5.1
0840	346.8	+10.6	2046	0340	83.5	8.7
0900	351.7	+10.7	2106	0400	86.5	12.5
0920	356.5	+10.8	2126	0420	89.6	16.2
0940	1.4	+10.9	2147	0440	92.7	19.9
1000	6.2	+10.9	2207	0500	95.9	23.6
1020	11.1	+11.0	2227	0520	99.2	27.4
1040	15.9	+11.1	2247	0540	102.7	31.0
1100	20.8	+11.2	2307	0600	106.4	34.6
1120	25.7	+11.3	2327	0620	110.4	38.2
1140	30.6	+11.3	2347	0640	114.7	41.7
1200	35.5	+11.4	0007	0700	119.4	45.0
1220	40.3	+11.5	0027	0720	124.6	48.2
1240	45.2	+11.6	0047	0740	130.3	51.2
1300	50.1	+11.6	0107	0800	136.8	54.0
1320	55.0	+11.7	0127	0820	144.1	56.4
1340	59.9	+11.8	0147	0840	152.1	58.5
1400	64.8	+11.9	0207	0900	161.0	60.0
1420	69.7	+11.9U	0227	0920	170.5	61.0
1440	74.6	+12.0U	0247	0940	180.4	61.4
1500	79.5	+12.1U	0307	1000	190.4	61.1
1520	84.4	+12.2U	0327	1020	200.0	60.2
1540	89.3	+12.2U	0348	1040	208.9	58.7
1600	94.2	+12.3U	0408	1100	217.1	56.7
1620	99.1	+12.3W	0428	1120	224.4	54.4
1640	103.9	+12.4W	0448	1140	231.0	51.7
1700	108.8	+12.5W	0508	1200	236.8	48.7
1720	113.7	+12.5W	0528	1220	242.1	45.6
1740	118.6	+12.6W	0548	1240	246.9	42.3
1800	123.5	+12.7W	0608	1300	251.3	38.9
1820	128.3	+12.7W	0628	1320	255.3	35.4
1840	133.2	+12.8J	0648	1340	259.1	31.8
1900	138.1	+12.8J	0708	1400	262.6	28.2
1920	142.9	+12.9J	0728	1420	266.0	24.6
1940	147.7	+12.9J	0748	1440	269.3	21.0
2000	152.6	+13.0J	0808	1500	272.5	17.3
2020	157.4	+13.1J	0828	1520	275.6	13.7
2040	162.3	+13.1	0848	1540	278.7	10.1
2100	167.1	+13.2	0908	1600	281.8	6.5
2120	171.9	+13.2	0928	1620	284.9	3.0

R.A. OF MOON = 0308 PATH-LOSS INCREASE + 1.3 DB

fig. 1. Moon coordinate printout for W2WD on 8 June 1983.

ing increment. You'll get ten different lines of data for every minute of readout, but the GMT, LST, and your local time will not change until the data for the following minute appears, etc. If this is an important use, you can modify the program to print out the time(s) to the resolution needed.

For EME operators who use ground reflection for added gain or cannot elevate their antennas, the pro-

```
WHAT ARE THE STATION CALL LETTERS ? W2WD
WHAT IS YOUR LOCAL TIME (EST, EDT, PST, ETC.) ? EST
HOW MANY HOURS/MINUTES DIFFERENCE FROM GMT?
    USE + IF EARLIER, - IF LATER
    (EG., EST WOULD BE -0500)? -0500
WHAT IS YOUR LATITUDE (DEGREES,MINUTES)
+ NORTH  /  - SOUTH  ? 40,39.0
WHAT IS YOUR LONGITUDE (DEGREES,MINUTES)
+ WEST  /  - EAST  ? 74,22.5
WHAT IS THE DESIRED PRINTING INCREMENT IN MINUTES (1-60)? 20
DO YOU ONLY WANT PRINTOUT WHEN THE MOON IS NEAR THE HORIZON
(YES/NO)? NO
DO YOU WANT HARDCOPY (YES/NO)? YES


INPUT - GMT MONTH, DAY, YEAR, TIME BEGINNING, TIME ENDING
USE 4-DIGITS FOR YEAR AND 24-HOUR CLOCK
ENTER DATA FOR UP TO 31 DAYS.
HIT <ENTER> AFTER LAST ENTRY

DATE 1  (MM,DD,YYYY,TTTT,TTTT) ? 6,8,1983,0,2400
DATE 2  (MM,DD,YYYY,TTTT,TTTT) ? 0_
```

fig. 2. Screen printout of input commands.

gram has an option to print or display data only when the moon is near the horizon. In this case you answer YES to DO YOU ONLY WANT PRINTOUT WHEN THE MOON IS NEAR THE HORIZON (YES/NO)? (I operate with a polar-mounted antenna so I answer NO to this question.) If you answer YES, you will be asked to reply to BELOW WHAT ELEVATION IN DEGREES DO YOU WANT PRINTOUT? You then enter the maximum elevation angle you're interested in.

If you don't have a printer or don't want hard copy, answer NO to DO YOU WANT HARDCOPY (YES/NO)? If you have a printer and answer YES but the printer is not turned on, the computer will reply PRINTER NOT READY. The program won't hang up, but will repeat the question until you either answer NO or turn on the printer.

At this point, you specify the date-time periods for the output data. These must be entered in GMT format. It is necessary to use *four* digits for the year; don't use 85, if you mean 1985! Similarly, time is inputted with four digits using a 24-hour clock for both the starting and ending points. You don't have to start the data at 0000 and end at 2400; they can be set according to your operating requirements. For the low-elevation-only option, you don't need to enter time spans because the computer will print only data meeting your criteria of when the elevation of the moon is below the angle you have specified earlier.

You can enter up to 31 dates for any given run. The dates do not have to be consecutive or in any given order, nor does the data requested have to be uniform from day to day. In fact, each entered date is a separate request. You may repeat the same dates with different start/end times if you wish.

program operation

Keep in mind that the calculations are relatively slow. Compiling data for 31 days could require hours of computer time. The actual time will vary greatly according to the time increments you specify, the printer speed, and other factors.

When you've completed the entry of all start/end times, hit <ENTER> to terminate the data input phase. The program prints the heading information and begins calculations. Each line of data requires roughly 15 seconds, even though the result of that calculation may be below the horizon and therefore not printed out. If you've asked for data every 20 minutes following midnight on a day when the moon does not rise until 0800 GMT, the computer will calculate the position of the moon 24 times before any results appear on the CRT or printer. Therefore, the program may seem to be hung up for several minutes even though it's actually hard at work calculating data that ends up below the horizon and is consequently not displayed.

fig. 3. Enhanced version moon coordinate program listing for TRS-80.

```
1000 '************ MOON COORDINATES ************
1010 '
1020 'PRIMARILY FOR USE IN EARTH-MOON-EARTH
1030 '(EME) COMMUNICATIONS BY RADIO AMATEURS.
1040 '
1050 'BASED ON PROGRAMS BY LANCE COLLISTER (WA1JXN/WA3GFL)
1060 'AND JAY LIEBMANN (K5JL).
1070 'SEE EIMAC PUBLICATIONS AS-49-6, AS-49-17 AND AS-49-24.
1080 'VARIAN, EIMAC DIVISION
1090 '301 INDUSTRIAL WAY
1100 'SAN CARLOS, CA 94070.
1110 '
1120 'MODIFIED FOR MODEL I TRS-80 LEVEL II AND DISK BASIC
1130 'WITH ENHANCED DISPLAYS AND ADDITIONS OF SIDEREAL TIME,
1140 'RIGHT ASCENSION AND DISTANCES TO THE MOON (CONVERTED
1150 'TO PATH-LOSS VARIATIONS IN DECIBELS)
1160 'BY WARREN BUTLER (W2WD).
1170 '
1180 'INPUT DATA: LATITUDE, LONGITUDE, GMT DATE/TIME AT SITE.
1190 'DATA FOR UP TO 31 DIFFERENT DAYS CAN BE INPUTTED AT
1200 'ONE TIME.  ENTER DATA IN THE FORMAT REQUESTED.  AFTER
1210 'THE LAST INPUT, INSERT ZEROS OR HIT <ENTER>.
1220 '
1230 'OUTPUT DATA: GHA, DECLINATION, AZIMUTH AND ELEVATION OF
1240 'MOON, SIDEREAL TIME (ST) AND LOCAL TIME, UNIVERSAL
1250 'WINDOWS FOR EME COMMUNICATION, RIGHT ASCENSION OF MOON,
1260 'PATH-LOSS VARIATIONS (DB).
1270 '
1280 'HARDCOPY OUTPUT CAN BE SELECTED IF PRINTER IS AVAILABLE.
1290 '
1300 'UNIVERSAL EME WINDOWS ARE SHOWN BY LETTERS FOLLOWING DEC.
1310 'U = EUROPEAN UNIVERSAL WINDOW
1320 'W = W/VE UNIVERSAL WINDOW
1330 'J = J/VK/ZL UNIVERSAL WINDOW
1340 '
1350 'BE PATIENT, THE CALCULATIONS CAN TAKE SEVERAL MINUTES.
1360 '
1370 PRINT
1380 PRINT
1390 CLEAR 500
1400 DIM F(31), V(31), Y(31), Q(31), S(31)
1410 P5=2.0000000000*3.1415926535
1420 D5=360.0000000000/P5
1430 R5=P5/360.0000000000
1440 CLS
1450 'GOTO 1730
1460 PRINT @ 0, STRING$(64, 170)
1470 PRINT @ 148, "MOON COORDINATE PROGRAM"
1480 PRINT @ 278, "WARREN BUTLER (W2WD)"
1490 PRINT @ 343,"2305 MORSE AVENUE"
1500 PRINT @ 404,"SCOTCH PLAINS, NJ 07076"
1510 PRINT @ 473,"(201) 233-4460"
1520 PRINT @ 576, STRING$(64, 170)
1530 PRINT @ 833,"IF YOU NEED A DESCRIPTION OF PROGRAM, LIST LINES
     1000-1360"
1540 PRINT @ 978, "ELSE HIT <ENTER> TO CONTINUE";
1550 INPUT I%
1560 CLS
1570 REM: BEGIN INPUT DATA SEQUENCE
1580 PRINT"WHAT ARE THE STATION CALL LETTERS ";
1590 INPUT W$
1600 INPUT"WHAT IS YOUR LOCAL TIME (EST, EDT, PST, ETC.) ";TL$
1610 PRINT"HOW MANY HOURS/MINUTES DIFFERENCE FROM GMT?"
1620 PRINT"     USE + IF EARLIER, - IF LATER"
1630 INPUT "     (EG., EST WOULD BE -0500)";TD
1640 IF TD<-1200 OR TD>1200 THEN 1610
1650 PRINT"WHAT IS YOUR LATITUDE (DEGREES,MINUTES) "
1660 PRINT"+ NORTH  /  - SOUTH  ";
1670 INPUT L5, U5
1680 IF L5>90 OR L5<-90 OR U5>60 THEN 1650
1690 PRINT"WHAT IS YOUR LONGITUDE (DEGREES,MINUTES) "
1700 PRINT"+ WEST   /  - EAST   ";
1710 INPUT L6, U6
1720 IF L6>180 OR L6<-180 OR U6>60 THEN 1690
1730 'W$="W2WD": L5=40: U5=39.0: L6=74: U6=22.5: TL$="EST": TD=-0500
1740 L5=(L5+U5/60)*R5
1750 L6=(L6+U6/60)*R5
1760 INPUT"WHAT IS THE DESIRED PRINTING INCREMENT IN MINUTES (1-60)";I
1770 IF I<=0 OR I>60 THEN 1760
1780 B$=""
1790 INPUT"DO YOU ONLY WANT PRINTOUT WHEN THE MOON IS NEAR THE HORIZON
       (YES/NO)";B$
```

(continued on page 43)

This happens in the example shown in **fig. 1**. On June 8, 1983, the moon rose at about 0800 GMT. The computer did not start printing out data until *six minutes* after the input phase had been completed and the <ENTER> key had been pressed. From that point the printing continued until moonset had been reached at about 2120 GMT. However, the program did not stop calculations until the specified ending time of 2400 GMT. The program required about 15 minutes to generate the data shown in **fig. 1**.

Similarly, if data ends above the maximum angle selected in the low-elevation mode, the program will also appear to be "hung-up." If you already know the times of moonrise and moonset, the start and end times can be set accordingly and much time can be saved in the printout process.

printouts and displays

Now let's look at the output data (**fig. 1**) in more detail. After the GMT time column, the GHA, or Greenwich Hour Angle, is shown. It is the angle subtended by the moon and the Greenwich meridian. This angle can be translated to your geographical meridian (Local Hour Angle) by adding your longitude if east, or subtracting if west. The third column provides the declination of the moon or its position north (+) or south (−) of the celestial equator. Using the GHA and declination information, the polar-mounted EME antenna can be kept trained on the moon.

Note that some of the values for declination are followed by the letter U, W, or J. These are indicators that the moon is in one of the universal EME windows established to allow use of fixed antenna arrays. This was intended to permit large antennas to be built at lower cost by eliminating the need to position the array in azimuth or elevation. The U, or universal window, was set up for contacts between European and USA stations. The W window is better situated for USA/USA and the J window is for JA/VK/W contacts.[4]

The next two columns show local time readouts. Standard or daylight-savings time needs no further discussion. LST (Local Sidereal Time) is tied to the stars and used primarily by astronomers. It is included here to help readers locate some of the stellar noise sources for checking EME systems performance. Or, for that matter, to locate a "cold" area of the sky for the same purpose. Many EME operators use the sun as a noise source to check antenna and system performance. When system performance is improved beyond the norm, smaller noise sources in the heavens such as Cassiopeia A, Cygnus A, and others can prove useful. Knowing the R.A. (right ascension of the moon) and declination of these sources and your LST can help in training your antenna to the proper direction in the sky.

For EME operators using AZ-EL antenna mounts, the computer calculates the azimuth and elevation angles to the moon from their QTH. Azimuth is the angle with respect to true north, (not magnetic north as would be indicated with a compass); elevation is the angle above the local horizon.

Below the positional data in **fig. 1** are two other useful parameters for EME operation. The first is the R.A., which indicates where the moon is in relation to other objects in the sky. Some of the stars are noise sources that can reduce the signal-to-noise ratio of the communication path when they are in line with the moon at the time of a QSO. Earlier, I suggested that these same kinds of noise sources could be useful for system calibration. For better EME contacts, however, they should be avoided. Reference 1 contains radio sky maps over the frequency range of 64 to 910 MHz. These show the noise temperature of the sky as a function of right ascension and declination. It should be mentioned that the right ascension of the moon given at the bottom of the printout listings is the first calculated value. The R.A. varies about one hour over a 24-hour day, but this usually provides sufficient accuracy to judge whether noise sources from behind the moon are going to be a problem.[2]

Because the moon is traveling in an elliptical path around the earth, its distance from the earth varies. At apogee it is roughly 407,000 kilometers from the earth; at perigee, 356,000 km. The variation in range is sufficient to add about 2 dB of additional path loss at apogee as compared to when the moon is at the perigee position. Apogee to perigee time spans are roughly 13 days. For each day of a printout, the path-loss difference in dB, as compared to the perigee position, is printed at the bottom of the listing. Equations used for calculating the earth-to-moon distances were taken from reference 3.

accuracies

Considering the number and types of calculations this program processes, it is reasonable to ask whether the results obtained on a personal computer are sufficiently accurate for use by the typical Amateur EME station. To answer this question, I compared the output data to the most accurate, reliable data available to me.

For GHA and declination I was unable to locate a suitable source of topocentric values to confirm the output of this program. However, the *Nautical Almanac* provides geocentric data with errors less than 0.0005 degrees and thus served as a satisfactory standard for data referenced to the center of the earth.[5,6] In order to compare "oranges to oranges," I had to bypass the translational calculations I had inserted to convert from geocentric to topocentric coordinates. I then used the random number generator of the

fig. 3, continued

```
1800 IF LEFT$(B$, 1)="Y" THEN 1830
1810 I6=100
1820 GOTO 2030
1830 INPUT"BELOW WHAT ELEVATION IN DEGREES DO YOU WANT PRINTOUT";I6
1840 WW$=" "
1850 INPUT"DO YOU WANT HARDCOPY PRINTOUT (YES/NO)";WW$
1860 IF LEFT$(WW$, 1)="Y" WW$="YES"
1870 IF WW$="YES" AND PEEK(14312) > 127 THEN PRINT "PRINTER NOT READY":
     WW$="": GOTO 1850
1880 PRINT
1890 PRINT"WHAT ARE THE GMT MONTH, DAY, YEAR DESIRED?"
1900 PRINT"*** NOTE - USE 4-DIGITS FOR YEAR (EG., 1983) ***"
1910 PRINT"ENTER DATA FOR UP TO 31 DAYS"
1920 PRINT"HIT <ENTER> AFTER LAST ENTRY"
1930 PRINT
1940 N=0
1950 FOR N=1 TO 31
1960 PRINT"DAY ";N;"  (MM,DD,YYYY)";
1970 INPUT F(N), V(N), Y(N)
1980 IF F(N)=0 THEN 2220
1990 IF F(N)<1 OR F(N)>12 OR V(N)<1 OR V(N)>31 OR Y(N)<1900 OR Y(N)>2000
     THEN 1960
2000 IF N=31 THEN 2220
2010 NEXT N
2020 GOTO 1950
2030 WW$=""
2040 INPUT"DO YOU WANT HARDCOPY (YES/NO)";WW$
2050 IF LEFT$(WW$, 1)="Y" WW$="YES"
2060 IF WW$="YES" AND PEEK(14312) > 127 THEN PRINT "PRINTER NOT READY":
     WW$="": GOTO 2040
2070 PRINT
2080 PRINT"INPUT - GMT MONTH, DAY, YEAR, TIME BEGINNING, TIME ENDING"
2090 PRINT"USE 4-DIGITS FOR YEAR AND 24-HOUR CLOCK"
2100 PRINT"ENTER DATA FOR UP TO 31 DAYS."
2110 PRINT"HIT <ENTER> AFTER LAST ENTRY
2120 PRINT
2130 N=0
2140 FOR N=1 TO 31
2150 PRINT"DATE";N;" (MM,DD,YYYY,TTTT,TTTT) ";
2160 INPUT F(N), V(N), Y(N), Q(N), S(N)
2170 IF F(N)=0 THEN 2220
2180 IF F(N)<1 OR F(N)>12 OR V(N)<1 OR V(N)>31 OR Y(N)<1900 OR Y(N)>2000
     OR  Q(N)<0 OR Q(N)>2359 OR S(N)<0001 OR S(N)>2400 THEN 2150
2190 IF N=31 THEN 2220
2200 NEXT N
2210 GOTO 2140
2220 IF N=31 THEN N5=N ELSE N5=N-1
2230 FOR N=1 TO N5
2240 IF LEFT$(B$, 1)="Y" THEN 2260
2250 GOTO 2290
2260 E1=2400
2270 B=0
2280 GOTO 2310
2290 E1=S(N)
2300 B=Q(N)
2310 M=F(N)
2320 D=V(N)
2330 Y=Y(N)
2340 Y1=Y-( INT(Y/100)*100)
2350 REM: SETUP OUTPUT FORMAT
2360 PRINT
2370 IF WW$="YES" THEN LPRINT""
2380 PRINT
2390 IF N=1 THEN CLS
2400 IF WW$="YES" THEN LPRINT""
2410 PRINT"POSITION OF THE MOON ON ";M;"/";D;"/";Y;" GMT FROM"" "
     W$
2420 IF WW$="YES" THEN LPRINT"POSITION OF THE MOON ON ";M;"/";D;"/";Y;"
     GMT FROM"" " W$
2430 PRINT
2440 IF WW$="YES" THEN LPRINT""
2450 PRINT"GMT" TAB(8)"GHA" TAB(17)"DEC" TAB(27)"LST" TAB(37)TL$
     TAB(47)"AZ" TAB(56)"EL"
2460 IF WW$="YES" THEN LPRINT"GMT" TAB(8)"GHA" TAB(17)"DEC" TAB(27)"LST"
     TAB(37)TL$ TAB(47)"AZ" TAB(56)"EL"
2470 I1=2
2480 REM: CALCULATION OF JULIAN DATE
2490 IF M>=3 THEN 2570
2500 IF INT((Y-1853)/4)<11 THEN 2530
2510 C1=-1
2520 GOTO 2540
2530 C1=0
2540 J1=365*(Y-1853)+D+30*(M+9)+INT((M+10)/2)
2550 J2=INT((Y-1853)/4)+1+C1
2560 GOTO 2680
2570 IF INT((Y-1852)/4)<11 THEN 2600
2580 C1=-1
2590 GOTO 2610
```

(continued on page 44)

fig. 3, continued

```
2600 C1=0
2610 IF M=9 THEN 2650
2620 IF M=11 THEN 2650
2630 C2=0
2640 GOTO 2660
2650 C2=1
2660 J1=365*(Y-1852)+D+30*(M-3)+INT((M-2)/2)
2670 J2=INT((Y-1852)/4)+C1+C2
2680 J=J1+J2
2690 JD#=J+2397547.5
2700 GOSUB 4450        'FOR MOON DISTANCE CALCULATIONS
2710 T1=J-17472.5
2720 REM: MAIN CALCULATIONS BEGIN
2730 D9=(B-INT(B/100)*100)+INT(B/100)*60
2740 D6=(E1-INT(E1/100)*100)+INT(E1/100)*60
2750 D7=D9-D6
2760 D8=D7-I
2770 IF D7>0 THEN 2790
2780 GOTO 2820
2790 IF D8>=0 THEN 4250
2800 B=E1
2810 REM: CALCULATION OF LATITUDE AND LONGITUDE OF MOON
2820 T=(B-INT(B/100)*100)/1440+INT(B/100)/24
2830 T5=T1+T
2840 K1=((.751213+.036601102*T5)-INT(.751213+.036601102*T5))* P5
2850 K2=((.822513+.0362916457*T5)-INT(.822513+.0362916457*T5))* P5
2860 K3=((.995766+.00273777852*T5)-INT(.995766+.00273777852*T5))
     *P5
2870 K4=((.974271+.0338631922*T5)-INT(.974271+.0338631922*T5))* P5
2880 K5=((.0312525+.0367481957*T5)-INT(.0312525+.0367481957*T5))
     *P5
2890 L8=K1+.658*R5*SIN(2*K4)+6.289*R5*SIN(K2)
2900 L8=L8-1.274*R5*SIN(K2-2*K4)-.186*R5*SIN(K3)
2910 L8=L8+.214*R5*SIN(2*K2)-.114*R5*SIN(2*K5)
2920 L8=L8-.059*R5*SIN(2*K2-2*K4)-.057*R5*SIN(K2+K3-2*K4)
2930 K6=K5+.6593*R5*SIN(2*K4)+6.2303*R5*SIN(K2)-1.272*R5* SIN(K2-2*K4)
2940 L7=5.144*R5*SIN(K6)-.146*R5*SIN(K5-2*K4)
2950 REM: CALCULATION OF RIGHT ASCENSION (R1) AND DECLINATION (D1)
2960 D1=COS(L7)*SIN(L8)*.397821+SIN(L7)*.917463
2970 D1=ATN(D1/( SQR(1-D1[2)))
2980 G1=50+.5+((D1)/(.792))*D5
2990 G2=80+((D1)/(.808))*D5
3000 G3=141.5-((D1)*(.738)*D5)
3010 G4=170.5-((D1)*(.857)*D5)
3020 A2=COS(L7)*COS(L8)/COS(D1)
3030 A1=( COS(L7)*SIN(L8)*.917463-SIN(L7)*.397821)/COS(D1)
3040 A=ATN(A1/A2)
3050 GOSUB 3390
3060 R1=A
3070 R2=R1*57.295779*24/360
3080 L1=.065709822*T1
3090 L=T*24*1.002738+6.646055+(L1-INT(L1/24)*24)
3100 LA=L-(L6*24*57.295779/360)*1.002738
3110 L=(L-INT(L/24)*24)
3120 REM: CALCULATION OF GREENWICH HOUR ANGLE (G) FROM LOCAL SIDEREAL
     TIME
3130 G=(L/24)*P5-R1
3140 IF G<P5 THEN 3170
3150 G=G-P5
3160 GOTO 3210
3170 IF G<0 THEN 3190
3180 GOTO 3210
3190 G=G+P5
3200 REM: CALCULATION OF LOCAL HOUR ANGLE (H) FROM GHA (G)
3210 H=L6-G
3220 GOSUB 4740        FOR PARALLAX CORRECTIONS
3230 REM: CALCULATION OF ELEVATION (E) OF OBJECT
3240 E3=COS(L5)*COS(H)*COS(D1)+SIN(D1)*SIN(L5)
3250 E2=SQR(1-(E3*E3))
3260 E=ATN(E3/E2)
3270 F=E
3280 IF E<0 THEN 4200
3290 IF E>I6*R5 THEN 4200
3300 REM: CALCULATION OF AZIMUTH (A) OF OBJECT
3310 A2=SIN(D1)/( COS(L5)*COS(F))
3320 A2=A2-( SIN(L5)/COS(L5))*( SIN(F)/COS(F))
3330 A1=SIN(L5)*SIN(D1)+COS(L5)*COS(D1)*COS(H)
3340 A1=( SIN(H)*COS(D1))/SQR(1-A1[2)
3350 A=ATN(A1/A2)
3360 GOSUB 3390
3370 GOTO 3540
3380 REM: REMOVAL OF AMBIGUITIES INCURRED WITH ATN FUNCTION
3390 IF A=0 THEN 3410
3400 GOTO 3450
3410 IF A2<0 THEN 3430
3420 GOTO 3530
3430 A=P5/2
3440 GOTO 3530
3450 IF A>0 THEN 3510
```

(continued on page 45)

fig. 3, continued

```
3460 IF A2<0 THEN 3490
3470 A=P5+A
3480 GOTO 3530
3490 A=P5+(A-P5/2)
3500 GOTO 3530
3510 IF A2=>0 THEN 3530
3520 A=A+P5/2
3530 RETURN
3540 IF(T-I1)>(2*I)/1440 THEN 3560
3550 GOTO 3580
3560 PRINT
3570 IF WW$="YES" THEN LPRINT""
3580 Z1=INT(A*D5*10+.5)/10
3590 Z2=INT(E*D5*10+.5)/10
3600 Z3=INT(G*D5*10+.5)/10
3610 Z4=INT(D1*D5*10+.5)/10
3620 IF Z4<0 THEN 3750
3630 IF Z3<G1 THEN 3750
3640 IF Z3>G2 THEN 3660
3650 GOTO 3690
3660 IF Z3<G3 THEN 3710
3670 IF Z3>G4 THEN 3750
3680 GOTO 3730
3690 Y$="U"
3700 GOTO 3760
3710 Y$="W"
3720 GOTO 3760
3730 Y$="J"
3740 GOTO 3760
3750 Y$=" "
3760 AT$="000"
3770 BT$="00"
3780 CT$="0"
3790 BS=INT(B+.5)
3800 IF BS<10 THEN BS$=AT$+RIGHT$(STR$(BS), 1): GOTO 3840
3810 IF BS<100 THEN BS$=BT$+RIGHT$(STR$(BS), 2): GOTO 3840
3820 IF BS<1000 THEN BS$=CT$+RIGHT$(STR$(BS), 3): GOTO 3840
3830 BS$=RIGHT$(STR$(BS), 4)
3840 IF TD>0 OR TD/100=INT(TD/100) TC=TD: GOTO 3860
3850 TC=TD+2360
3860 ES=BS+TC
3870 IF ES>2400 THEN ES=ES-2400
3880 IF ES<=0 THEN ES=ES+2400
3890 IF ES<10 THEN ES$=AT$+RIGHT$(STR$(ES), 1): GOTO 3930
3900 IF ES<100 THEN ES$=BT$+RIGHT$(STR$(ES), 2): GOTO 3930
3910 IF ES<1000 THEN ES$=CT$+RIGHT$(STR$(ES), 3): GOTO 3930
3920 ES$=RIGHT$(STR$(ES), 4)
3930 IF LA<0 THEN LA=LA+24
3940 IF LA>24 THEN LA=LA-24
3950 LB=100*INT(LA)
3960 LC=60*(LA-INT(LA))
3970 IF LC-INT(LC)=>0.5 LC=INT(LC)+1 ELSE LC=INT(LC)
3980 IF LC=>60 LC=0: LB=LB+100
3990 LD=LB+LC
4000 IF LD>2400 THEN LD=LD-2400
4010 LB$=STR$(LD)
4020 IF LD<10 THEN LB$=AT$+RIGHT$(LB$, 1): GOTO 4060
4030 IF LD<100 THEN LB$=BT$+RIGHT$(LB$, 2): GOTO 4060
4040 IF LD<1000 THEN LB$=CT$+RIGHT$(LB$, 3): GOTO 4060
4050 LB$=RIGHT$(LB$, 4)
4060 Z1$="###.#"
4070 Z2$="##.#"
4080 Z3$="###.#"
4090 Z4$="+##.#"
4100 PRINT USING"%  %";BS$;
4110 PRINT TAB(7)USING Z3$; Z3;
4120 PRINT TAB(16)USING Z4$; Z4;
4130 PRINT Y$;
4140 PRINT TAB(27)LB$;
4150 PRINT TAB(37)USING"%  %";ES$;
4160 PRINT TAB(45)USING Z1$; Z1;
4170 PRINT TAB(55)USING Z2$; Z2
4180 IF WW$="YES" THEN LPRINT USING"%  %";BS$;: LPRINT TAB(7)USING
     Z3$; Z3;: LPRINT TAB(16)USING Z4$; Z4;: LPRINT Y$;: LPRINT TAB(27)LB$;
     : LPRINT TAB(37)USING"%  %";ES$;: LPRINT TAB(45)USING Z1$; Z1;:
     LPRINT TAB(55)USING Z2$; Z2
4190 I1=T
4200 B=B+I
4210 Z=(B-INT(B/100)*100)-60
4220 IF Z<0 THEN 2730
4230 B=INT(B/100)*100+100+Z
4240 GOTO 2730
4250 RX=R2
4260 IF RX<0 THEN RX=RX+24
4270 IF RX>24 THEN RX=RX-24
4280 RA=100*INT(RX)
4290 RB=60*(RX-INT(RX))
4300 IF RB-INT(RB)=>0.5 RB=INT(RB)+1 ELSE RB=INT(RB)
4310 IF RB=60 RB=0: RA=RA+100
```

(continued on page 47)

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TRS-80 to select days and months for 1978 and 1983. Why did I select those particular years? 1983 was the current year; 1978 is 5 years before. In addition, copies of the *Nautical Almanac* for each of those years were easily available. I printed out the moon coordinates to three decimal places for the selected dates. Then I modified an existing statistical program so that I could efficiently use the TRS-80 to calculate the average, or arithmetical means, standard deviations, and ranges of errors.

Average errors of less than 0.01 degree were obtained with maximum errors under 0.2 degree in sample sizes approaching 100. This inherent accuracy of the basic program equations should be adequate for many Amateur EME stations.

The above analysis is for geocentric GHA and declination data with the *Nautical Almanac* used as an accurate standard for comparison. I could find no comparable standard for azimuth and elevation error analysis. However, at least one checkpoint each day can be used to gain some confidence in the AZ/EL calculations. When the moon's GHA equals your longitude, the azimuth should be 180 degrees. At that same time the elevation should be equal to 90 degrees minus your latitude (+north/−south) plus the declination (+north/−south). Checking these points for the days used in the preceding analysis, it was found that the azimuth differences at 180 degrees were less than 0.011 degrees and elevation under 0.001 degrees, maximum. Average errors were 0.00037 and 0.00013, respectively. (These are for geocentric values.)

Having gained confidence in the ability of the fundamental equations to provide accurate results, the parallax corrections which had been bypassed for the above analyses were reinstated to provide output values referred to the surface rather than the center of the earth. These are given in lines 4750 through 4850 and were based on information found in reference 3. Again, the lack of an accurate standard precludes a statistical error analysis but a straight comparison of GHA, declination, azimuth and elevation values before and after the parallax corrections may be useful.

When this was done, all corrections were less than 1.0 degree. Those for azimuth were "in the noise level," i.e., less than 0.005 degree. At moonset and moonrise, the elevation correction factors were very close to 1 degree and at azimuths near 180-degrees about 0.6 degree. (A detailed error analysis is available from *ham radio*, Greenville, NH 03048. Send an SASE with two first-class stamps attached. Request W2WD Moon Coordinate Error Analysis.)

optimizing the program

The program as presented is arranged for universal use. For frequent use at one location, it should be streamlined to eliminate the time-consuming task of keying in the repetitive portions of the input data. To do this, the following changes should be made:

- At **line 1450**, remove the apostrophe ('). This has the effect of activating this GOTO 1730, which bypasses the repetitive input data requirements.
- At **line 1730**, remove the apostrophe (') and activate this line, replacing my standard input data with your standard input data:

W$	=	"your call letters" (use quotation marks)
L5	=	your latitude, degrees [negative (−) for south latitude]
U5	=	your latitude, minutes
L6	=	your longitude, degrees [negative (−) for east longitude]
U6	=	your longitude, minutes
TL$	=	"your time zone" (use quotation marks)
TD	=	time differential from GMT to your zone [negative (−) for zones west of GMT]

When these changes have been made, the program, when called up, will immediately go to the question WHAT IS THE DESIRED PRINTING INCREMENT IN MINUTES (1-60)? Thus, you have avoided the logo, the need to input your call letters, local time, time differential, latitude, longitude. This saves a lot of data entry time. I use two tailored programs . . . one set for Eastern Standard Time, for which I use the filespec "MOONEST" and the other "MOONEDT" for Eastern Daylight Savings Time. If you never use the low-elevation mode, this section can be bypassed. If you do not have a printer, then that question can likewise be skipped. With some knowledge of BASIC programming, making changes to suit your individual requirements are not difficult.

The program was rewritten for the level II dialect of the Model I TRS-80 from versions presented in reference 4 by Lance Collister (WA1JXN) and Jay Liebmann (K5JL). Enhancements have been added in the form of readouts for Local Sidereal Time, Right-Ascension of the moon and path loss variations. Provisions have also been made for the direct inputting of time zones and time differentials. Keyboard input statements have been error-trapped to reduce the chance for "cockpit" errors.

For those wishing to avoid the task of manually keying in the 10,000-byte program listing (see **fig. 3**), a 500-baud, level II BASIC cassette tape is available from the author for $15.00 including postage. For Disk BASIC the taped program may be "cloaded" into memory after first shutting off the interrupts with the CMD"T" command. From memory it may then be "saved" onto a disk for future use. Two versions are on the tape. The first is the version as presented in this article. It was stretched out to make it easier to list and modify to your own requirements. The second

fig. 3, continued

```
4320 RC=RA+RB
4330 RY$=STR$(RC)
4340 IF RC<10 THEN RY$=AT$+RIGHT$(RY$, 1)
4350 IF RC<100 THEN RY$=BT$+RIGHT$(RY$, 2)
4360 IF RC<1000 THEN RY$=CT$+RIGHT$(RY$, 3)
4370 RY$=RIGHT$(RY$, 4)
4380 PRINT
4390 IF WW$="YES" THEN LPRINT""
4400 PRINT"R.A. OF MOON = ";RY$; "      PATH-LOSS INCREASE  +";DB
     " DB"
4410 IF WW$="YES" THEN LPRINT "R.A. OF MOON = ";RY$;"      PATH-LOSS
     INCREASE  +";DB;" DB"
4420 PRINT
4430 NEXT N
4440 END
4450 REM: CALCULATE DISTANCE TO THE MOON
4460 DD#=JD#-2444238.5
4470 AA=0.98564733
4480 ED=-3.76286
4490 MS=(AA*DD#)+ED
4500 IF MS<0 THEN MS=MS+360: GOTO 4500
4510 IF MS>360 THEN MS=MS-360: GOTO 4510
4520 AE=0.1858*SIN(MS*0.0174533)
4530 AF=0.37*SIN(MS*0.0174533)
4540 LS=(AA*DD#)+(1.9157417*SIN(0.0174533*((AA*DD#)+(-3.76286))))
     +278.833540
4550 IF LS<0 THEN LS=LS+360: GOTO 4550
4560 IF LS>360 THEN LS=LS-360: GOTO 4560
4570 LL=(13.1763966*DD#) + 64.975464
4580 IF LL <0 THEN LL=LL+360: GOTO 4580
4590 IF LL >360 THEN LL=LL-360: GOTO 4590
4600 CC=LL-LS
4610 MM=LL-(0.1114041*DD#)-349.383063
4620 IF MM<0 THEN MM=MM+360: GOTO 4620
4630 IF MM >360 THEN MM=MM-360: GOTO 4630
4640 EV=1.2739*SIN(((2*CC)-(MM))*0.0174533)
4650 MN=MM+EV-AE-AF
4660 EC=6.2886*SIN(MN*0.0174533)
4670 MD=383242.41/(1+0.054900*COS((MN+EC)*0.017453292))
4680 REM: CONVERT DISTANCE VARIATION TO PATH-LOSS CHANGE (DB)
4690 DB=MD/356334
4700 DB=40*LOG(DB)/LOG(10)
4710 DB=INT(DB*10+0.5)/10
4720 RETURN
4730 END
4740 REM: CORRECTIONS FOR PARALLAX
4750 H1=H
4760 R#=MD/6378.16
4770 U=ATN(0.996647*TAN(L5))
4780 P1=0.996647*SIN(U)
4790 P2=COS(U)
4800 HC=ATN((P2*SIN(H))/(R#*COS(D1)-P2*COS(H)))
4810 H=H+HC
4820 G=L6-H
4830 D1=ATN((COS(H))*((R#*SIN(D1))-P1)/((R#*COS(D1)*COS(H1))-P2))
4840 RETURN
```

is a customized version which will be set to your station parameters if you include the necessary information. I will need your station call letters, your latitude, your longitude, your local time zone, and the difference between your time and GMT. Also include whether you want the low-elevation and printing routines bypassed.

The two programs are also available on a diskette ($18.00) formatted to run on TRSDOS compatible operating systems. It has been checked on single-density TRSDOS 2.3, NEWDOS 2.1, NEWDOS-80 1.0, DOSPLUS 3.3 and LDOS 5.0. A double-density version will run on NEWDOS-80 2.0 or DBLDOS 4.23 and perhaps others using a Percom doubler with a Model I machine. Be sure to specify single or double-density and include the same information as specified in the paragraph above.

references

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ham radio

WHAT THE COMPETITION CALLS "NO LOST WORDS OR SYLLABLES" WE CALL NOISE

THE COMPETITION: "HI HONEY I(NOISE BURST)M FIXING YOUR (NOISE BURST)VORITE SUPPE(NOISE BURST) BARBECUE HAMB(NOISE BURST)GERS. THEY WI(NOISE BURST) BE READY AT 6:(NOISE BURST) WILL YOU MA(NOISE BURST) IT IN TIME?(NOISE BURST)"

Samples (heard as bursts of noise) displace your phone party's audio for as long as it takes your transceiver to T/R. The above example assumes a transceiver T/R time of about 150 mS (typical)

PRIVATE PATCH II: " HI HONEY I AM FIXING YOUR FAVORITE SUPPER. . . BARBECUE HAMBURGERS. THEY WILL BE READY AT 6:30. WILL YOU MAKE IT IN TIME?"

Thousands of PRIVATE PATCH II owners are enjoying the commercial communications quality that only a VOX based, simplex autopatch can deliver.

PRIVATE PATCH II IS PRICED AT LESS THAN HALF OF OUR COMPETITORS "FAVORITE COMMERCIAL SIMPLEX VOX PATCH"

SAMPLING VS. PRIVATE PATCH II

The performance of a sampling patch is totally dependent on the T/R speed of your radio. Such is not the case with PRIVATE PATCH II. PRIVATE PATCH II will give excellent results with any radio. Synthesized and relay switched types included.

PRIVATE PATCH II requires only three connections to your radio (MIC, PTT and SPKR audio). If these connections are made inside your radio PRIVATE PATCH II does not interfere with the normal use of your radio. Otherwise for a quick and easy interface you may plug PRIVATE PATCH II Into the MIC and SPKR jacks. A 10 minute job! Unlike sampling patches, connections are not required to the squelch, discriminator or power. And best of all, modifications are never required.

Controlling and talking through PRIVATE PATCH II is almost always quicker and easier than using a sampling patch. This is because you may talk or send control commands the instant you press the PTT button. The ability to break in or take control is assured by interrupt control logic. The interrupt controller creates a window (similar to sampling) but is seldom heard in normal quick back and forth communication. With a sampling patch you press the MIC button for one to five seconds before talking on each and every transmission. This is very frustrating for the mobile, and causes confusion for each party.

The sampling process reduces the effective range of your base radio. This is because if a sample, and a signal fade coincide, the sampling patch thinks the mobile is not transmitting. This causes a sampling patch to become erratic at ranges still very useable by PRIVATE PATCH II. PRIVATE PATCH II will not diminish the range of your system.

PRIVATE PATCH II has two more range extending tricks not available to a noise sampled autopatch.

1. You may use a linear amplifier with no loss of performance
2. You can operate through any repeater from your base station.

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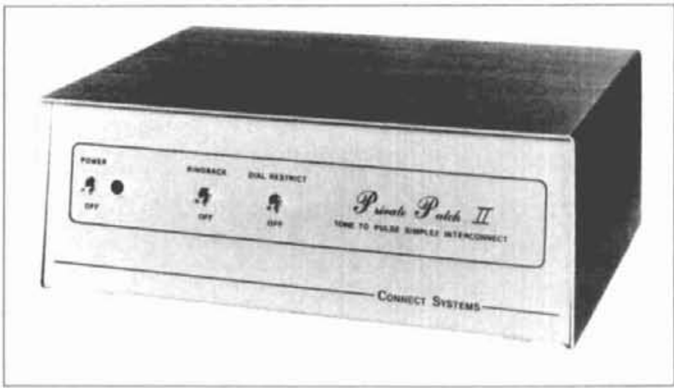

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More Details? CHECK—OFF Page 128

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Use this device
to determine Yagi
or other antenna patterns

a sensitive field strength meter

While attempting to make some meaningful relative field strength measurements on an antenna system some months ago, I found my years-old field strength meter to be quite inadequate. Its lack of sensitivity meant that measurements had to be made close to the antenna, where the induction field introduced error in the radiation field I was actually interested in. Its non-linearity was also a problem.

I decided to design a new unit that would include a stable, sensitive linear amplifier, provisions for remote monitoring, and a meter calibrated in decibels. The final results proved so satisfactory that it seemed other Amateurs would be interested in this project.

Figure 1 shows the schematic. The two-pole, five-position switch, coils and 365 pF variable capacitor cover a range from 1.5 to 30 MHz. (The combination of parts came from an old low-power transmitter.)

Almost all handbooks show coil-capacitor combinations that can be used for field strength meters, so Amateurs can build the kind of unit that best suits their needs. In addition, they can incorporate any kind of pickup coupling system for the pickup antenna of their choice. My own measurements were sufficiently satisfactory using a short antenna only a few feet long connected directly to the ANT binding post. When a longer antenna for pickup was desired, it was connected to CAP, binding post C, where it connected to C2, a small 50 pF variable capacitor.

The amplifier uses a couple of Darlington NPN transistors whose high beta, 5000, provides high sensitivity with S1 used as the amplifier ON/OFF switch. Switch S2 in the left position allows the output of the 1N34 diode to be fed directly into the 50 μA meter (M) for direct reading. When S2 is in the right position, the amplifier is switched into the circuit. Switch S3 is for LOCAL or REMOTE monitoring. At full GAIN setting the input signal is adjusted to give a full-scale reading of fifty microamperes on the meter. Then with the amplifier switched out of the circuit, the meter reading drops down to about half a microampere.

A 2.5 mH RF choke and capacitors C3, C4, and C5 effectively keep RF out of the amplifier circuit.

two balance adjustments are required

Because of the high sensitivity of the amplifier circuit, it's best to have two balancing controls, R7 (BAL, front panel adjustable) and R8 (an internal screwdriver-adjustable potentiometer). When initially setting R8, adjust it in conjunction with R7. Doing this allows the GAIN control to be varied from zero to full gain with the meter indication remaining at zero.

Because the amplifier circuit is basically a balanced bridge type, a zero-center meter could have been used. But none was available, so R3, an 82-ohm offset resistor, was added. Its function is to prevent the meter from swinging too hard to the left below zero, which might occur if the panel balance control were inadvertently turned fully counterclockwise with the amplifier gain turned fully on. The best safe operating procedure to employ is always to turn the gain con-

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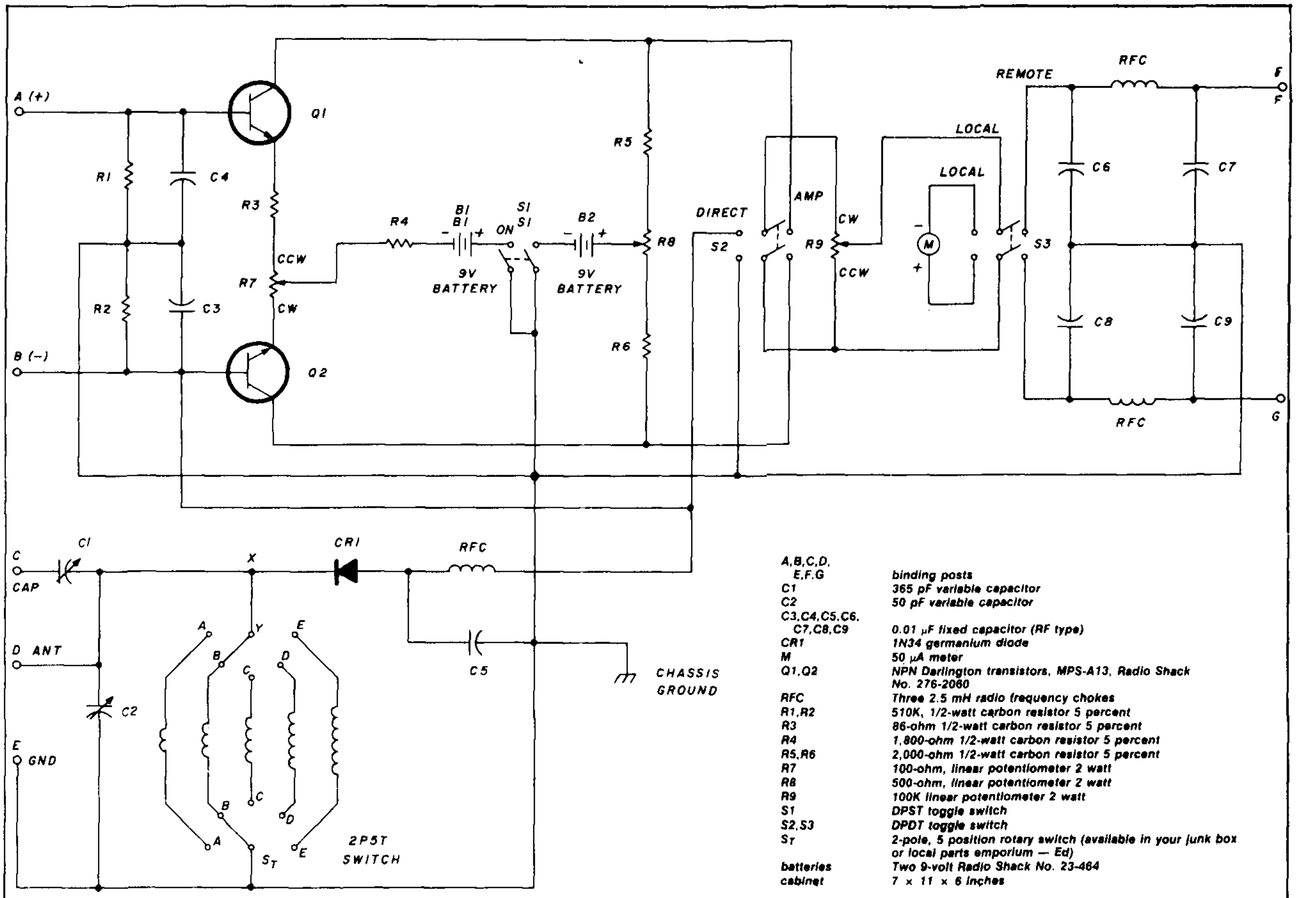


A,B,C,D, E,F,G	binding posts
C1	365 pF variable capacitor
C2	50 pF variable capacitor
C3,C4,C5,C6, C7,C8,C9	0.01 μF fixed capacitor (RF type)
CR1	1N34 germanium diode
M	50 μA meter
Q1,Q2	NPN Darlington transistors, MPS-A13, Radio Shack No. 276-2060
RFC	Three 2.5 mH radio frequency chokes
R1,R2	510K, 1/2-watt carbon resistor 5 percent
R3	86-ohm 1/2-watt carbon resistor 5 percent
R4	1,800-ohm 1/2-watt carbon resistor 5 percent
R5,R6	2,000-ohm 1/2-watt carbon resistor 5 percent
R7	100-ohm, linear potentiometer 2 watt
R8	500-ohm, linear potentiometer 2 watt
R9	100K linear potentiometer 2 watt
S1	DPST toggle switch
S2,S3	DPDT toggle switch
S_T	2-pole, 5 position rotary switch (available in your junk box or local parts emporium — Ed)
batteries	Two 9-volt Radio Shack No. 23-464
cabinet	7 × 11 × 6 inches

fig. 1. Field strength meter schematic.

trol down to zero before turning the power switch ON, and then initially balance the circuit while slowly advancing the gain control. This is a common precaution used in all sensitive bridge circuits.

additional features

The addition of binding posts A and B allows the amplifier to be used by itself as a high impedance, high gain amplifier, where binding posts A and E are used for a + input and ground.

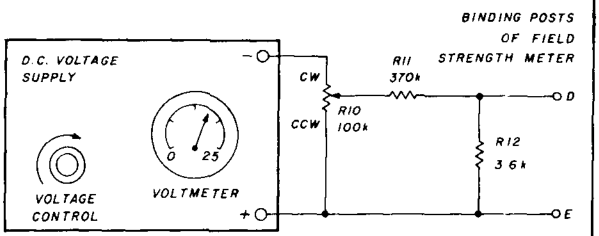
Switch S3, a remote metering circuit in my original field strength meter, was also incorporated in the new, more sensitive unit. However, it was found necessary to add the RF filter circuit to help eliminate RF pickup from the remote metering wires to the unit. This remote metering feature is not really necessary, but is handy if you want to elevate the field strength meter a few feet when making beam antenna measurements. Care should be taken that the remote meter wires are both shielded and properly grounded, because it doesn't take much RF getting back into the field strength meter to upset the initial balance.

Although the meter readings by themselves are a good indication of relative field strength, it was thought desirable to provide a theoretical calibration curve relating meter deflection to dB change, and then to do an actual calibration using the test circuit shown in **fig. 2**.

calibrating is easy

The theoretical dB calibration curve can be best understood if, for example, we use full scale on the meter 50 μA as 0 dB with a given input signal. Now if the input signal were reduced so that the meter reads 30 μA, then the dB drop in signal level is $dB = 20 \log_{10} (30/50) = 4.44$ dB. This calculation is for a perfect linear system; and by using this mathematical procedure it is easy to calculate and develop the theoretical dB curve of **fig.3** for different values of meter current.

This curve is very easy to use. For example, with the field strength meter set up away from your station, if you rotated your beam and saw that the maximum meter reading went over 50μA, just cut back on the GAIN control on the field strength meter, so that the maximum reading is either 50 μA or somewhat lower. The maximum reading does not have to be exactly 50 μA, as shown by the following example.



item	description
PS	0-25 volt DC power supply (see text)
R10	100 kilohm 2-watt potentiometer
R11	370 kilohm, 1/2-watt carbon resistor
R12	3.6 kilohm, 1/2-watt carbon resistor

fig. 2. Calibration test unit.

If the maximum reading were 45 μA and the minimum reading 10 μA, then from **fig. 3**, 45 μA is equal to −0.915 dB, and 10 μA is equal to −13.98 dB and the numerical difference is −13.065 dB, which can be rounded off to −13.1 dB.

The reason that it is not necessary to have the microammeter set exactly at 50 μA when reading the maximum signal is that decibels can be added or subtracted from each other anywhere along the curve to give the correct dB difference.

It should be pointed out, however, that better accuracy is obtained if the maximum reading is set initially close to 50 μA. This is because, as in most metering systems, the percentage accuracy obtainable is less at the lower end of a scale than at the top. Also, the curve shown was not extended below 5 mA.

I happened to have a small variable 0-25 VDC power supply with a reasonably accurate built-in voltmeter. Almost any kind of variable supply with a voltmeter can be used, as the power supply voltage is factored into the calibration procedure to be described. The attenuator network is used to reduce the power supply voltage so that full scale on the microammeter could readily be obtained at full gain of the amplifier, with the power supply set at 25 volts.

Before beginning the calibration procedure, it is first necessary to isolate the resonant circuits from the input terminals; otherwise the coils would provide a DC short between the RF input binding posts D and E used for the calibration input voltage. This can be accomplished by temporarily opening the circuit between points X and Y of **fig. 1**. An alternate method is just to slip a piece of paper between the coil contact on the turret and the arm of the rotary switch.

The test unit is then connected to the input binding posts D and E, as shown, and R10 rotated to the counterclockwise, or zero-output position. The amplifier is now turned on, properly balanced to zero, and the GAIN control set to its maximum position. The balance should again be checked, and it should still be at zero. The power supply is turned on, and the voltage adjusted to read 25 volts. R10 is rotated in a clockwise direction until the field strength microammeter indicates a full scale reading of 50 μA. R10 should not be adjusted further, since its setting now has established the zero dB level.

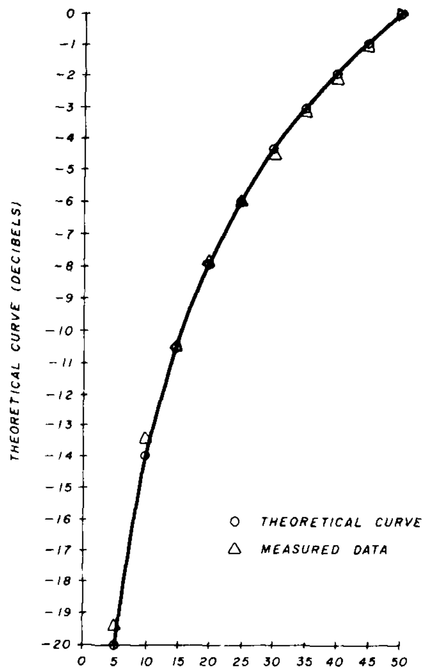
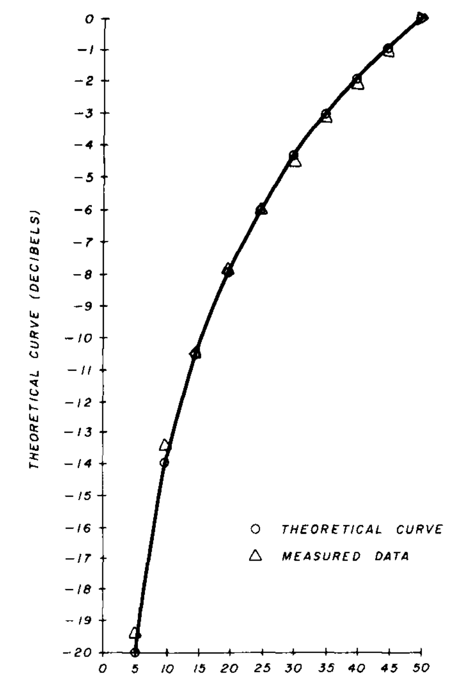


fig. 3. Measured data showing close correlation to theoretical values indicates good amplifier linearity.

The power supply voltage is now reduced until the microammeter reads 45 μA. The power supply voltage is now read and recorded. This procedure is again repeated for 40 μA, again reading and recording the power supply voltage. Similar successive steps, each 5 μA lower are done, until the final microampere reading of 5 μA is made and data recorded. The actual recorded data is indicated in **fig. 3**.

The actual calculated dB for a measured reading of voltage is done as follows: for example, at 30 μA, the power supply voltage was found to be 15.2 volts. The actual calculated dB for a measured reading of, for example, 30 μA, and a recorded power supply voltage of 15.2 volts is equal to $20 \log_{10} (15.2/25) = -4.32$ dB. This is quite close to the theoretical value of −4.44 dB previously shown. The resulting data and curve is shown in **figs. 3** and **4**. Good correlation between theoretical and observed data shows that the amplifier had a good linear characteristic. Although the amplifier was calibrated using DC voltage, the results obtained showed the unit to be very useful as a device for measuring relative dB levels. And that is what most Amateurs are really interested in. The measure of absolute field strength normally requires extremely good



MICROAMPERES	THEORETICAL dB	TEST POWER SUPPLY VOLTAGE	CALCULATED MEASURED dB
50	0	25	0
45	-0.915	22.4	-0.954
40	-1.94	19.7	-2.07
35	-3.10	17.6	-3.05
30	-4.44	15.2	-4.32
25	-6.02	12.5	-6.02
20	-7.96	10.1	-7.87
15	-10.46	7.6	-10.34
10	-13.98	5.3	-13.47
5	-20	2.6	-19.66
0	∝	——	——

fig. 4. Measured and theoretical data.

commercial equipment and is not generally used in Amateur-type measurements.

Although the calibration curve was shown for the full maximum gain of the amplifier, several similar curves were run at lower gains, down to 20 percent of maximum gain, and no noticeable variations were encountered. This indicates that the curve shown can be used between the limits of full gain down to 20 percent of full gain. My own experience is that if your signal is very strong, just reduce the size of your pickup antenna or move your field strength meter further away from your transmitting antenna. Actually, the further you are away from your transmitting antenna, the better patterns you will obtain.

field testing

After experimentation at my own station, the unit was taken to the home of Russell Forsyth, K4YS, who had recently put up a large 20-meter beam. The azimuthal pattern in dB was readily obtained, with a front-to-back ratio measured at 14 dB, corresponding to a power ratio of a little over 25:1, which seemed reasonable for the beam he used. A rather interesting anomaly was observed, in that at a certain direction when the beam was rotated, the field strength meter showed a marked irregular intermittent variation. Shaking the tower seemed to aggravate the condition, indicating that perhaps one of the elements was loose, or perhaps the mast was not perfectly vertical, and at certain positions, perhaps loose bolts or something else was causing the problem. At the moment, we still do not know what the problem is, but it does show that the field strength meter can also be used as a diagnostic tool in problems of this type.

Other tests near my home, where we have a high-voltage line nearby, showed that there was some sort of pickup near the power poles on rainy days, particularly when corona could be heard. The noise was tunable and verified as line noise interference by using a small transformer and earphones connected to the REMOTE binding posts instead of a meter. Apparently the lightning protection system, which incorporates a ground at each pole and a overhead ground line, was being shock-excited by the corona discharge and causing those loop circuits to have induced noise oscillations. (At least that's what it seemed to me.) Perhaps later I can use the meter to track down excessive line noise that occasionally is quite high on my Yaesu FT-101B.

There were no problems in the design or construction of the basic unit and its modifications. Though the field strength meter, as shown in the photograph, is quite large, it could certainly be reduced in size. I'm hoping that its use will provide Amateurs with a more sensitive method of making relative antenna measurements, so essential to good antenna experimentation. I'll be glad to answer any questions or comments you might have. Just send an SASE to the author at the address indicated.

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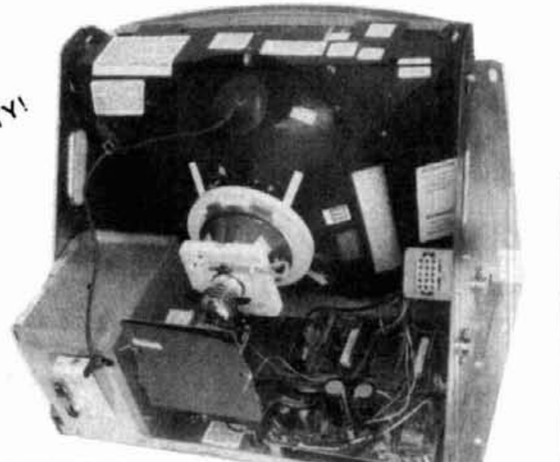

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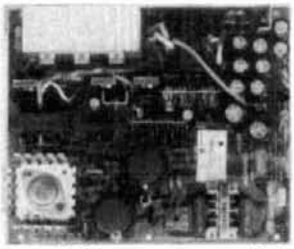

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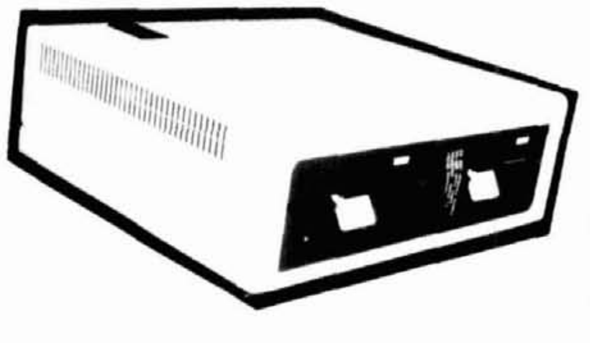

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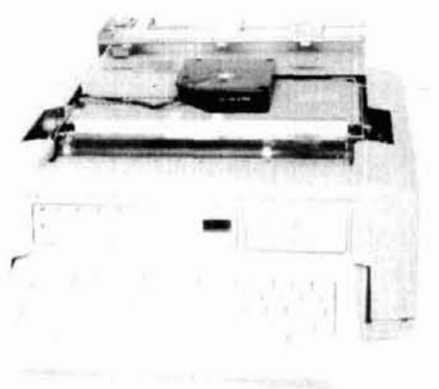

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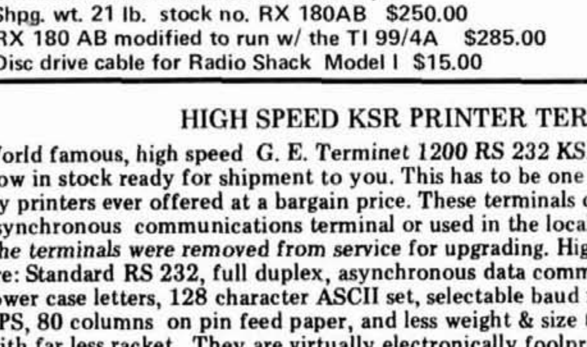

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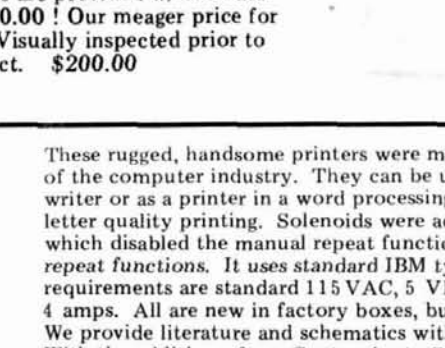

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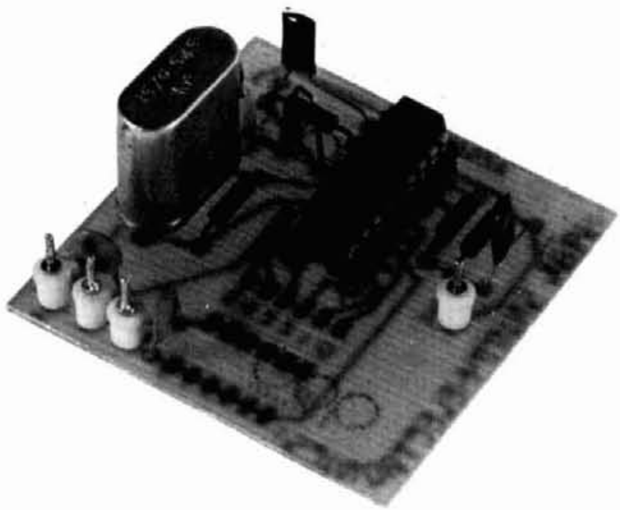

an integrated circuit low-pass filter

Operational amplifiers, developed and popularized in the late 1960s, have all but eliminated audio amplifiers designed using discrete transistors, resistors, and capacitors. Amplifiers built with discrete components are usually designed for special applications. In an analogous manner, audio frequency filters have evolved. First came the LC filters. In widespread use since before 1920, they are characterized by large physical size and weight, complexity of design procedure, and critical component values. Next came the active filters. Active filters were first implemented using vacuum tubes, then discrete transistors, and today using integrated circuits — typically quad op amps. These circuits did away with bulky inductors and complicated design procedures. Any adjustments needed on an active filter could be accomplished by "tweaking a pot" or two. As manufacturers of integrated circuits saw large numbers of op amps being used to manufacture the same type of active filters again and again, they saw the opportunity to profit from the manufacture of dedicated filtering devices. The use of one of these integrated circuit filters is the subject of this article.

The device discussed in this article is the S3528 programmable low-pass filter, manufactured by American Microcircuits, Inc., 3800 Homestead Road, Santa Clara, California 95051. Contained within one eighteen-pin dual inline package (DIP) is a complete seventh-order elliptical low-pass filter; its passband ripple is less than 0.1 dB, and its stopband attenuation is greater than 51 dB for frequencies greater than 1.3 f_c (cutoff frequency).

In addition, this IC contains two uncommitted operational amplifiers that can be used to provide additional filtering and gain. The device is also progammable for cutoff frequency. This means that the cutoff frequency is not fixed, but can be changed by providing switch or logic level inputs to the device. The S3528 also has a built-in oscillator for use with an external crystal,

By Robert L. Martin, WB2KTG, 45 Salem Lane, Little Silver, New Jersey 07739

Block Diagram **Pin Configuration**

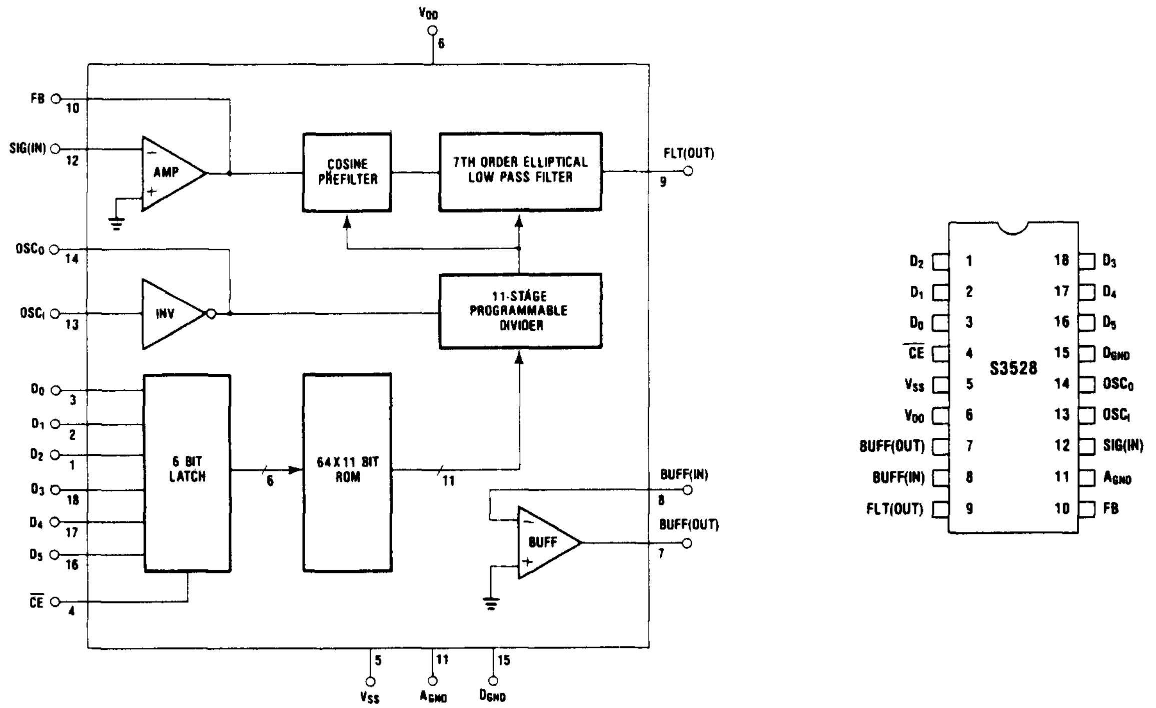

fig. 1. S3528 programmable low-pass filter. *(Courtesy American Microsystems, Inc., 1984.)*

such as the low-cost 3.58 MHz TV crystal. This feature allows very stable performance over a wide range of temperature and voltage. If desired, an external oscillator may be used. The frequency range of operation of this low-pass filter is 40 Hz to 20 kHz with the TV crystal, or 10 Hz to 20 kHz with an external oscillator. A block diagram and pin layout are shown in **fig. 1**.

The six data inputs, DØ-D5, are tied to either +5 VDC or GND to program the cutoff frequency into the device according to the coding shown in **table 1**. These data inputs, latched by an input buffer, are used to preset the internal divider to give the proper frequency to the switched capacitor filter.

In our demonstration circuit, a 6-position jumper "patch panel" provides the frequency selection. If you connect the DØ through D5 inputs to the buffered data bus of a microcomputer, the cutoff frequency can be varied by simply outputting the proper code per **table 1**.

switched capacitor filters

This integrated circuit filter uses CMOS op amps, CMOS transmission gates (switches), and CMOS capacitors to synthesize the required filtering function. The technique used is that of a switched capacitor filter. The following discussion will explain the functioning of a simple SCF circuit.

Figure 2 illustrates a simple RC active first order low-pass filter and **fig. 3** an SCF version of the same circuit. If the switch in **fig. 3** is in position "A," the voltage on capacitor C_r becomes equal to the input voltage. If the switch is then changed to position "B," the charge on C_r can be considered to have been transferred to capacitor C. The basic function illustrated being that of integration or low-pass filtering.

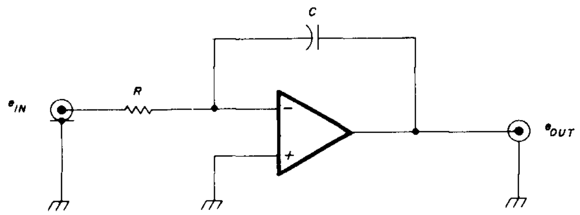

fig. 2. Simple RC-active low-pass filter.

The value of the simulated resistor (C_r and the SPDT switch) depends only on the capacitance of C_r and the switching frequency. Since the ratio of capacitor sizes is proportional to their surface areas in an integrated circuit, the ratio remains stable over a wide temperature and voltage range. If the switching frequency is crystal controlled, or otherwise stabilized, a very frequency-stable filter results.

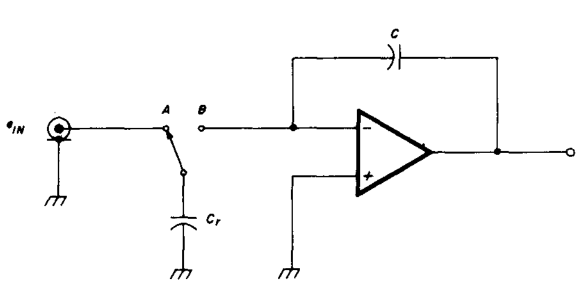

fig. 3. Switched capacitor equivalent to *fig. 2*.

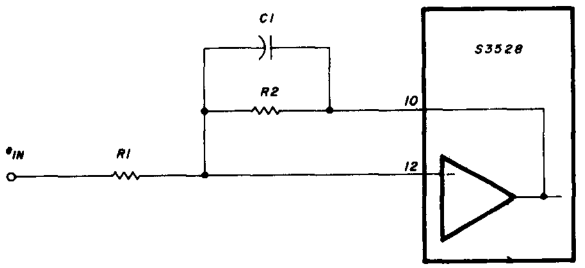

fig. 4. Input anti-alias low-pass filter.

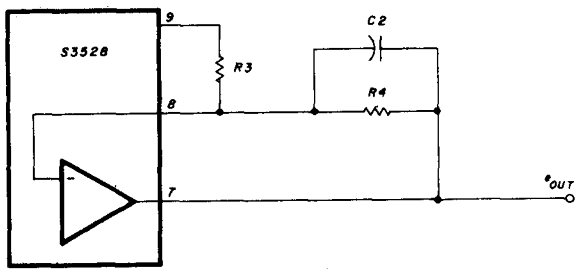

fig. 5. Output smoothing low-pass filter.

one problem with SCFs

The S3528 filter uses a switching frequency approximately 80 times the progammed cutoff frequency. As long as the input signals are far removed, no greater than 10 percent of the switching frequency, there are no problems. If, however, some portion of the input signal contains energy at the switching frequency, a phenomenon known as "aliasing" will occur. An alias is a spurious signal that appears at the output, indistinguishable from the desired low-frequency signal.

One way to visualize what is happening is to think of the stroboscopic effect observed in a Western movie when the frame flicker rate is approximately equal to the rotational speed of the spoked wagon or stage coach wheels: the wheels may appear to be stationary, or even going backward. This is an example of visual aliasing.

To prevent this from happening in our circuit, we apply a simple active RC low-pass filter to the input. We use the input op amp as shown in **fig. 4**.

To properly size R1, R2, and C1 the following method is used:

Determine the required amplification factor of the filter. For ease of application, limit gains to between 1 and 10. Assume R1 = 10 kilohms. This establishes the input impedance of this filter. R2 = n·R1 where n is the required gain.

Determine the desired cutoff frequency. Multiply this frequency by 2 and apply this to the following formula:

$$C = \frac{1}{6.28 \cdot f \cdot R2} \qquad (1)$$

For example:

Gain = 3.5

Cutoff frequency = 100 Hz

R2 = 35 kilohms

table 1. Frequency vs. input coding for S3528.

NOMINAL CUTOFF FREQ. (HZ)	INPUT CODE BINARY (HEX) D5-D0	DIVIDER RATIO	ACTUAL CUTOFF FREQ. (HZ)
40	000000 (00)	2048	44
100	000001 (01)	895	100
200	000010 (02)	447	200
250	001011 (0B)	358	250
300	000011 (03)	298	300
400	000100 (04)	224	399
475	001010 (0A)	188	476
500	000101 (05)	179	500
600	000110 (06)	149	601
700	000111 (07)	128	699
800	001000 (08)	112	799
900	001001 (09)	99	904
1000	001100 (0C)	90	994
1000	010000 (10)	89	1005
1030	001101 (0D)	87	1028
1050	001110 (0E)	85	1053
1100	010001 (11)	81	1105
1150	001111 (0F)	78	1149
1200	010010 (12)	74	1209
1300	010011 (13)	69	1297
1400	010100 (14)	64	1398
1470	011010 (1A)	61	1467
1500	010101 (15)	60	1491
1540	011011 (1B)	58	1542
1600	010110 (16)	56	1598
1700	010111 (17)	53	1688
1720	011100 (1C)	52	1721
1800	011000 (18)	50	1790
1900	011001 (19)	47	1904
1950	011101 (1D)	46	1945
2000	100000 (20)	45	1989
2030	011110 (1E)	44	2034
2100	100001 (21)	43	2081
2200	100010 (22)	41	2183
2240	011111 (1F)	40	2237
2300	100011 (23)	39	2295
2350	101010 (2A)	38	2350
2400	100100 (24)	37	2418
2500	100101 (25)	36	2686
2560	101011 (2B)	35	2557
2600	100110 (26)	34	2632
2700	100111 (27)	33	2711
2800	101000 (28)	32	2797
2900	101001 (29)	31	2887
3000	110000 (30)	30	2983
3100	110001 (31)	29	3086
3200	110010 (32)	28	3196
3300	110011 (33)	27	3314
3400	110100 (34)	26	3442
3600	110110 (36)	25	3579
3700	110111 (37)	24	3728
3900	111001 (39)	23	3891
4000	101100 (2C)	22	4067
4470	101101 (2D)	20	4474
5000	101110 (2E)	18	4971
5600	101111 (2F)	16	5593
6000	110101 (35)	15	5965
6400	111000 (38)	14	6392
7500	111010 (3A)	12	7457
9000	111011 (3B)	10	8949
10000	111100 (3C)	9	9943
15000	111101 (3D)	6	14915
18000	111110 (3E)	5	17897
22000	111111 (3F)	4	22372

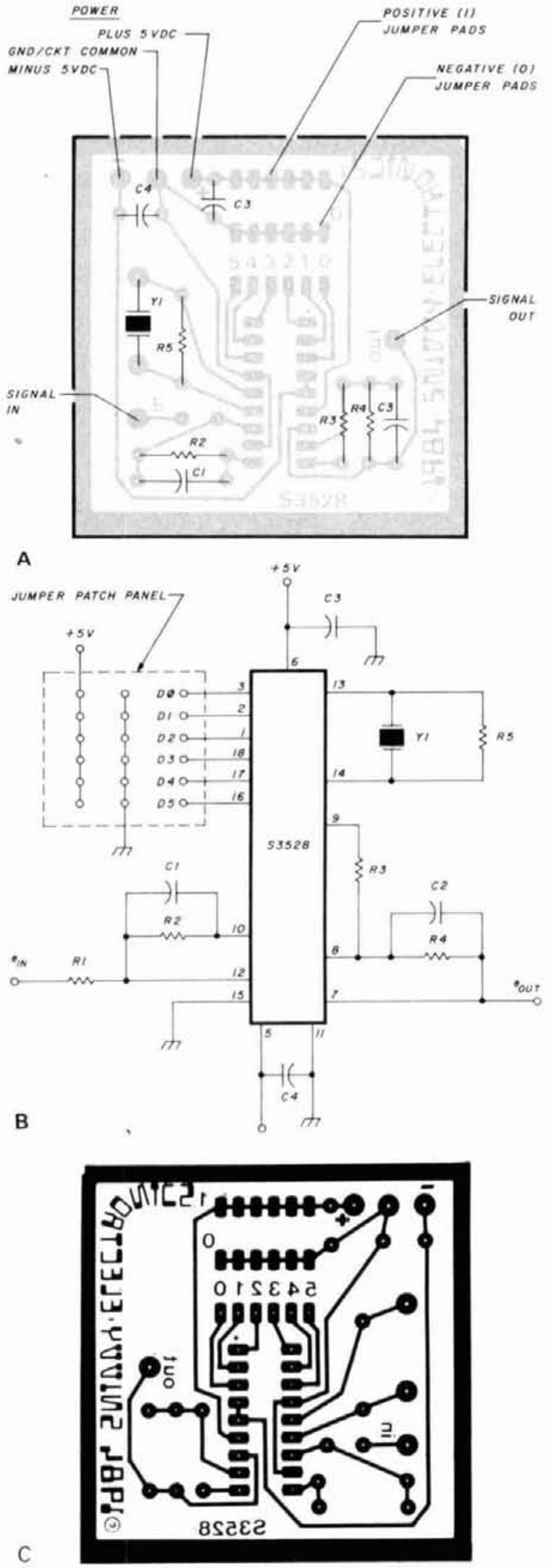

fig. 6. (*A*) shows parts layout; (*B*), wiring diagram; (*C*), 1:1 negative PC layout.

item	description
C1,C2	determined by the desired frequency, polister, 35 VDC
C3,C4	0.01 μF, disc ceramic, 35 VDC
IC1	S3528 AMI, Inc.
R1,R2,R3,R4	10 kilohm, 1/4-watt, carbon film resistors
R5	10 megohm, 1/4-watt, carbon film resistor
socket	18 pin, solder tab type
Y1	3.58 MHz "color burst" crystal, HC-6-U holder
Either printed circuit card or perforated wiring board	
Miscellaneous: solder, 12-inch tinned, bare jumper wire	

$$C = \frac{1}{6.28 \cdot 200 \cdot 35000} = 0.02\ \mu F$$

The values of R and C calculated are not critical. You can use the nearest available device without significant performance degradation.

The output of the S3528, pin 9, should not be used directly to drive a load of less than 10 kilohms. To buffer the output and to provide smoothing of the output signal, the op amp at pins 6 and 7 can be configured as a low-pass filter. Refer to **fig. 5** for details.

To properly size R3, R4, and C2, use the following guidelines and example:

gain of output op amp/filter = 1
R3 = R4 = 10 kilohms
output low-pass filter corner frequency
= 2·elliptic filter cutoff frequency

Calculate:

$$C = \frac{1}{6.28 \cdot R3 \cdot 2 \cdot f} = 0.079\ \mu F$$

(0.1 μF is close enough)

For applications requiring a variable cutoff frequency over a 10:1 (or less) range, the input and output networks should be sized for the highest desired frequency. For applications over a wider range, external filtering or using switched filtering components may be required, depending on the nature of the signals being processed.

As with any new components, it is wise to experiment with the S3528 on a solderless breadboard or a prototype evaluation board prior to incorporating it into a new design.

Readers who wish to etch their own PC cards should refer to **fig. 6** for a full-size copy of the negative artwork and components layout. Printed circuit cards and kits are available from the author; see page 63 for details.

circuit applications

The circuit and PC card have been used in numerous applications around the ham shack and laboratory. As a low-pass filter for preprocessing the signals from the station microphone, it gives good performance while having extremely fast rolloff above the cutoff frequency. As an audio output processor from the station receiver, it is capable of narrowing the effective width of the receiver's internal crystal filters without degrading the shape factor of the crystal filters.

In low-frequency signal processing, it can completely eliminate 60 Hz audio hum from low-level signals. It could be used in a modem or terminal unit for this purpose. Similarly, it can be used in biomedical or biofeedback circuits to eliminate hum and higher frequency products while extracting EEG or ECG signals.

conclusion

The device described in this article is only one of many switched capacitor filter integrated circuits available for use as general-purpose SCF building blocks for producing literally any complex filtering function. Designers of commercial and military equipment are making good use of these devices for lower cost, higher reliability, reduced size, reduced weight, and ease of design.

Judgment must be used when using this filter with pulse or digital circuits. As with any high Q filter, ringing and pulse shape distortion may occur. Again, experiment with the filter before committing to a new design.

The following are available from Snivvy Electronics, 45 Salem Lane, Little Silver, N. J. 07739.

Printed circuit card, glass epoxy	**$ 6.50**
S3528 integrated circuit	**$12.50**
Complete kit of all parts including capacitors C1, C2 for 100 Hz, 1000 Hz, 2000 Hz, 2500 Hz	**$22.50**

N. J. residents please add 6 percent N. J. Sales Tax. All sales include a copy of the S3528 applications brief and complete circuit data.

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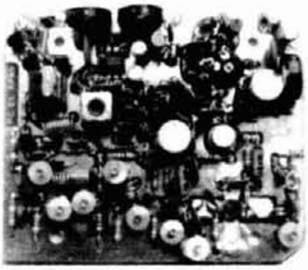

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the reduced-size, full-performance, corner-fed delta loop

For broadband DX performance, try this one on for size

The full wave loop, whether in delta or quad form, has enjoyed popularity over many years. It forms the basis for the well-known and very effective 2-element quad antenna, the history of which can be traced back over 30 years.

A single loop in quad (diamond shape) or delta (apex up) has the following advantages: it requires only a single support, matches easily into low impedance coax, and offers broadband performance.

In May, 1974, L.V. Mayhead, G3AQC, published a very interesting article on the operation of loop antennas close to ground.[1] By modelling the antennas at UHF, he found that the angle of radiation of a delta loop close to ground could be significantly lowered if it were fed at either side corner instead of at the center of its base leg. This had important and useful implications for the use of such antennas on the lower frequencies.

The corner-fed delta loop has also been mentioned as an effective DX antenna by ON4UN in his book *80-Meter DXing*,[2] and references to it have appeared in many popular Amateur journals over recent years.

Space for a full-sized 7 MHz delta loop was not available at my station in North Sydney, Australia. However, previous experiments carried out in England showed very little deterioration in performance of two-thirds size, side-loaded quads at HF compared to full-sized quads. I decided, therefore, to see whether a corner-fed, reduced-size delta loop for 7 MHz could be made to perform as efficiently as a full-sized loop.

14-MHz model first compared

Instead of experimenting at 7 MHz, I chose to first experiment by reducing the size of a full-sized 14-MHz corner-fed delta loop so that a standard of comparison would be available. The 14-MHz loop had been in use for some time and had shown itself to be an effective antenna despite a base height of only 6 feet (1.8 meters). In comparison tests with a half-wave dipole at 30 feet (9.2 meters), it would generally give a 1 S-point improvement in Europe on long path, and seemed about equal to the dipole on the short path to Europe (from Australia). Both antennas were broadside to Europe on long and short paths.

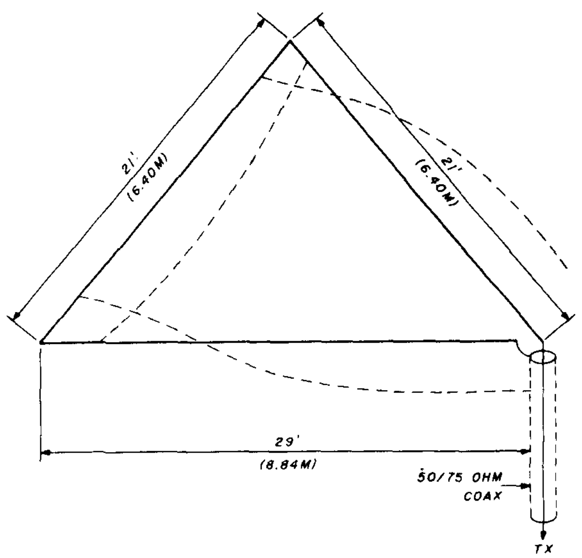
It is worth remembering that even when the delta loop and dipole delivered equal results, the dipole had the advantage of being nearly a half wavelength high on 14 MHz. To achieve the same effective height for a dipole operating on 7 MHz or 3.5 MHz would mean heights of over 60 feet (18 meters) and 130 feet (40 meters), respectively!

current distribution determines polarization

In the original article on the corner-fed delta loop, the loop was not an equilateral triangle, but instead had sides in the ratio 1:1:1.4, where 1.4 represents the base of the apex-up triangle. This configuration means that the two sloping sides meet at a right angle to each other, and the vertical height of the triangle formed is not as great as it would be if the triangle were equilateral in shape.

The current distribution of a delta loop fed in one corner is shown in **fig. 1**. The phase of the currents in the two sloping legs is such as to make it resemble two vertical antennas fed in phase so that maximum radiation would take place in a plane broadside to the plane of the antennas. Although the sloping sides of the delta loop are at 45 degrees to the horizontal, the

By **V.C. Lear, G3TKN**, 53 Chaplains Avenue, Cowplain, Portsmouth PO8 8QH England



DIMENSIONS FOR FULL-SIZE CORNER-FED DELTA LOOPS

frequency	total loop circumference		length of each side		length of base	
(MHz)	(feet)	(meters)	(feet)	(meters)	(feet)	(meters)
3.600	279 ft. 2 in.	(85.09)	82 ft.	(24.99)	115 ft. 2 in.	(35.10)
7.075	142 ft. 0 in.	(43.28)	42 ft.	(12.80)	58 ft. 0 in.	(17.68)
14.150	71 ft. 0 in.	(21.64)	21 ft.	(6.40)	29 ft. 0 in.	(8.84)

fig. 1. Current distribution for a full size 14 MHz corner-fed delta loop. The inner conductor of coax connects to sloping leg.

phase of the currents in both sloping legs produces a predominately vertically polarized signal.

My objective with the smaller loop was to try to recreate a similar current distribution to that described for the full-sized loop.

A loop two-thirds the size of the normal 14 MHz loop was made. It had sides of 14 feet (4.27 meters), and a base of 20 feet (6.1 meters). However, before I made a serious attempt to load this loop to obtain the conditions of current distribution previously mentioned, I decided to simply series-load the loop with a coil at its feedpoint and observe the effects.

When the loop was loaded in this way, it was possible to lower its resonant frequency from 20.9 MHz to 16.5 MHz by adding small amounts of inductance. However, tuning the loop lower than 16.5 MHz required increasingly larger amounts of inductance, until finally a coil of 17 turns close wound on a 2-inch (5-cm) diameter form was needed for resonance at 14 MHz. This coil had a measured inductance of 17.1 microhenries, indicating that the antenna was 1500 ohms capacitive reactive at 14 MHz. The radiation resistance of the loop was too low to allow a good match into 75-ohm coax, and even with the feeder tapped into the coil to obtain a match, results were very poor. This was not really surprising, because by virtue of being placed at the feedpoint, the coil would have been carrying high current, introducing high loss into the system. The current distribution of the loop loaded in this way would not resemble the full-sized version. At best, then, the loop can be lowered in frequency by up to *20 percent* of its natural resonant frequency by simple series loading with a coil. This is probably not the best way to load the antenna, but it may still give useful results. To bring the 20.9 MHz resonant loop to resonance on 14 MHz represents a *33 percent* lowering of frequency.

the effects of base loading

The next experiment consisted of increasing the sides of the loop from 14 feet (4.27 meters) to 17 feet 9 inches (5.41 meters), but keeping the base at 20 feet (6.1 meters) and to try loading the base wire. If the base could be loaded, maximum current would appear in both sloping sides, with a voltage point at the top of the loop. The sides of 17 feet 9 inches (5.41 meters) represent a quarter wave each on the loop at 14 MHz, since the loop circumference in feet is given by 1005/frequency in MHz (bear in mind that there is no *end effect* on the wire of a loop).

The objective with the loop in **fig. 2** was to make the base look like an *electrical half wavelength*, with a voltage point in the middle of the base wire. This would then result in each leg carrying high currents in phase, and a low impedance point at each corner of the loop.

The loading of the base wire was quite easily achieved by connecting two five-foot (1.52-meter) lengths of 300-ohm twin ribbon feeder, shorted at their far ends, 5 feet (1.52 meters) in from each corner of the loop, as shown in **fig. 2**. The ribbon feeder was used instead of a coil because it was felt losses would be lower, and ribbon feeder proved easier to trim and was less bulky than a coil. A 3-foot 6-inch (1.07 meter) piece of stiff wire was connected at the voltage point midway along the base and brought the loop to resonance at 14.15 MHz.

The loop matched well to 75-ohm coax, with an SWR of less than 1.5:1 across the band. The performance of the loop compared well to the full-size version. However, the two hanging stubs would obviously present a problem when scaled up, so I decided to hang two wires from the middle of the base, and pull them back on themselves, as shown in **fig. 3**. This idea worked extremely well, and with each wire 9 feet 3 inches (2.82 meters) long, the loop was again brought to resonance at 14 MHz, but without the inconvenience of hanging stubs.

This method of loading has the advantages of low loss and ease of trimming. Performance was again similar to the full-sized loop.

reducing leg length

Now that the base had been successfully loaded so that it looked like an electrical half wavelength, an attempt was made to reduce the two sloping legs back to 14 feet (4.27 meters) by taking up the extra wire in the form of a closed stub at the top of the antenna. Various closed stubs using open wire line and 300-ohm ribbon feeder were tried, and although the antenna could be resonated each time, the radiation resistance at the feedpoint was too low to match into 75- or 50-ohm coax.

The stubs were dispensed with, and a single wire 7 feet (2.13 meters) in length connected to the apex of the loop, and hanging vertically brought the antenna to resonance. Although this proved to be a much simpler way of resonating the loop, the radiation resistance at the feedpoint was still too low.

A 4:1 matching transformer at the feedpoint connected in such a way so as to step up the impedance of the loop did provide a match, but the bandwidth of the loop was too narrow. It was felt that the introduction of a matching transformer would start to make the whole exercise rather cumbersome and introduce extra losses apart from the unacceptable bandwidth.

After some experimentation, the sides were increased to 16 feet (4.88 meters), with a 3-foot 6-inch (1.07-meter) wire hanging from the apex. This produced an SWR of less than 1.5 to 1 across the whole band even when the base of the loop was only 4 feet (1.2 meters) high.

Results with this loop were still comparable to the full-sized loop, and represented a reduction of the full-sized loop of 27 percent in terms of circumference, with a corresponding reduction in the perpendicular height of the triangle by 18 percent.

40-meter model

The final experiment was to double the dimensions for 7 MHz operation. This meant that the sides were each 32 feet (9.75 meters), base 40 feet (12.19 meters), apex vertical loading wire 7 feet (2.13 meters), and base loading wires each 18 feet 4 inches (5.59 meters). These dimensions resulted in resonance at around mid-band on 7 MHz with an SWR of no greater than 1.5 to 1 across the band.

A thin, light nylon cord was attached between the bottom of the apex loading wire and the midpoint on the base. This was necessary to keep the apex loading wire from blowing around and also to help prevent sag in the base wire, the latter which was also supporting the bottom loading wires. The loop was pulled slightly away from the mast so as to keep any interaction between the partially metal mast and itself to a minimum.

The results with the 7 MHz loop were good. I worked many continents—including America and Europe — with good reports.

It should be remembered, however, when assessing the performance of any antenna, there is no one antenna that will give excellent results on all paths, during all types of propagation, and over all distances. The corner-fed delta loop is a predominately low-angle radiator and as such should generally be at its best over longer distances. If one is only interested in working, for example, up to 1000 miles, a dipole with its higher angle of radiation could be expected to give better results. If a corner-fed delta loop is erected, this point should be remembered in any comparison checks.

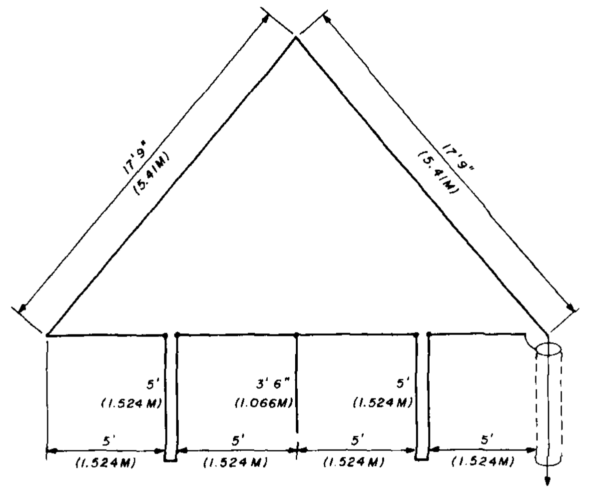


fig. 2. Initial method of loading loop using lengths of 300-ohm ribbon feeder shorted at far end.

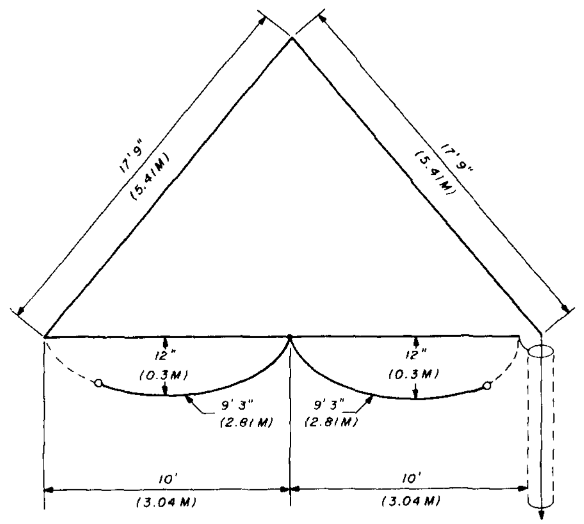


fig. 3. Improved method of loading using wires connected to center of base and pulled out as shown. Light nylon cord is used to secure their ends.

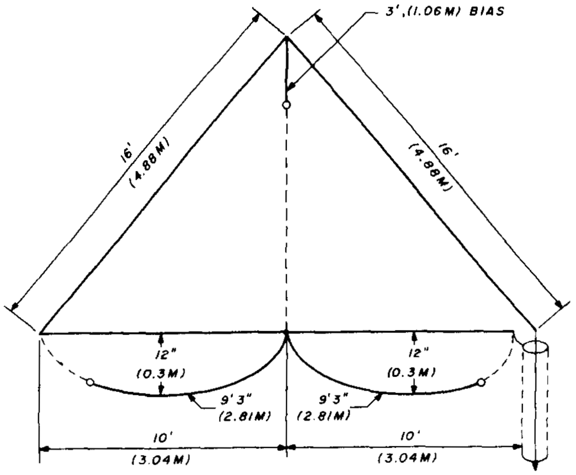


fig. 4. The final loaded loop for 14 MHz. All dimensions are doubled for operation on 7 MHz. Base loading wires are each 18 feet 4 inches (5.59 meters). The top loading wire is made taut by securing to base with thin light nylon cord.

how about 80 meters?

There is no reason those with sufficient space and mast height could not assemble a 3.5 MHz version of this antenna by simply doubling the dimensions given. A 3.5 MHz version would require a mast height of 56 feet (17 meters), which would allow the base of the loop to be 6 feet (1.8 meters) above ground. The horizontal distance taken up would be 80 feet (24.4 meters). These dimensions represent a considerable saving on the full-sized loop and are feasible for many Amateurs.

Final trimming of the loop should be done by adjusting the lengths of the bottom loading wires simultaneously so that their respective lengths are always equal. The top vertical loading wire should not be trimmed. Any trimming should be carried out with the antenna at its normal height, because bringing the base wire closer to ground tends to reduce the resonant frequency of the antenna. The base wire should preferably be a minimum of 6 feet (1.8 meters) above ground. While good results have been achieved with lower positioning than this, it is not recommended; in the original article[1] on the corner-fed delta loop, the base was 10 feet (3 meters) high and this, of course, would be a better height to aim for.

The radiation resistance of a loop antenna drops as it is reduced in size and also as its height above ground decreases. Consequently, 50-ohm coax is recommended for the feeder, although this is not critical. Good results have been obtained with 75-ohm cable.

further experimentation

I hope that by describing some of the experiments carried out, and results obtained, others might be encouraged to experiment further. There is no reason, for example, why those with a little extra space might not erect another similar loop 0.12 to 0.2 of a wavelength *behind* the driven loop, with this additional loop tuned to act as a reflector. To obtain reflector operation of this second loop, the bottom loading wires would have to be increased slightly to bring the loop to resonance some 5 percent lower in frequency from the driven loop.

It should be noted that any attempt to reduce the size of an antenna will be accompanied by a corresponding reduction in radiation resistance, bandwidth, and overall system efficiency. The corner-fed delta loop loaded in this way represents what I believe to be a reasonable compromise between these parameters and acceptable size.

references

1. L.V. Mayhead, G3AQC, "Loop Aerials Close to Ground," *Radio Communication*, May, 1974, page 298.
2. J. Devoldere, ON4UN, *80-meter Dxing*, Communications Technology, Inc., Greenville, New Hampshire, 1978.

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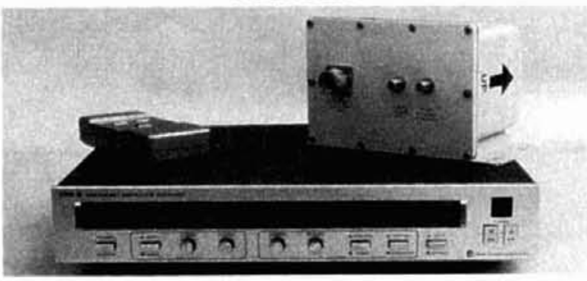

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ham radio TECHNIQUES

the 5/8-wave VHF antenna revisited

In April, 1935, Gihring and Brown published their classic study on the field strength pattern along the ground for vertical antennas of different heights.[1] Their work was part of an ongoing effort to develop a good anti-fade antenna for the broadcast band. One result of their experiments was the popularization of the 5/8-wave vertical antenna; its design combined high radiation efficiency with a power gain of nearly 3 dB over a 1/4-wave comparison vertical antenna.

Since these classic experiments, the 5/8-wave antenna, in combination with an extensive ground radial system, has become a broadcast industry standard antenna.

Amateurs, however, have used this interesting antenna on the HF and VHF bands with mixed results. When an elaborate ground system is used, the antenna performs as might be expected. But when used with a radial system (as is often the case on the VHF bands), the antenna often proves to be a disappointment.

I found this out last summer when, in an attempt to get into a distant 2-meter repeater, I switched from a 1/4-wave ground plane antenna to a 5/8-wave antenna. Any improvement in signal strength at the far-distant repeater was a product of the imagination.

Why didn't it work? What happened to the 3 dB signal increase I was supposed to get from the bigger antenna? Good question.

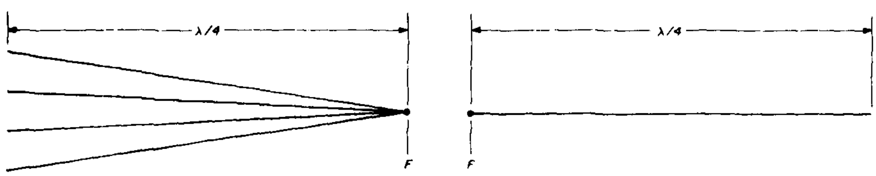


fig. 1. Quarter-wave ground plane, radials collapsed and mounted on its side, transforms itself into a half-wave horizontal dipole.

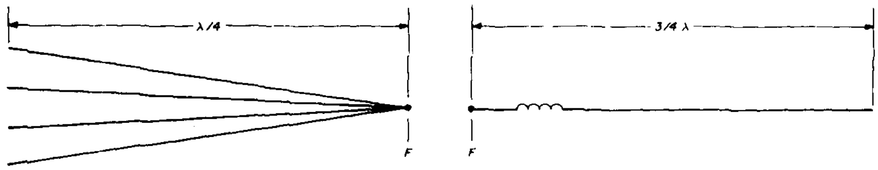


fig. 2. The "5/8-wave" vertical consists of an electrical 3/4-wavelength element and four 1/4-wavelength radials. With radials collapsed and mounted on its side it transforms itself into an asymmetrical antenna.

the W8HXC tests

In my November column I reported on the tests that W8HXC (Ralph) and AF8B (Don) had run on various 2-meter vertical antennas. They found that under many circumstances the feedline became part of the antenna, despite attempts to make the radials isolate the antenna from the feedline, as they are supposed to do.

The simple 1/4-wave ground plane would provide good isolation between antenna and coax line if the line were wrapped into a two-turn RF choke coil about 1 1/2 inches (3.8 cm) in diameter. The coil was placed directly below the antenna. Examining the coaxial line with an "RF sniffer" revealed the line was "cold," provided it dropped directly down beneath the antenna.

The 5/8-wave antenna exhibited current maxima along the outside of the coaxial line until (by cut-and-try) the quarter-wave radials were positioned about 3/8-wavelength below the base of the antenna.

The final conclusion of these tests

was that radials could be placed *any* distance below the antenna as long as the sum of radial length and distance from the antenna base totalled 5/8-wavelength.

the W6SAI tests

This information sounded good to me, so I repeated the tests that Ralph and Don reported. What they said was borne out in fact, but the 5/8-wave vertical antenna still didn't show any appreciable gain over a simple 1/4-wave ground plane antenna.

While I was debating whether I should relegate the antenna to the junk yard, I remembered something I'd read in an old edition of the ARRL *Antenna Book*.[2] The example cited was a horizontal antenna, but the point was made that feeding an antenna in an asymmetrical manner tilted the pattern away from the feedpoint. This could be the clue I was looking for!

If the 1/4-wave ground plane is drawn with the radials in-line with the radiator, it transforms itself into a dipole, as shown in **fig. 1**. If the 5/8-wave vertical is drawn with the usual 1/4-wave radials in-line, it resembles **fig. 2**. The radiator portion of the antenna is really a 3/4-wave resonant section, with 1/8-wavelength wound into a coil so that the advantageous characteristics of the 5/8-wave radiator are retained. The feedpoint (F-F) feeds a lopsided antenna configuration!

Many years ago I had a horizontal "Zepp" antenna that consisted of exactly this configuration (**fig. 3**). I remember that it was impossible to balance the antenna currents and RF in the halves of the antenna. I solved the problem by making the short section the same length as the long one. All my loading problems went away.

Enough circumstantial evidence existed at this point for me to substitute 3/4-wavelength radials for the 1/4-wavelength radials on my 2-meter vertical antenna. This is quickly and easily done, as shown in **fig. 4**. This antenna provided improved performance over the 1/4-wavelength ground plane and also over the 5/8-wavelength extended antenna with 1/4-wavelength radials.

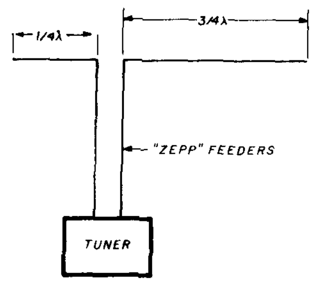


fig. 3. This old "Zepp" antenna at W6SAI looked good on papar but never tuned up properly.

adjusting the 5/8-wavelength vertical

The new antenna was quite easy to adjust. The four radials were precut to 57 inches (145 cm) and the whip was cut two inches longer than estimated length to allow for some pruning. The feedpoint was arbitrarily tapped on the base coil and the antenna was temporarily erected in the air, free and clear of nearby objects.

An SWR check made across the band revealed that the SWR curve canted toward the low frequency end of the band and was quite constant between 1.9-to-1 and 1.7-to-1. The vertical whip was trimmed 1/4-inch (0.6 cm) at a time and the SWR curve checked after each "snip." The SWR seemed to bottom out at about 1.5-to-1, with a very broad response, indicating good bandwidth performance.

Squeezing and expanding the bottom base coil helped a bit and the point of lowest SWR was zeroed in at 146.0 MHz. Unfortunately, the SWR was still too high for us (W6SAI, W6EMD, K6KCM).

The final step was to adjust the feedpoint tap, a quarter-turn along the coil at a time. Now, we were really getting somewhere! After two or three trials, a tap point was found where the SWR on the transmission line was 1.1-to-1, or better, at 146 MHz, rising to about 1.6-to-1 at the band edges.

We observed that the coil tap, number of turns, and vertical antenna length were interrelated. If the antenna was too short, increasing coil induc-

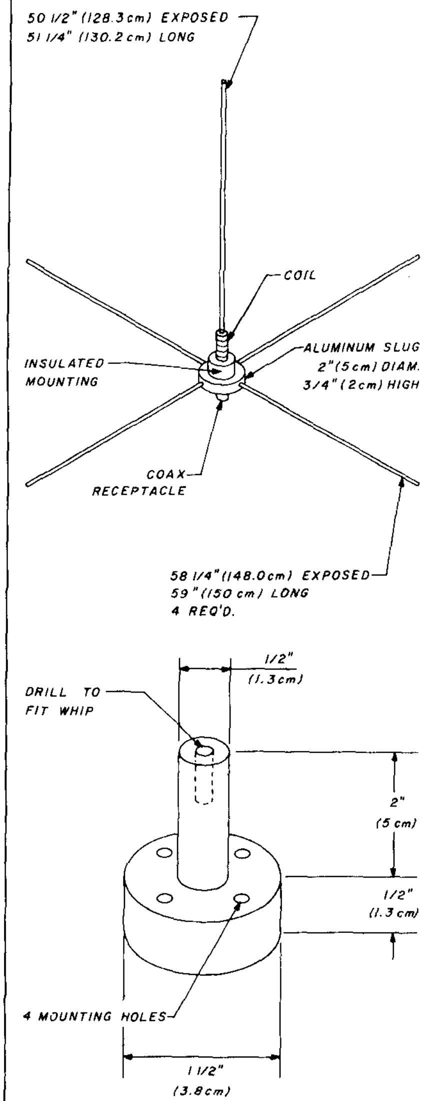


fig. 4. The W6SAI 5/8-wave expanded ground plane for 144 MHz.*

***Note: Elements are built from lengths of 5/16 inch (7.9 mm) diameter aluminum tubing. The vertical whip section is 51-1/4 inches (130.2 cm) long, with 50-1/2 inches exposed above the coil mount. Each radial is 59 inches (150 cm) long, with 58-1/4 inches (148.0 cm) protruding from the aluminum slug. The coil form and support are cut on a lathe from a single block of LEXAN,® a polycarbonate material having good resistance to ultra-violet light (sunlight). It is drilled at the top to accept the whip and at the bottom (4 places) to fit the aluminum mounting slug. The base coil is wound on the top portion of the form. The coil consists of 6-1/2 turns, No. 14 bare copper wire, spaced to 1-1/8 inch (2.86 cm) length. Bottom of the coil is grounded to the aluminum mount. Top end of the coil is attached to the vertical whip by a short strap which encircles the base of the whip and is held in place with 4-40 hardware. Feedline is tapped on the coil 2-1/3 turns from the grounded end. A coaxial receptacle is mounted on the bottom of the mounting slug and the whole antenna is mast-mounted by means of an L-shaped bracket.**

band (MHz)	antenna electrical length (half wave)	physical length of radiating element (feet)	resonant frequency (MHz)	2:1 SWR antenna bandwidth (MHz)	ARRL design frequency (MHz)	Amateur band frequency allocation (MHz)
30	5	241	10.105	9.953 to 10.256	10.12	10.100 to 10.15
20	7	241	14.188	13.975 to 14.400	14.17	14.000 to 14.350
16	9	241	18.270	17.995 to 18.544	18.11	18.068 to 18.168
15	11	254	21.210	20.892 to 21.528	21.22	21.000 to 21.450
12	13	254	25.080	24.707 to 25.460	24.94	24.890 to 24.990
10	15	254	28.950	28.523 to 29.305	28.60 29.20	28.000 to 29.700

fig. 5A. Dimensions and specifications of K4EF six-band long-wire array.

tance brought it back into resonance. The whole thing was very forgiving and, when an extra-eager "snip" cut the antenna too short, a slight compression of the base coil brought the SWR back to where it had been before.

No antenna gain measurements were made as no antenna range was available at the time. On-the-air tests indicate that the antenna is doing the job it was intended to do: provide a good signal at distant repeaters that could not be triggered with a conventional ground plane antenna.

who was first?

Amateur Radio historians often come up with unusual and interesting items. The 5/8-wave vertical antenna popularized by Gehring and Brown in 1935, was patented in the United States by Andrew Alford on April 2, 1940. However, the same antenna concept had been patented in Germany on October 24, 1932, and even this was an extension of a previous patent on the same antenna design issued earlier in Russia.

To top it off, the popular coaxial balun (so-called "Collins Balun") was patented by W.B. Bruene with U.S. patent 2,777,996 in 1954. An identical balun was patented in Germany by Telefunken in August, 1942.

It looks as if a great deal of effort has been expended in re-inventing the wheel!

My thanks to Dipl. Eng. A. Krischke, DJ0TR, OE8AK, for these details. This well-known historian has a private collection of over 1000 radio patents dating back to the early antenna patents of Marconi (1896), Lodge (1898), and Braun (1898).

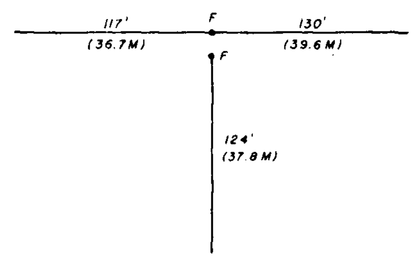


fig. 5B. Top view of K4EF antenna array.

the K4EF antenna for 10-30 MHz

This simple antenna has consistently outperformed a straight long-wire antenna on long distance DX contacts. It was developed by Ev (K4EF) to cover the new ham bands at 18 and 24 MHz, as well as the regular 30, 20, 15, and 10 meter bands. The feedpoint impedance is close to 200 ohms on all bands, so a good quality 4-to-1 balun provides a convenient match to a 50-ohm transmission line. No antenna tuner is necessary, except on the low (CW) end of the 10-meter band.

Antenna dimensions are given in **fig. 5A**. A view of the array is shown in **fig. 5B**. Wire length is critical and should be duplicated to within 2 inches (5 cm). Copper-clad steel wire is recommended, as pure copper wire will stretch over such a span. If kept reasonably in the clear, away from large metal objects, and thirty feet (9 meters) or higher above ground, the electrical characteristics will closely match those listed in the illustration. The antenna may be mounted on a mast or tower at the feedpoint and the three legs suspended in a flat-top or inverted-V configuration. The apex angles of the wire legs are not critical and the sum of the angles may be varied from 180 degrees to 120 degrees, or less.

lightning protection

Lightning is a problem in many areas of the country where frequent electrical storms occur. This antenna is no more vulnerable than other antennas of its size and it has the advantage of being easily protected. The balun (which can be destroyed by a nearby lightning strike) and coax are moved to the base of the tower and the antenna is fed with a 200 ohm open-wire line which can be switched to ground at the tower base when the antenna is not in use (**fig. 6**.)

The line is made up of parallel-connected lines. K4EF's line used two insulators: one at the top and the other at the bottom. Thirty-pound weights were attached to the bottoms of each wire to keep the whole line under tension. This eliminated intermediate insulators. Surplus, heavy-duty 208 ohm "ribbon" line can also be used for the transmission line.

receiver overload

The large capture area of the antenna results in large signal voltages (sometimes from unwanted stations) which may overload the front end of

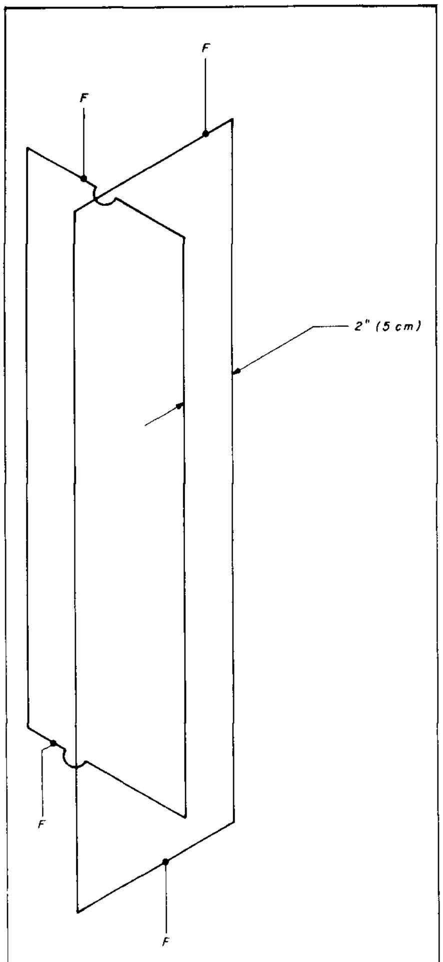


fig. 6. Construction of a 200-ohm line. Diagonally opposite wires are joined at each end of line. Spacing to centers of wires on both sides of square is 2 inches (5 cm) for No. 10 wire. Base of line can be grounded when antenna is not in use.

the receiver. A simple RF attenuator will help, if your receiver does not have some sort of input protection. The circuit of **fig. 7** is suggested.

protecting wooden masts at or below ground level

Fourteen years ago my friend Stu, W2LX, erected a wooden mast on a ground post sunk into moist soil. The post was a 4 × 4, about five feet long, with over 3 feet sunk into the ground. The problem was how to protect the portion of the post buried in the ground.

A previous mast had been treated heavily with wood preservative, but it has rotted out after a few years in the

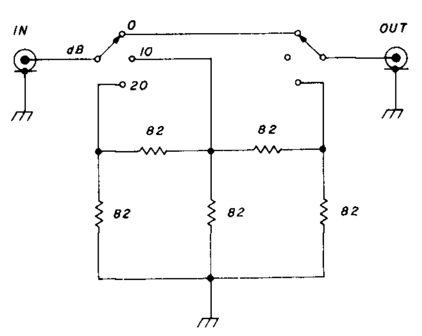


fig. 7. Simple two-step attenuator.

ground. This time, Stu had a better idea.

An untreated mast post was wrapped with three layers of aluminum foil — the heavy-duty type used in the kitchen. The layers of foil were arranged so that the seams did not overlap, and the foil was carefully folded around the base of the ground post. The foil was wrapped with vinyl tape at several points. The mast was then placed in the ground hole and the hole filled with dirt and tamped down.

The metal foil protruded a few inches above ground to protect the post from surface water.

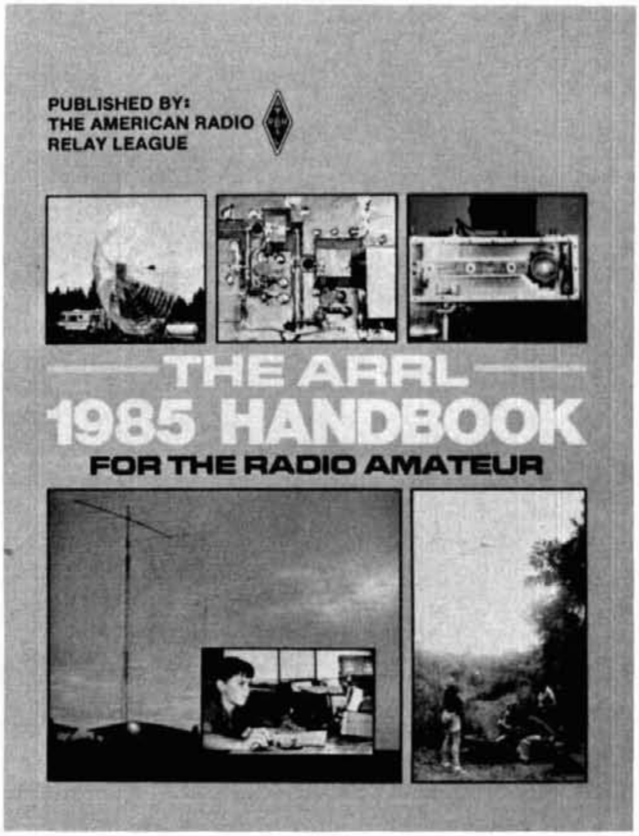
This fall — *fourteen years later* — W2LX moved to a new location. Because the new owner of his home would have no use for the mast, it was taken down. When the ground post uprooted, Stu found the unusual protective technique had been an unqualified success! Once the soil had been washed from the post, the foil looked practically new. Carefully removing the tape and unwrapping the foil, he found that the wood *also* looked almost new. There was no sign of water damage, termites, or rot.

W2LX plans to use this technique when he erects his wooden mast at his new QTH and passes the idea along to other Amateurs who may be interested in erecting a wooden mast.

references

1. H. Gihring and G. Brown, "General Considerations of Tower Antennas," *Proceedings of the IRE*, April, 1935.
2. *The ARRL Antenna Book*, 12th Edition, 1970.

ham radio



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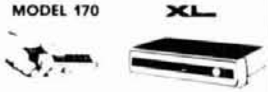

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The PIN diode is a semiconductor device that operates in a manner similar to that of a variable resistor at RF frequencies. The amount of forward DC bias applied to the PIN diode determines its resistance (impedance) to the passage of RF signals.

PIN diodes are attractive for RF switching because they have no moving parts; can "hot switch" large RF currents; can control large amounts of RF current with a relatively small DC bias control current; and don't introduce any significant RF waveform distortion.

PIN diodes are formed from three distinct types of silicon wafers: an *I*ntrinsic layer (pure, non-doped); a *P* doped layer, and an *N* doped layer. **Figure 1** shows a typical PIN diode with its layers identified. A practical package for a PIN diode has the leads attached to the P and N layers and the whole unit encapsulated in either epoxy or glass. It is the thickness of the intrinsic layer which determines the "geometry" of the PIN diode and gives the diode manufacturer the ability to create PIN diodes with different characteristics for special applications.

The external physical appearance of PIN diodes is determined by their intended applications. **Figure 2** shows two Unitrode PIN diodes: a 7300 series and a 4000D series. The 7300 series, about the same size as a 1N4148, is used for microwave attenuators because of its low internal capacitance (0.7 pF). The larger one, series 4000D, is an insulated stud mounted unit with ribbon leads. This unit is used for high power RF switching and certain applications at a 500 *Kilowatts* pulsed power level (1 μs pulse). Tests have shown that with proper heatsinking and DC biasing, the Unitrode 4000D series PIN diode can handle in excess of 3000 watts.*

PIN diode parameters and specifications

All PIN diodes, regardless of their application, share certain common characteristics. One is a forward resistance, R_s, that varies inversely with DC forward bias. This is usually shown in graphical form. (See **fig. 3**, a graph for a typical Unitrode UM4000 series PIN diode.) Note that at 1000 mA (1 amp) forward bias, the UM4000 R_s is approximately 0.1 ohm. At 1 mA the R_s rises to 20 ohms and at 1 μA forward bias, the R_s is in excess of 10 kilohms. The standard for rating most PIN diodes is to provide 100 mA forward bias current at 100 MHz.

Figure 4 shows a set of equivalent circuits for PIN diodes in both a forward and reverse bias state. The forward-biased PIN diode can be considered as equivalent to a series resistor (R_s) and inductor (lead inductance). The reverse-biased PIN diode is equivalent to a resistor (R_p-parallel resistance) in parallel with a capacitor (the C_T . . . total package capacitance) with both of these in series with an inductor (the

By J.R. Sheller, KN8Z, Design Electronics Ohio, 4925 S. Hamilton Road, Groveport, Ohio 43125

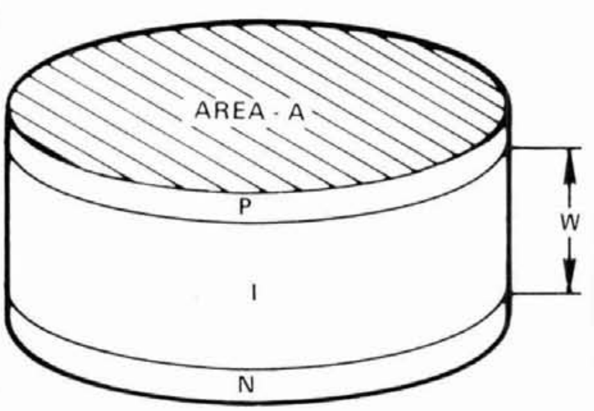

fig. 1. PIN diodes consist of three distinct layers: *P* type, *N* type, and a pure non-doped (*I*ntrinsic) layer.

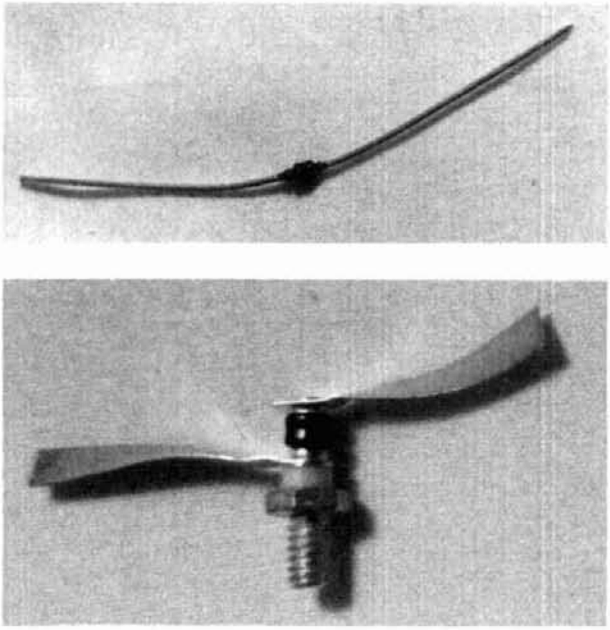

fig. 2. One of the PIN diodes is used in microwave attenuators while the larger unit can actually handle 500 kW of pulsed power.

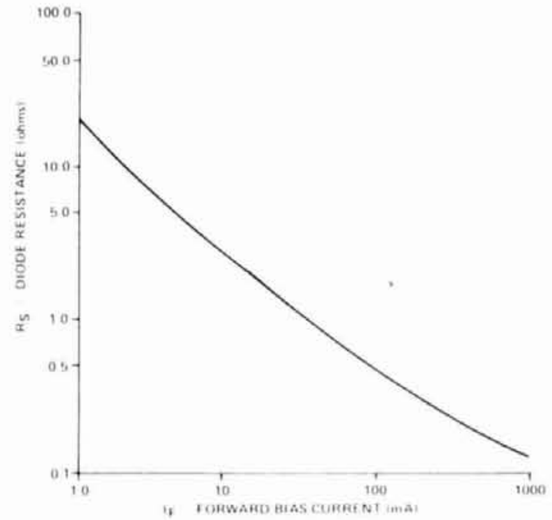

fig. 3. R_s versus current for UM4000 PIN diode.

leads). When PIN diodes are reverse biased, they exhibit a blocking effect to RF as shown in the equivalent circuit of **fig. 4**. As the reverse DC bias is increased, R_p increases and C_T decreases. The limit of the reverse DC bias is determined by the value of V_r (reverse breakdown voltage) of the particular diode. **Figure 5** shows a typical curve of reverse bias voltage (V_r) versus parallel resistance (R_p) for a Unitrode 7300 series PIN diode. **Figure 6** shows a typical curve of reverse bias voltage (V_r) versus total capacitance (C_T) for a Unitrode 4300 series PIN diode.

power handling capability

The maximum power rating of a PIN diode is a function of the forward resistance, R_s, and the amount of RF current flowing through the diode. A PIN diode rated to dissipate 12 watts at 25 degrees C can safely switch 1500 watts of RF. A typical calculation shows why this is so. Assume:

$$\begin{aligned} &RF\ load &&= 50\ ohms \\ &R_S &&= 0.2\ ohm \\ &RF\ power\ level &&= 1500\ watts \end{aligned}$$

$$I = \sqrt{P/R} = \sqrt{\frac{1500}{50}} = 5.48\ amperes$$

The power dissipated by the diode is equal to forward resistance times current squared or:

$$0.2 \cdot (5.48)^2 = 6\ watts$$

The power dissipation of a PIN diode is therefore a function of the load impedance, the forward resistance and the RF current flowing. Consequently, a PIN diode rated to dissipate only 12 watts can, with proper heat sinking, easily handle RF in excess of 1500 watts.

PIN diodes versus vacuum relays

The use of vacuum relays for "high speed" RF switching has been well covered in the literature over the past 20 years (see bibliography). Several commercial amplifiers use vacuum relays for RF switching in order to obtain full break-in "QSK" operation. But using vacuum relays to obtain the high-speed switching times required for full QSK has several disadvantages. These include high cost; the inability to "hot switch"; the necessity for complex switching and protective circuitry; mechanical sound and vibration as the relay opens and closes on each character of CW

*Tests performed at Design Electronics Ohio, Groveport, Ohio.

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Options - continued

		Regular	SALE
CF-1	Cooling fan for PS-15	45.00	
EX-310	Voice synth for 751, R-71A	39.95	
SP-3	External base station speaker	49.50	
	Speaker/Phone patch - specify radio	139.00	129^{95}
BC-10A	Memory back-up	8.50	
EX-2	Relay box with marker	34.00	
AT-100	100w 8-band automatic ant tuner	349.00	314^{95}
AT-500	500w 9-band automatic ant tuner	449.00	399^{95}
AH-1	5-band mobile antenna w/tuner	289.00	259^{95}
PS-30	Systems p/s w/cord, 6-pin plug	259.95	233^{95}
OPC	Optional cord, specify 2 or 4-pin	5.50	
GC-4	World clock	99.95	94^{95}

HF linear amplifier

		Regular	SALE
IC-2KL w/ps	160-15m solid state amp	1795.00	1299

VHF/UHF base multi-modes

		Regular	SALE
IC-251A*	2m FM/SSB/CW transceiver	$749.00	499^{95}

***$50 Factory Rebate - until gone!**

		Regular	SALE
IC-551D	80 Watt 6m transceiver	$699.00	599^{95}
EX-106	FM option	125.00	112^{95}
BC-10A	Memory back-up	8.50	
SM-2	Electret desk microphone	39.00	
IC-271H	100w 2m FM/SSB/CW xcvr	899.00	759^{95}
PS-35	Internal power supply	160.00	144^{95}
PS-15	external power supply	149.00	134^{95}
CF-1	Cooling fan for PS-15	45.00	
EX-144	PS-15/CF-1 fan adaptor	6.50	
AG-25	Mast mtd. GaSFET preamp	84.95	
IC-471H	75w 430-450 SSB/CW/FM xcvr	1099.00	989^{95}
PS-35	Internal power supply	160.00	144^{95}
PS-15	20A power supply	149.00	134^{95}
CF-1	Cooling fan for PS-15	45.00	
EX-144	PS-15/CF-1 fan adaptor	6.50	
AG-35	Mast mounted preamp	84.95	
IC-271A	25w 2m FM/SSB/CW xcvr	699.00	619^{95}
PS-25	Internal power supply	99.00	89^{95}
AG-20/EX-338	2m preamplifier	56.95	
IC-471A	25w 430-450 SSB/CW/FM xcvr	799.00	699^{95}
AG-1	Mast mounted 15dB preamp	89.00	
PS-25	Internal power supply	99.00	89^{95}

Common accessories for 271A/H and 471A/H

		Regular	SALE
SM-6	Desk microphone	39.00	
EX-310	Voice sythesizer	39.95	
TS-32	CommSpec encode/decoder	59.95	
UT-15	Encoder/decoder interface	12.50	
UT-15S	UT-15S w/TS-32 installed	79.95	

VHF/UHF mobile multi-modes

		Regular	SALE
IC-290H	25w 2m SSB/FM xcvr, TTP mic	549.00	489^{95}
IC-490A	10w 430-440 SSB/FM/CW xcvr	649.00	579^{95}

VHF/UHF/1.2 GHz FM

		Regular	SALE
IC-22U	10w 2m FM non-digital xcvr	299.00	249^{95}
EX-199	Remote frequency selector	35.00	
IC-27A	Compact 25w 2m FM w/TTP mic	369.00	329^{95}
IC-27H	Compact 45w 2m FM w/TTP mic	409.00	369^{95}
IC-37A	Compact 25w 220 FM, TTP mic	449.00	359^{95}
IC-47A	Compact 25w 440 FM, TTP mic	469.00	419^{95}
UT-16/EX-388	Voice synthesizer	29.95	
IC-120	1w 1.2 GHz FM transceiver	499.00	449^{95}
ML-12	10w amplifier	339.00	299^{95}

6m portable

		Regular	SALE
IC-505	3/10w 6m port. SSB/CW xcvr	$449.00	399^{95}
BP-10	Internal Nicad battery pack	79.50	
BP-15	AC charger	12.50	
EX-248	FM unit	49.50	
LC-10	Leather case	34.95	
SP-4	Remote speaker	24.95	

Hand-held Transceivers

Deluxe models		Regular	SALE
IC-02AT	for 2m	349.00	299^{95}
IC-04AT	for 440 MHz	379.00	339^{95}

Standard models		Regular	SALE
IC-2A	for 2m	239.50	189^{95}
IC-2AT	with TTP	269.50	199^{95}
IC-3AT	220 MHz, TTP	299.95	239^{95}
IC-4AT	440 MHz, TTP	299.95	239^{95}

Accessories for Deluxe models

		Regular
BP-7	425mah/13.2V Nicad Pak - use BC-35	67.50
BP-8	800mah/8.4V Nicad Pak - use BC-35	62.50
BC-35	Drop in desk charger - all batteries	69.00
BC-60	6-position gang charger, all batts	SALE 359.95
BC-16U	Wall charger - BP7/BP8	10.00

Accessories for both models

		Regular
BP-2	425mah/7.2V Nicad Pak - use BC35	39.50
BP-3	Extra Std. 250 mah/8.4V Nicad Pak	29.50
BP-4	Alkaline battery case	12.50
BP-5	425mah/10.8V Nicad Pak - use BC35	49.50
CA-2	Telescoping 2m antenna	10.00
CA-5	⅝-wave telescoping 2m antenna	18.95
FA-2	Extra 2m flexible antenna	10.00
CP-1	Cig. lighter plug/cord - BP3 or Dlx	9.50
DC-1	DC operation pak for standard models	17.50
LC-02AT	Leather case for Dlx models w/BP-7/8	39.95
LC-2AT	Leather case for standard models	34.95
LC-11	Vinyl case for standard models	17.95
LC-14	Vinyl case for Deluxe models w/BP-7/8	17.95
RB-1	Vinyl waterproof radio bag	30.00
HH-SS	Handheld shoulder strap	14.95
HM-9	Speaker microphone	34.50
HS10	Boom microphone/headset	19.50
HS-10SA	Vox unit for HS-10 (deluxe only)	19.50
HS-10SB	PTT unit for HS-10	19.50
ML-1	2m 2.3w in/10w out amplifier	SALE 79.95
ML-25	2m 2.3w in 20w out amplifier	SALE 179.95
SS-32M	Commspec 32-tone encoder	29.95

Shortwave receivers

		Regular	SALE
R-71A	100 Khz-30 Mhz digital receiver	$799.00	689^{95}
FL-32	500 Hz CW filter	59.50	
EX-310	Voice synthesizer	39.95	
RC-11	Wireless remote controller	59.95	
CR-64	High stability oscillator xtal	56.00	
R-70	100 Khz-30 Mhz digital receiver	749.00	569^{95}
EX-257	FM unit	38.00	
IC-7072	Transceive interface, 720A	112.50	
FL-44A	SSB filter (2nd IF)	159.00	144^{95}
FL-63	250 Hz CW filter (1st IF)	48.50	
SP-3	External speaker	49.50	
CK-70 (EX-299)	12v DC option	9.95	
MB-12	Mobile mount	19.50	

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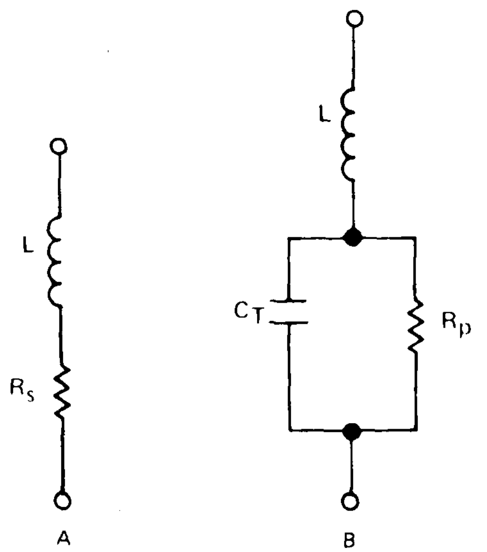

fig. 4. PIN diode equivalent circuits, (A) forward bias, and (B) reverse bias.

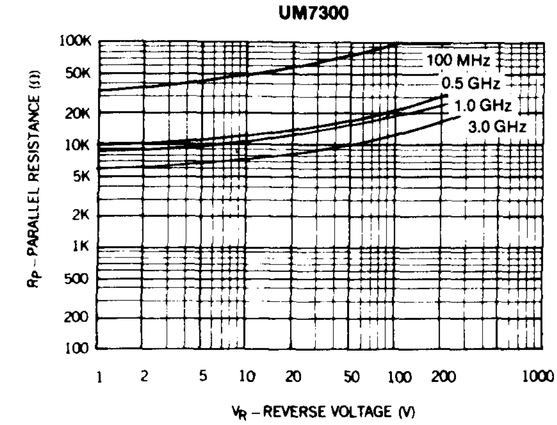

fig. 5. Parallel resistance of a Unitrode UM7300 PIN diode versus reverse voltage over the 100 MHz to 3 GHz frequency range.

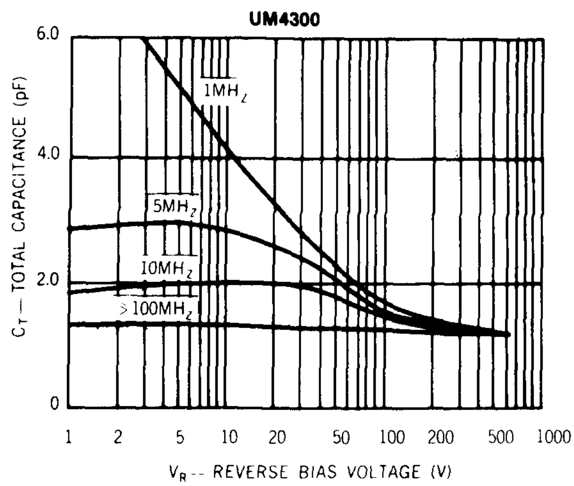

fig. 6. Unitrode 4300 series PIN diode capacitance versus reverse voltage response.

or AMTOR; and maximum switching time of approximately 1-2 milliseconds. Amateur use of PIN diodes for RF switching is a relatively recent occurrence. However, nearly all currently available transceivers that offer full QSK use PIN diodes for their T/R function. Several articles have appeared in the literature over the past 10 years concerning the use of PIN diodes in "low power" T/R switching. Using PIN diodes to perform RF switching functions offers several advantages, including relatively low cost; the ability to be "hot switched"; silent operation; rugged construction; and no complex protective or peripheral circuitry are required. In addition, switching times less than 1 μs are possible.

making PIN diodes work at HF

PIN diodes were designed for and operate best in the VHF and UHF regions. Their use below 30 MHz was delayed because of the need for high inductance and high current RF chokes. The need for capacitors capable of handling 5-10 amps of RF current, with values up to 100,000 pF was also a barrier. PIN diodes can be used at HF, however, and a detailed description of a commercial application that successfully utilizes PIN diodes for RF switching follows.

the QSK 1500

The QSK 1500 switch was developed to allow owners of QSK transceivers to operate full break-in CW or AMTOR at the legal power limit (1500 watts). It uses pin diodes to provide ALL the T/R switching functions associated with the relays in the existing amplifier, with no modifications to either the QSK transceiver or the RF amplifier needed. However, this unit is not intended to make a non-QSK transceiver operate in the QSK mode, nor will it work with a separate transmitter/receiver combination.

A block diagram of the QSK 1500 is shown in **fig. 7** and the schematic of the RF switching section in **fig. 8**.

receive signal path

The receive signal from the antenna travels through the OUTPUT RECEIVE LINE BLOCKER, the RECEIVE LINE PROTECTOR, and the INPUT RECEIVE LINE BLOCKERS, then into the front end of the QSK TRANSCEIVER. With the QSK

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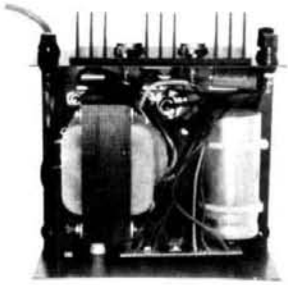

INSIDE VIEW – RS-12A

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PERFORMANCE SPECIFICATIONS

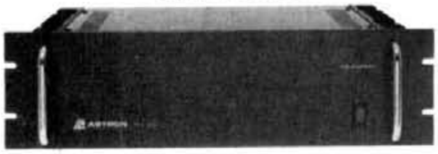
- INPUT VOLTAGE: 105 - 125 VAC
- OUTPUT VOLTAGE: 13.8 VDC ± 0.05 volts (Internally Adjustable: 11-15 VDC)
- RIPPLE: Less than 5mv peak to peak (full load & low line)

MODEL RS-50A

MODEL RS-50M

MODEL VS-50M

RM-A Series



MODEL RM-35A

19" X 5¼ RACK MOUNT POWER SUPPLIES

Model	Continuous Duty (AMPS)	ICS* (AMPS)	Size (IN) HXWXD	Shipping Wt. (lbs.)
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50

RS-A SERIES

MODEL RS-7A

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W X D	Shipping Wt (lbs)
RS-4A	3	4	3¾ x 6½ x 9	5
RS-7A	5	7	3¾ x 6½ x 9	9
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

RS-M SERIES

MODEL RS-35M

- Switchable volt and Amp meter

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt (lbs)
RS-12M	9	12	4½ x 8 x 9	13
RS-20M	16	20	5 x 9 x 10½	18
RS-35M	25	35	5 x 11 x 11	27
RS-50M	37	50	6 x 13¾ x 11	46

VS-M SERIES

MODEL VS-20M

- Seperate Volt and Amp Meters
- Output Voltage adjustable from 2-15 volts
- Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps) @13.8VDC	@10VDC	@5VDC	ICS* (Amps) @13.8V	Size (IN) H x W x D	Shipping Wt (lbs)
VS-20M	16	9	4	20	5 x 9 x 10½	20
VS-35M	25	15	7	35	5 x 11 x 11	29
VS-50M	37	22	10	50	6 x 13¾ x 11	46

RS-S SERIES

MODEL RS-12S

- Built in speaker

MODEL	Continous Duty (Amps)	ICS* Amps	Size (IN) H x W x D	Shipping Wt (lbs)
RS-7S	5	7	4 x 7½ x 10¾	10
RS-10S	7.5	10	4 x 7½ x 10¾	12
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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1500, the "receive signal" never passes through the RF amplifier, but is instead always bypassed around it by the PIN diodes in the QSK 1500. The receive signal from the antenna, is prevented from seeing the tank circuit of the RF amplifier by PIN diode CR2, which is reverse biased. This diode prevents "suck out" or attenuation of the receive signal. PIN diodes CR3, CR4, and CR5 are forward biased and offer a very low impedance path for the receive signal to reach the QSK transceiver. Capacitors C1 and C4 are selected so that receive signal attenuation is typically less than 0.5 dB.

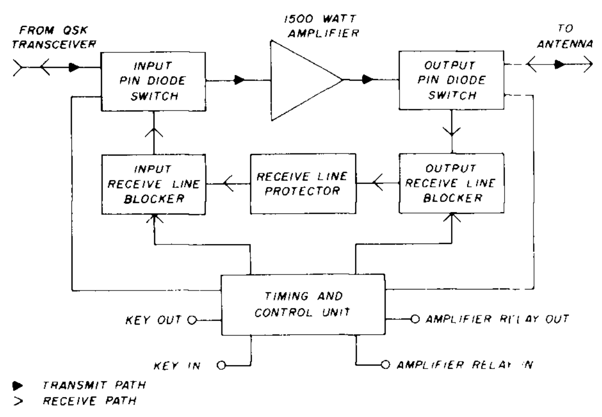

fig. 7. Block diagram of QSK 1500.

transmit signal path

The sequence of events is more complicated in the transmit mode. Let's follow the progress of two transmitted "dots" of CW. (The same pattern also occurs for AMTOR.)

The initial key closure causes the following to happen:

- The timing circuit is triggered on the control board. (This timer is a retriggerable one-shot with a time out of 1.32 seconds.)
- The AMP RELAY OUT line closes and the relays in the RF amplifier are activated.
- The INPUT AND OUTPUT RECEIVE LINE BLOCKERS (PIN diodes CR3, CR4, CR5) are reverse biased with 525 volts DC.
- The input and output PIN diodes switch (CR1 and CR2) are forward biased. CR1 (the input PIN diode) is forward biased with 125 mA DC current, while CR2 (the output PIN diode) is forward biased with 950 mA DC current.
- The KEY OUT LINE from the QSK 1500 triggers the CW KEY JACK on the QSK transceiver and a "dot" of CW is generated.
- The RF from the QSK transceiver flows through C1,

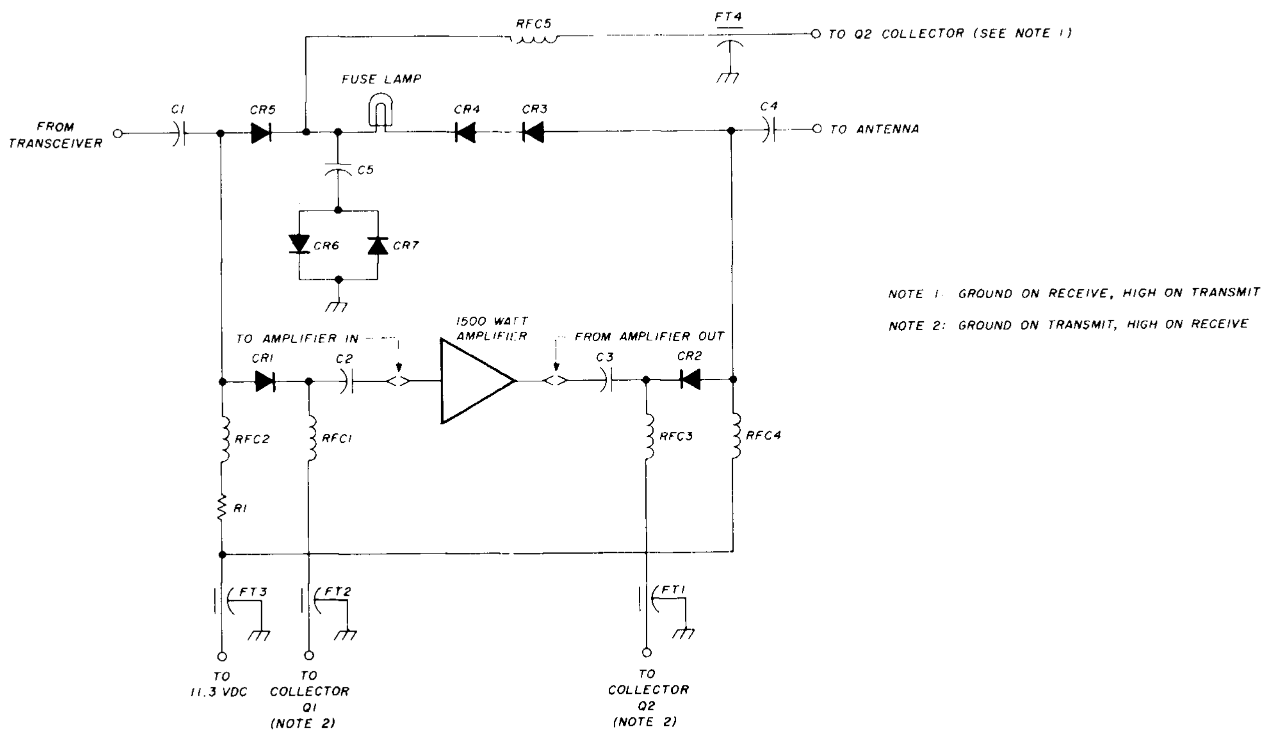

fig. 8. Schematic of RF switching section of QSK 1500.

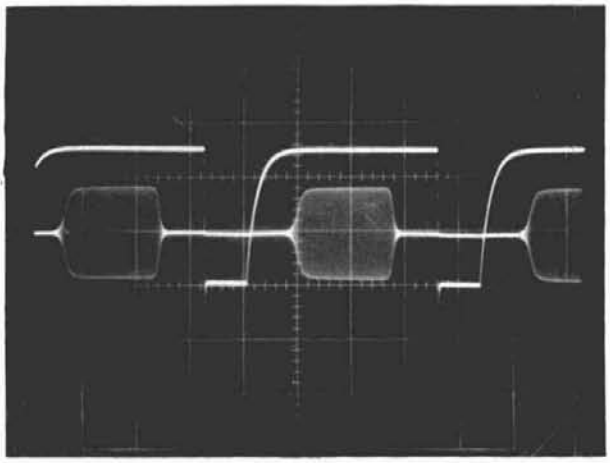

fig. 9. CW waveform inside 500V blocking voltage. (See Note below.)

CR1, and CR2 into the RF amplifier, where it is amplified and then passed through C3, CR2, and C4 into the antenna.

- The 525 volt DC reverse bias applied to PIN diodes CR3, CR4, and CR5 prevents any transmitted RF signal from passing through the RECEIVE LINE.

As soon as the "key" opens the following occurs:

- PIN diodes CR1 and CR2 switch state and go from a forward bias condition to a reverse bias condition.
- PIN diodes CR3, CR4, and CR5 also switch state and go from a reverse bias state to a forward state.
- With the reversal of the bias states of all five PIN diodes, we have now returned to the receive condition.

Figure 9, an oscilloscope photo of a CW waveform (58 WPM dots) inside the 500 volt blocking voltage, shows the "turn on"/"turn off" of the PIN diodes in relation to the blocking voltage.

With every key opening and closure a change of state of the PIN diodes occurs. The relays in the RF amplifier *remain* closed until no key closure has occurred for 1.32 seconds. This assures that no RF is being switched by the RF amplifier's internal relays, and that all switching between transmit and receive is accomplished by the PIN diodes in the QSK 1500.

The RECEIVE LINE PROTECTOR ensures that no RF can reach the transceiver or the RF amplifier input in the unlikely event of an output blocker PIN diode (CR3 or CR4) failure.

Note: Photo shows transmission of a string of dots at 58 WPM. At this rate the receive time between pulses is 8 ms. The built-in delay time between turn-on of the 500V blocking voltage and the beginning of the CW envelope is 7 ms. The CW pulse (dot) is 21 ms in duration; the built-in delay between the end of the CW pulse and turn off of the 500V blocking voltage is 7.5 ms. The CW waveform shows no trace of distortion. Power level, 1380 watts; transmitter, TS930S; amplifier, Drake L7; QSK unit, QSK 1500; scope, TEK 556 dual-trace; wattmeter, Bird 43 with scope coupler; load, Bird Termaline™ 2 kW; keyer, Accu-keyer II. Vertical scale = 200V/cm (× 10 probe with 20V scale); horizontal scale = 10 ms/cm.

conclusion

The appearance of high power PIN diodes represents the dawn of a new era in RF switching. The vacuum relay — the former "king of RF switching" — will slowly but surely give way to the solid state PIN diode. PIN diodes now make possible relatively low-cost, ultra high-speed RF switches, which if properly designed and constructed can operate at power levels exceeding 1500 watts. A totally silent switch in a small package, the PIN diode is here to stay.

acknowledgement

Thanks are due to the technical and engineering staff at Unitrode for assistance with technical data and help with solving various problems that arose.

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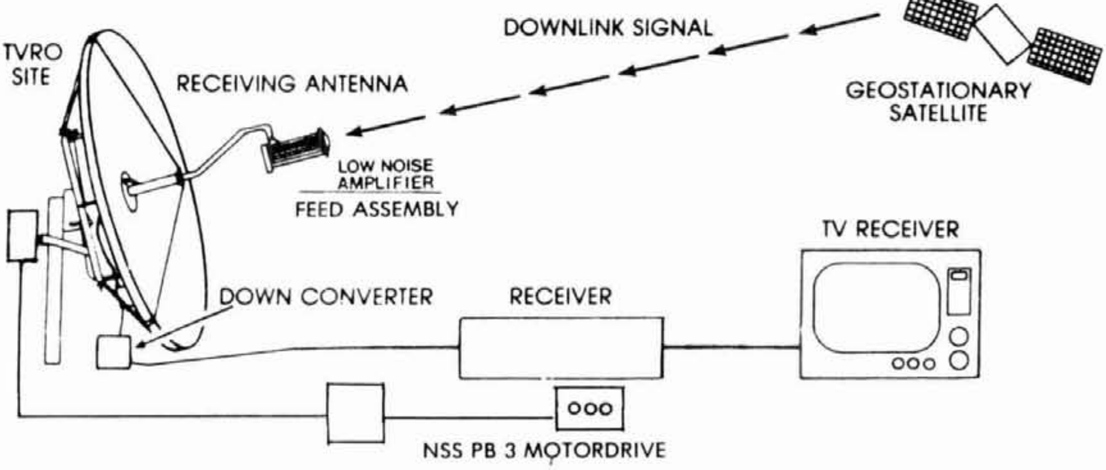

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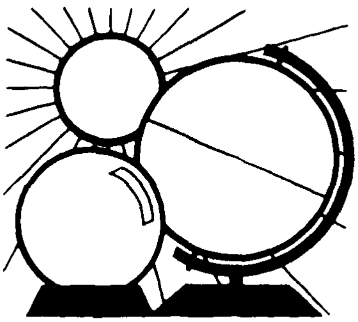

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ionosphere matching

Over the years a few hams have tried to vary the height of horizontal antennas above ground in order to vary the take-off-angle (TOA). By varying the angle at which the signal reflects from the ionosphere, you can control the length of the hops (and consequently, distance) and by orienting the antenna (rotating the beam), the direction the signal takes. This is known as "matching to the ionosphere."

If you can't vary your antenna's height above ground, then fix it permanently at a height that emphasizes your area of interest. (Antenna handbooks include graphs that relate take-off angle, ionospheric layer height, MUF and path length.)* **Table 1** shows a short BASIC program for calculating the take-off angle of a horizontal antenna given the height above ground.

table 1. BASIC program for calculating take-off angle of a horizontal antenna given height above ground. (This program was written for the IBM 4341 in IBM's BASIC/VS.

```
110 R1=180/&PI
120 REM GET INPUT DATA
130 PRINT "ANTENNA HEIGHT ABOVE GROUND, FEET"
140 INPUT H1
150 PRINT "FREQUENCY TO BE USED, MHZ"
160 INPUT F1
170 REM CALCULATE A FULL WAVELENGTH, L1, AT THIS FREQ.
180 L1=984/F1
190 REM CALCULATE FIRST LOBE ANGLE, A1
200 A1=ASN(L1/(4*H1))
210 A1=A1*R1
220 PRINT "ANGLES OF MAXIMUM RADIATION FROM THIS ANTENNA"
230 PRINT
240 PRINT "FIRST LOBE ANGLE, DEG.=",A1
250 PRINT
260 REM CALCULATE 2ND LOBE ANGLE, A2
270 IF (4*H1)<(3*L1) GO TO 500
280 A2=ASN((3*L1)/(4*H1))
290 A2=A2*R1
300 PRINT "SECOND LOBE ANGLE, DEG.=",A2
310 PRINT
320 REM CALCULATE 3RD LOBE ANGLE, A3
330 IF (4*H1)<(5*L1) GO TO 500
340 A3=ASN((5*L1)/(4*H1))
350 A3=A3*R1
360 PRINT "THIRD LOBE ANGLE, DEG.=",A3
370 PRINT
380 REM CALCULATE 4TH LOBE ANGLE, A4
390 IF (4*H1)<(7*L1) GO TO 500
400 A4=ASN((7*L1)/(4*H1))
410 A4=A4*R1
420 PRINT "FOURTH LOBE ANGLE, DEG.=",A4
430 PRINT
500 PRINT "NO OTHER SIGNIFICANT LOBES"
510 END
```

last-minute forecast

The high probability of increased solar flux in the middle of January makes the second and third weeks of *the month favorable for* DX openings on the 10 through 30-meter bands. The openings may be transequatorial in nature, particularly if minor geomagnetic field disturbances occur at the same time as this greater solar activity. The lower frequency bands should be good during the first and last weeks of the month. Low noise and signal absorption during low solar flux periods account for good daytime openings on 40 and 80 meters this year. Geomagnetic field disturbances that produce ionospheric high latitude trough conditions (see December, 1984, *DX Forecaster*) should not be too significant this year except for periods around January 2, 11, 21 and 29.

Lunar perigee occurs on the 15th, with a full moon on the 7th this month. An intense but short-duration meteor shower, the Quadrantid, will reocccur between January 2nd and 4th and last a few hours.

band-by-band summary

Ten and fifteen meters will be open worldwide from sunrise until after sunset during the solar flux peaks this month. Skip distances of 2500 miles (4000 km) (or multiples) are possible, and will occur on the daylight paths.

Twenty meters will be open to some area of the world for the entire 24-hour period many days of the month with the band conditions peaking in all directions just after local sunrise and again toward the east and south during the late evening hours. During hours of darkness the band will peak toward the west in an arc from southwest through northwest, encompassing the Pacific areas.

Thirty meters is a daytime and nighttime band. During the day it will re-

*Next month's *DX Forecaster* will feature such a graph.

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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	
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JANUARY

WESTERN USA

GMT	PST	N Asia Far East	NE Europe	E S. Africa	SE S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	4:00	30	30	15	15*	15	10	10	20
0100	5:00	20	30	15	10	15	10	10	20
0200	6:00	20	30	15	10	15	10	10	20
0300	7:00	20	30	20	10	15	15	10	20
0400	8:00	20	30	20	10	15	15	15	20
0500	9:00	20	30	20	10	15	15	15	20
0600	10:00	20	40	20	15	20	20	15	30
0700	11:00	30	40	20	15	20	20	20	30
0800	12:00	30	40	20	15	20	20	20	30
0900	1:00	30	40	20	20	20	20	20	30
1000	2:00	30	40	20	20	20	20	20	30
1100	3:00	30	40	30*	20	20	20	20	30
1200	4:00	30	40	20	20	20	20	20	30
1300	5:00	30	40	20	20	20	20	20	40
1400	6:00	40	30	20	20	20	20	20	40
1500	7:00	40	30	15	20	20	20	20	40
1600	8:00	40	30	15	20	20	20*	20	40
1700	9:00	40	20	15	20	15	15	20	40
1800	10:00	40	20	15	15	15	15	15	30
1900	11:00	40	20	15	15	15	15	15	30
2000	12:00	40	20	15	15	15	15	15	20
2100	1:00	30	20	15	15	15	15	15	20
2200	2:00	30	20	15	15	15	15	15	20
2300	3:00	30	30*	15	15	15	15	15	20

MID USA

GMT	MST	CST	N Asia Far East	NE Europe	E S. Africa	SE S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	5:00	6:00	30	30	15	15	15	15	10	20
0100	6:00	7:00	30	30	15	20	15	15	10	20
0200	7:00	8:00	40	30	20	20	15	15	15	20
0300	8:00	9:00	40	30	20	20	15	15	15	20
0400	9:00	10:00	40	30	20	20	20	15	15	20
0500	10:00	11:00	40	30	20	20	20	20	20	30
0600	11:00	12:00	40	40	20	20	20	20	20	30
0700	12:00	1:00	40	40	20	20	20	20	20	30
0800	1:00	2:00	40	40	20	20	20	20	20	30
0900	2:00	3:00	30	40	20	20	20	20	20	30
1000	3:00	4:00	30	40	20	20	20	20	20	30
1100	4:00	5:00	30	40	20	20	20	20	20	30
1200	5:00	6:00	30	40	20	15	20	20	20	30
1300	6:00	7:00	30	40	20	15	20	20	20	40
1400	7:00	8:00	20	30	15	15	20	20	20	40
1500	8:00	9:00	20	30	15	15	20	20	20	40
1600	9:00	10:00	20	30	15	15	20	15	20	40
1700	10:00	11:00	20	20	15	10	15	15	20	40
1800	11:00	12:00	20	20	15	10	15	15	15	30
1900	12:00	1:00	20	20	10	10	15	15	15	30
2000	1:00	2:00	30	20	10	10	15	15	15	20
2100	2:00	3:00	30	20	10	10	15	15	15	20
2200	3:00	4:00	30	20	15	10	15	15	15	20
2300	4:00	5:00	30	30	15	15	15	15	15	20

EASTERN USA

GMT	EST	N Asia Far East	NE Europe	E S. Africa	SE Caribbean S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	7:00	30	30	15	15	15	15	15*	20
0100	8:00	30	30	15	20	15	15	15	20
0200	9:00	40	30	20	20	15	15	15	20
0300	10:00	40	30	20	20	15	15	15	20
0400	11:00	40	30	20	20	20*	20	15	30
0500	12:00	40	30	20	20	20	20	20	30
0600	1:00	40	40	20	20	20	20	20	30
0700	2:00	40	40	20	20	20	20	20	30
0800	3:00	40	40	20	20	20	20	20	30
0900	4:00	30	40	20	20	20	20	20	30
1000	5:00	30	40	20	20	20	20	20	30
1100	6:00	30	40	20	20	20	20	20	40
1200	7:00	30	40	20	15	20	30*	20	40
1300	8:00	20	20	15	15	20	20	30*	40
1400	9:00	20	20	15	15	20	20	20	40
1500	10:00	20	20	15	15	20	20	20	40
1600	11:00	20	20	15	15	20	15	20	40
1700	12:00	20	20	15	10	15	15	15	40
1800	1:00	20	20	15	10	15	15	15	30
1900	2:00	30	20	15	10	15	15	15	30
2000	3:00	30	20	10	10	15	15	15	20
2100	4:00	30	20	10	10	15	15	15	20
2200	5:00	30	20	15	15	15	15	15	20
2300	6:00	30	20	15	15	15	15	15	20

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.

semble 20 meters, although signal strengths may decrease during midday during days of high solar flux values. This band will also be useful well into the night and often throughout the night. Once again exceptions to this are nights that follow very high solar flux value days. The poor period is usually the hour or so before dawn (diurnal MUF minimum). The workable distance may be expected to be greater than that of 80-meter DX at night and less than that of 20 meters during the day.

Forty and eighty meters will be the most useful nighttime DX bands. Forty during the daylight hours will be like 30 meters, with lower midday signal levels, *no* predawn propagation failure, and shorter skip distances overall. At night most areas of the world will be workable from slightly before dusk until a little after sunrise. Hops shorten on these bands to about 2000 miles (3218 km) for 40 meters and 1500 miles (2414 km) for 80 meters, but the number of hops can increase because signal absorption in the D-region of the ionosphere is low during the night. The path follows the direction of darkness across the earth, similar to the way in which the higher bands follow the sun.

One-sixty meters will be similar to 80 meters, providing good working conditions for enthusiastic DXers who like to operate during the night and early morning hours, especially at local dawn.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

VHF/UHF high power amplifiers: part 1

Until August 28, 1983, FCC Amateur Radio regulations limited transmitters to a maximum of 1000 watts DC plate input power on CW or 2000 watts PEP input on SSB. This was in contrast to most commercial radio regulations, which are specified in terms of output power. The regulations also specified that any drive power passed on to the amplifier output (such as in grounded grid or transistor amplifiers) had to be accounted for by lowering the DC input power to the final amplifier stage accordingly.

These archaic regulations particularly penalized VHF/UHF experimenters who wanted to take advantage of the maximum power limit for DX and for specialized communication modes such as EME. Klystron and transistor UHF amplifiers frequently have only 35 percent efficiencies; most UHF high power tube amplifiers have typical efficiencies of only 50 to 60 percent. Therefore, when the legal power limits were required, higher priced vacuum tubes were usually necessary in order to maximize efficiency — *often at the expense of linearity*!

These regulations are now history. On August 29, 1983 the FCC revised Part 97 of the Amateur Radio Rules and Regulations, authorizing the measurement of power in terms of maximum PEP (peak-envelope-peak) output, with 1500 watts the maximum limit (except in special subbands of the spectrum where other concerns may necessitate a lower power limit). SSB operation is not particularly affected because a 55 percent efficient VHF amplifier under the old regulations would have delivered up to 1100 watts of PEP output, only 1.35 dB down from 1500 watts. However, a 60 percent efficient CW amplifier would be, under the old regulations, about 4 dB below the new 1500 watts PEP output (CW and PEP output are the same power)!

If you want to run the new power limit in an Amateur VHF/UHF power amplifier, these regulatory changes have profound implications for the design of your next amplifier. Because most articles in the Amateur literature pertaining to VHF/UHF power amplifier design were written before the regulations were changed, this month's — and next month's — column address the broad subject of Amateur VHF/UHF high power amplifiers, with special emphasis on the effects of the new FCC regulations. (Many of the topics discussed will also apply to HF.) A list of recommended references will also be included so you can be sure your amplifier will be equal to the state-of-the-art. I'll concentrate on vacuum tube amplifiers because the typical solid-state devices available today can't economically furnish the maximum legal power in the VHF/UHF spectrum.

power tubes in general

Most high-power transmitting tubes used by Amateurs are technically classified as power grid tubes. Although some pentodes are used on 6 meters, most VHF/UHFers primarily use triode and tetrode power grid tubes. Some of the most popular types are listed, with some of their important specifications, in **table 1**.[1]

Triodes are most often used when a single high voltage supply is desired. With triodes, circuitry is usually less complicated than with tetrodes, especially if a cathode-driven grounded-grid circuit is used. Gain is low to moderate in this type of configuration, but the amplifier is usually easier to stabilize. Tetrodes, on the other hand, require a more complex high voltage supply with a screen grid bias voltage and may require neutralization. However, they usually have higher power gain than triodes, especially if grid-driven.

Most modern power tubes come in either a glass envelope with an internal anode or a ceramic package with an external anode. (The latter are more prevalent on 2 meters and above.) The

table 1. Popular VHF/UHF power amplifier tubes and typical parameters listed in order of plate dissipation.

triodes

types	dissipation[1]	maximum power[2]	PEP[3]	IMD[4]	F_{MAX}[5]	F_{EXT}[6]	other notes
8875	300	770	590	−35	500	900	cathode driven
3CX400A7/8874	400	770	590	−35	500	900	cathode driven
3-500Z	500	1400	740	−40	110	-	cathode driven
3CX800A7	800	1320	750	−36	30	450	cathode driven
3-1000Z/8164	1000	4800	1080	−29	110	-	cathode driven
3CX1500A7/8877	1500	4000	2050	−38	220	400	cathode driven
8938	1500	4000	2030	−44	500	-	cathode driven

tetrodes

types	dissipation[1]	maximum power[2]	PEP[3]	IMD[4]	F_{MAX}[5]	F_{EXT}[6]	other notes
4X150A (old)	150	375	200	-	500	-	discontinued
4X150A/7034	250	500	200	-	150	500	new style anode
4CX250B/7203	250	500	295	−25	500	-	
4CX250R/7580W	250	500	295	−25	500	-	"Ruggedized"4CX250B
4CX300A/8167	300	500	350	−25	500	-	
4CX350A/8321	350	750	300	−30	30	220	poor efficiency at UHF
8930	350	600	350	−27	500	-	formerly DX393
8122	400	660	380	−29	500	-	
4-400A/8438	400	1400	495	−35	110	-	
7650/7651[7]	600	1500	680	−31	1215	-	
4-1000A/8166	1000	4200	1540		110	-	
4CX1000A/8168[8]	1000	3000	1400	−23	110	220	
4CX1000K	1000	3000	1400	−23	110	400	Improved in/out isolation
7213/7214[7]	1500	2500	1250		1215	-	
GL6942	1500	2800			900	-	cathode driven
4CX1500B/8660	1500	2700	1160	−43	110	220	

Notes:

1. Rated plate power dissipation in watts if adequately cooled.
2. Maximum CCS DC input power.
3. Maximum useful peak envelope output power in watts at a low frequency, typical 2-30 MHz.[2]
4. Typical third-order intermodulation distortion level at rated PEP indicated, calculated in 2-30 MHz region.[2]
5. Upper frequency at which the maximum ratings apply.
6. Upper useful frequency if plate voltage and input power are reduced.
7. The second tube listed is a pulsed rated version. It will probably perform the same as the first version.
8. No grid current allowed.

older glass-sealed transmitting tubes with external anodes (for example, the 4X150A) were very efficient *if operated within their ratings* but were not very rugged. However, the newer ceramic insulated tubes, though perhaps somewhat less efficient, will usually take more severe punishment.

Transmitting tubes are usually designed for a particular mode of operation such as Class C, linear, or pulse. Class C operation is preferred for CW operation when high efficiency is required. Linear operation is required for SSB operation. Amateur Radio pulse, a form of Class C, is not permitted below 2.3 GHz.

The tubes designed for Class C operation typically have high efficiency in Class C but will not always be very linear when biased for linear operation, especially when operated near their maximum ratings. Furthermore, Class C amplifiers usually have lower power gain than equivalent linear amplifiers and often generate key clicks.

A good example of a frequently misused VHF/UHF Class C tube is the popular 4CX250B tetrode. This tube was very popular under the old FCC regulations of 1000-watts input power because a pair could be operated efficiently (at 60 to 70 percent efficiency) through 70 cm, a requirement for most EMEers. However, they are also often used in linear service. At the maximum rating of 500 watts input power per tube, the IMD (intermodulation distortion) is typically only −25 dB (see **table 1**) — not very good if you want to be popular with other local operators.

Many high power tubes designed for linear service are costly or less efficient than the Class C-type tubes. This was a significant consideration, especially on EME, before the FCC changed the method of measuring power. That's all different now, and we're no longer penalized for using tubes with less efficiency. As a result, many less efficient tubes, especially those used in commercial television transmitters, may now find their way into Amateur amplifiers after they're removed from commercial service during routine maintenance.

In summary, it would probably be best to design any new power amplifier with linear operation in mind. The small change in efficiency may easily be recovered by higher gain with less likelihood of generating key clicks. Using tubes other than those presently in wide use may also be advisable. **Table 1** and references 2, 3, and 4 should be consulted when choosing a suitable VHF/UHF transmitting tube.

configuration is important

One of the first considerations in designing a power amplifier is the tube type and quantity. Under the old FCC regulations, it was a common VHF/UHF practice to use two tubes in either a parallel or push-pull configuration. This was often done because efficient external anode tubes of moderate power (500 watts typical) were easily obtained, and usually at reasonable prices. Furthermore, many Amateur designs were already available in various publications. However, the new power limits are not easily achieved with the tubes that used to be popular — especially in linear operation.

If push-pull or parallel amplifiers are used, the tubes should be fairly well matched and from the same supplier. Extra metering and balancing circuitry are required to insure that both tubes properly share the load. If neutralization is used, this type of circuitry is doubly difficult. Therefore, it's better to use a single large tube that will do the entire job rather than multiple tubes that will share the load.

gain

The average amplifier gain in the VHF region is between 15 to 20 dB and 10 to 16 dB at UHF. As mentioned, gain is usually higher in linear operation than in Class C. A typical grounded-grid amplifier has 3 to 5 dB less gain but is usually easier to stabilize and match at the input. For example, on 2 meters a typical pair of grid-driven 4CX250Bs operating in Class C will typically deliver 400 to 600 watts of output with 10 watts of drive power, but only 200 to 300 watts on 70 cm (432 MHz). This will increase to 600 to 700 and 300 to 400 watts respectively in linear service. Hence, it should be obvious that an additional low-gain intermediate driver amplifier may be required if a typical 10-watt transverter is used to drive a power amplifier with a kilowatt or higher output capability.

input circuits

The input circuit for typical VHF/UHF amplifiers is one of two major types: either cathode or grid-driven. The cathode type usually requires only some simple "L" or "T" matching section (**fig. 1A**). Grid-driven amplifiers usually require a step-up or transformer type of input circuit with additional tuning and matching components (**figs. 1B** and **1C**).

If push-pull or parallel tube operation is used, circuitry should be designed to allow the drive to be tweaked for each tube (**fig. 1D**). If the input circuit doesn't have RF balancing capability, the input grid-biasing circuit should be designed to allow for independent adjustment of the DC bias voltage for each tube so that the load is shared equally between the tubes. The principal thing to remember when designing the input circuit of a power amplifier is that the configuration chosen primarily performs an impedance match between the input driver (usually a 50-ohm device) and the grid or cathode of the tube being driven. If the network is properly designed, the power amplifier driver will see a low VSWR (2:1 maximum).

stabilization

The screen grids of tetrode amplifiers must be properly bypassed. Most manufacturers and suppliers offer specially designed tube sockets with built-in bypass capacitors. These are highly recommended and, although usually expensive, will justify their cost with improved amplifier stability and efficiency. Some of the more modern tube socket designs (for

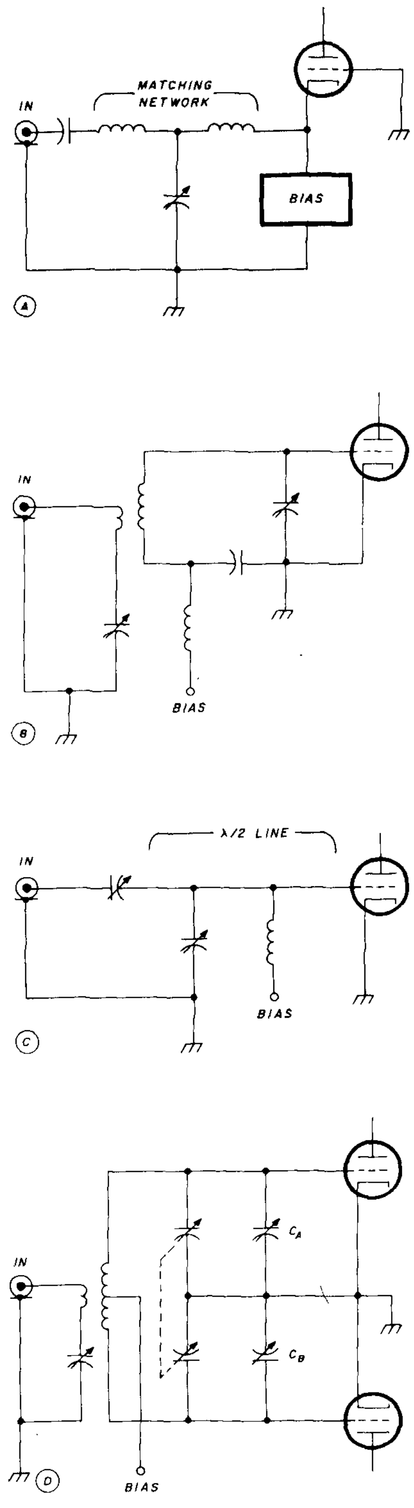

fig. 1. Typical input circuits used on high power VHF/UHF amplifiers. (*A*) grounded grid, cathode-driven, (*B*) single ended, grid-driven, (*C*) single ended 1/2 wavelength line, and (*D*) push-pull. C_A and C_B are balancing capacitors.

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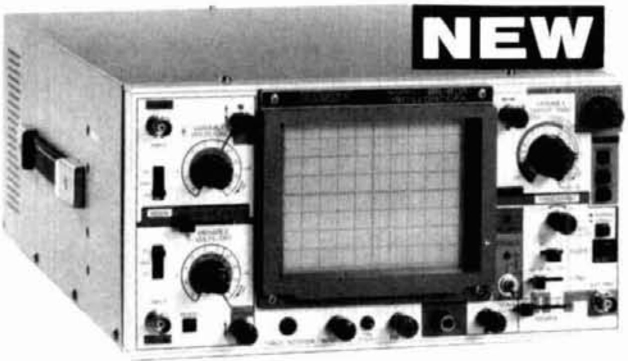

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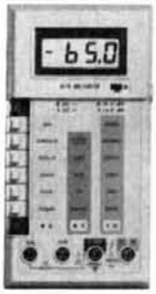

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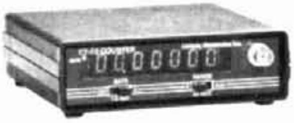

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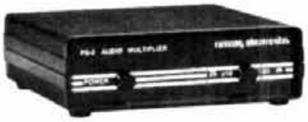

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example, the EIMAC SK-620 and SK-630 types) also offer more input to output isolation and are highly recommended for new designs.

Neutralization may be required if the amplifier gain is high and/or if the inter-electrode capacitance between the grid and plate of the tube is high. A typical bridge neutralization network is shown in **fig. 2A**. Push-pull operation requires a slightly more elaborate scheme (**fig. 2B**).

Many VHFers prefer to avoid neutralization by using grounded grid circuitry, which doesn't usually require neutralization if the inductance of the grid (and screen grid, if applicable) bypass capacitor assembly can be kept very low. Additional information on neutralization and amplifier stabilization techniques is available.[1]

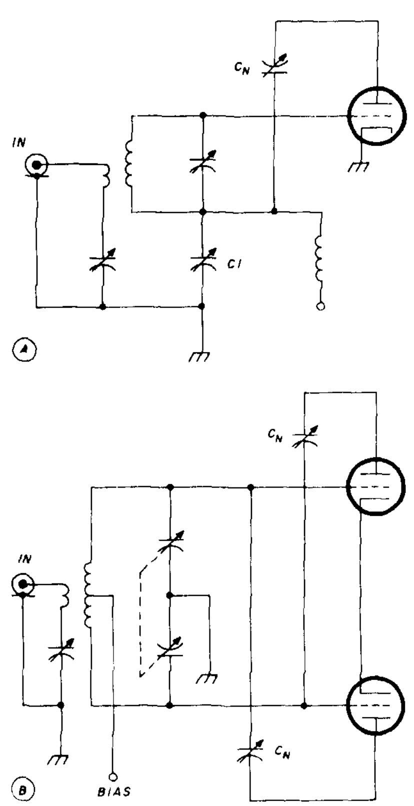

fig. 2. Typical VHF/UHF neutralization techniques. (*A*) single-ended "Bruene" circuit.[1] C1 should ideally be equal to the tube's input capacitance. C_N balances out the tube's feedback capacitance. (*B*) push-pull version.

output circuits

There seem to be as many stories about VHF/UHF transmitter output circuits as there are types. The principal types are the π network, the resonant 1/4, 1/2, or 3/4-wavelength tank, push-pull, the rectangular cavity, and variations on the latter.

On 6 meters the π network is still quite popular because it's versatile, matches a wide range of impedances, and has some inherent harmonic reduction capability (**fig. 3A**). The main problem in using this technique is that the minimum loaded *Q* in the tank circuit is determined by the output capacitance of the tube in parallel with the π network input capacitor, C1. At HF this is typically 25 to 250 pF of capacitance, but only 8 to 15 pF at 6 meters, the output capacitance of the typical tubes being used!

If a high (>15) loaded *Q* is used, the input tuning capacitor, C1, should be kept to a minimum value. This requirement can best be met by using a flapper-type capacitor. Because of their inherent inductance, vacuum variables are not recommended; standard transmitting variables usually have too high a minimum value of capacitance. Some designers have circumvented this problem by choosing a loaded *Q* based on the tube's output capacitance and tuning the inductor, L1, with a shorted turn loop.[5,6]

On 2 meters through 70 cm, most Amateur amplifier designs typically use tank circuits employing either a 1/4, 1/2, or 3/4-wavelength coax, microstrip or stripline (**figs. 3B** through **3E**). Sutherland points out that the 1/4-wavelength line is preferred because it exhibits the greatest bandwidth, and further recommends that the plate line impedance should be between one and two times the capacitive reactance (plus strays) of the tube in use.[7]

If the output capacitance of the tube is too high at the frequency of operation, a 1/4-wavelength line may be impractical because of its very short length. Then a 1/2- or 3/4-wavelength line would be usable. One advantage of the 1/2-wavelength line is that it need not be fed with the tube on one end, but may instead have the tube at the center, as shown in **fig. 3D**.[8,9,10] In this configuration, the tube capacitance is effectively divided between each side of the line. This also makes tuning and loading easier. Another advantage of the balanced 1/2-wavelength line is that the current through the tube is more evenly distributed.

Because they offer a quick way to obtain higher power with smaller tubes,[11,12] parallel amplifiers are also quite popular. However, they are often difficult to balance and are really no more efficient than a push-pull type amplifier with proper balance and output coupling.

Push-pull circuitry (see **fig. 3F**) used to be very popular, especially with the 500-watt ceramic tetrodes;[13-17] it is easier to balance because each tube is fed separately. In addition, the symmetrical output configuration inherently suppresses the second harmonic.

One factor often ignored is the current flowing through the tube itself. With the typical single-ended microstrip line approach, there is often a concentration of current through one side of the tube. This can result in decreased tube life. Consequently, at UHF and microwave frequencies, the recommended configuration is the rectangular cavity (**fig. 3G**). This configuration is particularly useful at 70 cm and above.[18] Multiple tubes can be easily paralleled for higher output power.[19] The layout of the rectangular cavity usually assures that current is uniformly distributed throughout the tube.

Output tank circuits — of which there are many — are usually chosen according to the frequency of interest, the tube type used, and the number of tubes used. In a symmetrical output tank, the use of a single tube that can deliver the required output power is probably the best choice.

output coupling

As previously mentioned, the π network, sometimes used on 6 meters, is very easy to couple to the load (**fig. 3A**).

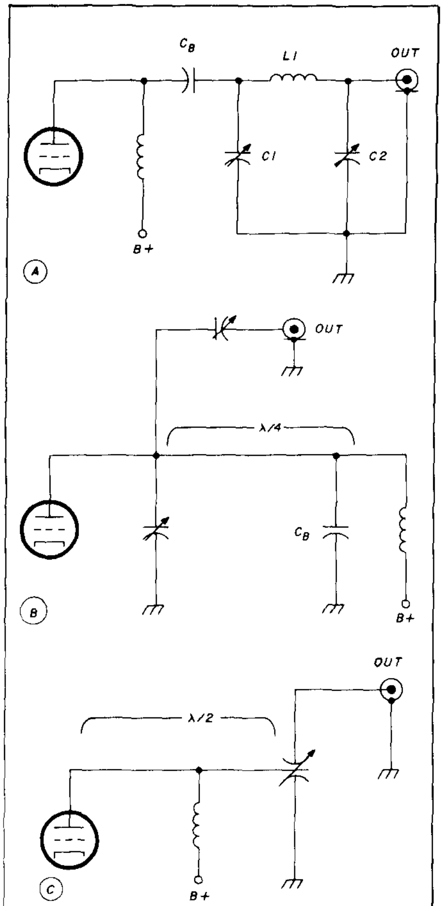
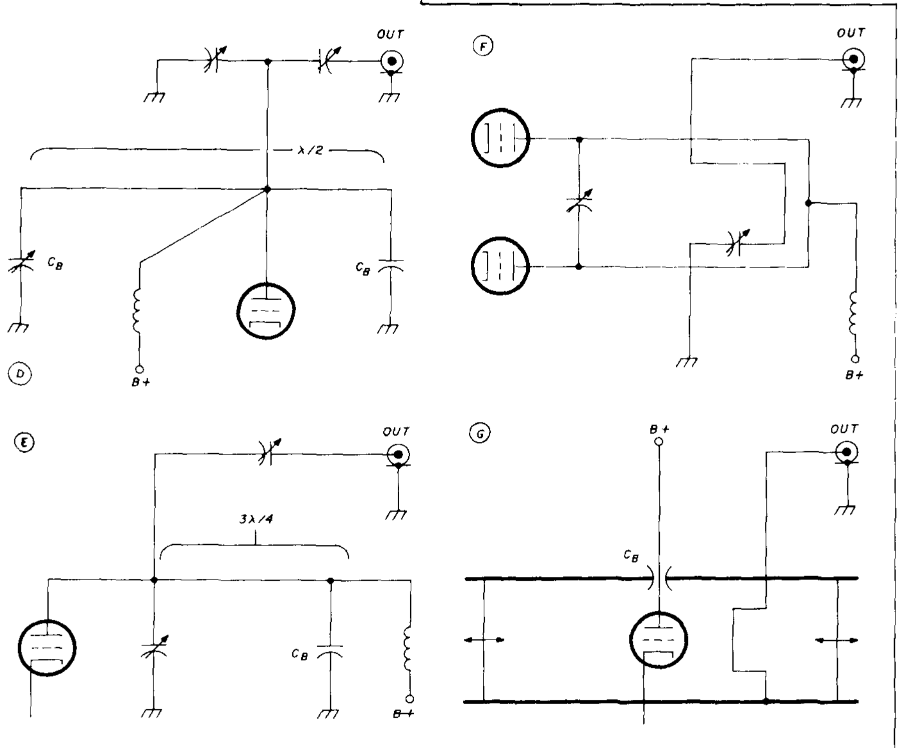

fig. 3. Typical output circuits. C_B is either a bypass or DC blocking capacitor. *(A)* π network, *(B)* 1/4 wavelength, *(C)* 1/2 wavelength, *(D)* 1/2 wavelength center-loaded, *(E)* 3/4 wavelength, *(F)* push-pull, and *(G)* rectangular cavity.

With the various types of tank circuits common in VHF/UHF amplifiers, output is often coupled by either an inductive link or by a series capacitor. While the former(**fig. 4A**) used to be popular, it has to be carefully designed and coupled to the tank circuit; this problem can be somewhat relieved by adding a series resonating capacitor in the loop (**fig. 4B**). If a loop *Q* of 2 to 5 is chosen, the capacitor will essentially operate as a loading capacitor, and the placement of the loop is not as critical. The resonant loop method is preferred because it reduces harmonic output.

Simple to design, use, and tune, the capacitance output coupling method shown in **fig. 4C** has become more popular for use in single-ended VHF/UHF amplifiers. But it can lead to problems, especially if the capacitance is increased beyond optimum, thereby causing overcoupling to the tank circuit. If this is done, harmonic output increases rapidly. Always use the least amount of coupling possible while maintaining acceptable output efficiency.

Let us not forget the push-pull configuration (**fig. 3F**). This is basically a balanced configuration, although its output network is frequently treated as unbalanced with the use of a series loop and capacitor, as just mentioned above. In fact, some users have claimed lower efficiency than expected. Efficiency can reportedly be increased by using capacitance output coupling in a balanced arrangement through a 4:1 half-wave balun.[20]

The rectangular cavity often uses a probe which is either a loop or a rod properly shunted across the cavity (**fig. 3G**). Once the proper location is found, tuning and loading can be accomplished by varying the spacing between the tube and the cavity walls, as shown in reference 19.

Finally, with the present-day designs, the proliferation of high power, and a congested VHF/UHF communications spectrum, we must pay more attention to out-of-band radiation and especially harmonics. Although a low-pass filter may reduce output power slightly. FCC regulations must be complied with. Remember also that if your amplifier has too much harmonic output power, it will appear to increase VSWR and yield false readings. In this regard, the push-pull configuration is preferred. Many harmonic suppression techniques are used, but the simplest is probably a shorted 1/4-wavelength coax stub shunted across the amplifier output connector.

summary

In this month's column I've only scratched the surface of VHF/UHF power amplifier design, emphasizing the new FCC output power regulations. Next month's column will explore the subject in greater depth, with emphasis on practical applications and construction. In the meantime, I'd suggest a review of the references listed — especially reference 1, because it deals with most of the problems prevalent in high power amplifiers and of-

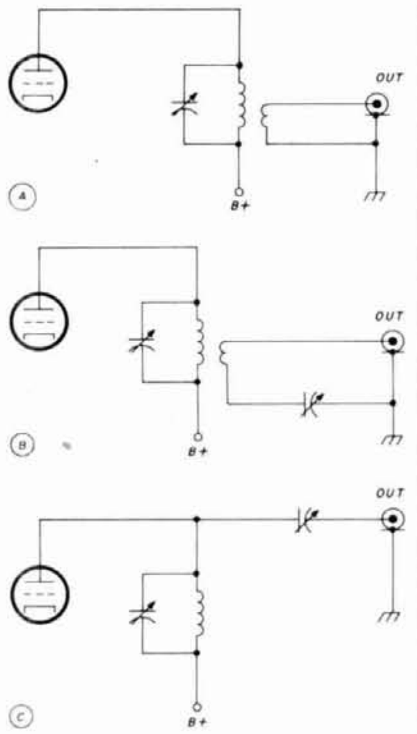

fig. 4. Typical output coupling techniques. (*A*) link coupling, (*B*) resonant link, and (*C*) capacitance probe.

fers many suggestions on solving them.

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January VHF/UHF events

January 3:	*0024 UTC, predicted peak of Quadrantid meteor shower*
January 12:	*EME perigee*
January 12-13:	*ARRL VHF Sweepstakes Contest*

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short circuit

VHF/UHF world

In W1JR's column, "VHF/UHF world: high dynamic range receivers," (November, page 97) the source of the NE41632B transistor and the balun core was correctly identified as PROTO-FAB. Since publication of the November issue, however, PROTO-FAB has changed its name to *PROTO-PARTS*.

PROTO-PARTS regrets that no telephone inquiries can be accepted. Mail inquiries, however, are welcome; please include SASE. Inquiries and orders should be addressed to PROTO-PARTS, 74 Wedgemere Drive, Lowell, Massachusetts 01852.

In the November column, certain paragraphs on page 100 were transposed. For a corrected copy of that page, send an SASE to *ham radio*, Greenville, New Hampshire 03048.

In **figs. 5** and **6** of W1JR's December column, a ground should be connected to the 5-volt, 3-terminal regulator.

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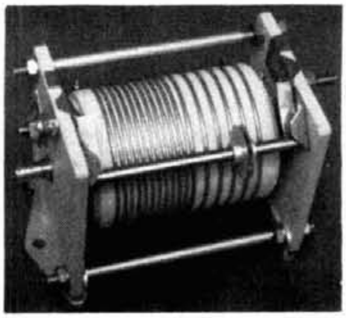

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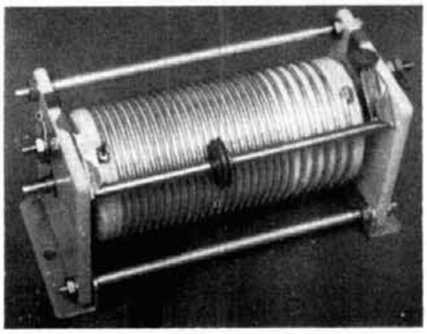

MODEL 229-0202-01
18UH at 5AMPS MAX. $44.99

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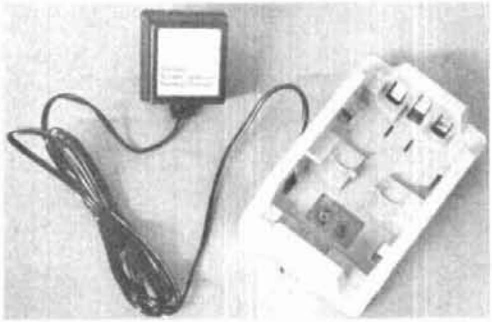

EVEREADY CHARGER $9.99

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Standard Size					Micro Size	
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4.7	15	34	68	180	6	27
5	18	36	75	200	7	33
5.1	20	38	80	220	8	36
6.8	22	39	82	240	9	51
7	24	40	100	250	12	62
8.2	25	43	110	300	15	
9.1	27	44	120	360	16	
10	27.5	47	123	470	18	
11	28	50	125	500	20	
12	30	51	140	820	22	
13	32	56	150	1000	24	

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SD1212-11	4.95	SD1307	3.00	SD1452-4	24.00	SRF1834	40.00	SD1263	15.00	SD1416	28.00	SD1542H1	170.00	SRF2919	15.00
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SD1214-11	5.00	SD1317	8.00	SD1477	35.00	SRF2147	22.00	SD1272-1	10.95	SD1428- 6084	12.00	SD1546H1	55.00	SS4152	15.00
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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
SD1220	8.00	SD1365-5	2.50	SD1484-6	1.50	SRF2371	15.00	SD1278-5	13.75	SD1430-2	18.00	SK3048	5.00	TA8563	15.00
SD1220-1	9.50	SD1375	7.50	SD1484-7	1.50	SRF2347	50.00	SD1279-1	18.00	SD1434	28.00	SL501-59	15.00	TA8564	15.00
SD1220-9	8.00	SD1375-6	7.50	SD1488	22.85	SRF2356	38.00	SD1279-3	18.00	SD1434-5	28.00	SL501-173	15.00	TA8894	15.00
SD1222-8	16.00	SD1379	15.00	SD1488-1	28.00	SRF2378	16.00	SD1281-2	8.00	SD1434-9	28.00	SM7714	5.00	TIS189	3.55
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SD1224-10	18.00	SD1380-3	1.00	SD1488-8	28.00	SRF2584	40.00	SD1283-2	10.60	SD1441	56.00	SRF395	50.00	TP1014	5.00
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
SD1225-1	15.00	SD1405	21.00	SD1511H3	75.00	SRF2741	40.00	SD1283-4	10.00	SD1444	3.25	SRF769H	20.00	TRW3	5.00
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SD1232	4.00	SD1410	18.00	SD1528-1	24.00	SRF2821	25.00	SD1290-7	15.00	SD1446	4.01	SRF1005	50.00	V4101E	20.00
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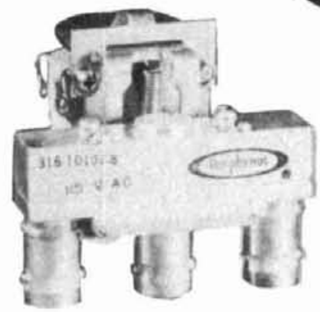

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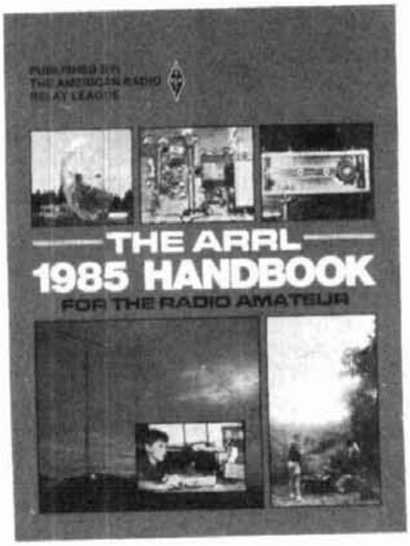

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product REVIEW

frequency counter kit

The handheld frequency counter designed by Roger Ray, (as published in the UK's *Radio and Electronics World* in November, 1982, and imported and sold by RADIOKIT) is a neat little project that shouldn't take up too much time. A valuable piece of test equipment for your ham shack, it runs off an inexpensive 9-volt battery, is quite light and easy to carry around, covers 20 Hz to 150 MHz, and has a five digit LCD display with a resolution of 1 Hz to 10 kHz depending upon range selected.

design

Ray went to great lengths to design this meter with simplicity in mind. An FC-177 LCD readout module, which also contains an OKI MSM-5527 frequency counter, is used for the display. Combining these two functions into one package greatly simplifies the construction of the counter. The FC-177 is designed to measure and display frequencies from 20 Hz to 3.999 MHz. To measure frequencies higher than 3.999 MHz, a divide-by-10 or a divide-by-100 prescaler is incorporated. To keep power consumption low, a count-and-hold technique is used when measuring frequencies above 3.999 MHz. In the LF range, 20 Hz - 10 kHz, the unit has a 50 mV sensitivity; in the MF range (10 kHz - 4 MHz) a 20 mV sensitivity; in the HF range (100 kHz - 40 MHz) a 20 mV sensitivity; in the VHF range (10 MHz - 150 MHz) a 100 mV sensitivity. Users will find that this unit, while not designed for laboratory precision, will give more than adequate readings for nearly all of their measurements.

circuit description

The input frequency is switched by the frequency range selector to one of four buffer stages. For measuring in the LF range, the signal is amplified, its frequency is multiplied by 100, and then used to control a VCO that is a part of a PLL circuit. The frequency displayed is that of the VCO with the decimal point properly positioned to account for the X100 factor. For example, an 800 Hz input signal would be changed to 80 kHz through the VCO, but counted and displayed as 800 by the FC-177.

Measuring in the MF range is within the FC-177's design. Signals are amplified and then directly fed to the display module. To measure in the HF range, signals are first amplified and then fed through a divide-by-10 prescaler (MSL-231RS) directly into the FC-177 for counting and display.

VHF range measurements are accomplished in a manner similar to that of HF signals, with the prescaler changed to divide-by-100.

As mentioned, this unit is designed for low current consumption. However, the MSL-231RS is designed with a current consumption of greater than 30 mA. To reduce current demand, a hold feature of the FC-177 is employed. Instead of constantly being counted, the incoming frequency is measured once every second. The designer calculates that this reduces consumption to below 15 mA, has little effect on accuracy, and adds several hours of battery life.

construction

The unit is built around a single-sided PC board and mounted to the enclosure by the four-way frequency switch. The FC-177 is mounted on the cover of the plastic equipment box. Parts placement is fairly straightforward, with connections to the FC-177 frequency counter module through short flexible wires. All IC's are mounted on sockets so they can be replaced with a minimum of effort in the unlikely event of failure.

Careful attention to the location of the display, four-way switch, momentary on-off switch and BNC input connector is a must if the unit is to function properly.

I'd estimate that overall time to build shouldn't be more than an evening or two, barring any unforeseen difficulties.

conclusion

At $74.95, it's really hard to beat this unit for ease of construction and usefulness in the ham shack. RADIOKIT has a number of other projects from various amateur radio publications. For a free catalog, contact RADIOKIT, Box 411H, Greenville, New Hampshire 03048.

Circle #179 on Reader Service Card.

— N1ACH

NEW products

microphone equalizer

Heath's new HD-1986 Microlizer is designed to improve the quality of transmitted speech and provide a better match between microphone and transceiver. This battery-powered microphone

equalizer fits in series with a microphone and transceiver using a standard 4-pin microphone jack and 1/4-inch phono output jack. It has continuously variable frequency controls to provide a ±12 dB (boost and cut) at 490 Hz and 2800 Hz. A gain control permits the user to increase or decrease the microphone signal fed to the transceiver for maximum efficiency and cleaner operation. The Microlizer can be bypassed to allow direct connection between microphone and transceiver by simply turning off the power switch.

For complete information and/or a copy of the current catalog, contact Heath Company, Department 150-405, Benton Harbor, Michigan 49022.

Circle #301 on Reader Service Card.

universal audio filter

Palomar Engineers has announced a new universal receiver audio filter. Model FL-4 — for SSB, CW, and RTTY — features switched capacitor filters. A 10-pole low-pass and an 8-pole high-pass can be moved anywhere in the 200-3500 Hz range to form a sharp bandpass filter at any frequency and of any bandwidth. A notch filter is also included.

It connects to the receiver phone jack and provides 2 watts of audio to drive a speaker. The on-off switch bypasses the filter when not in use. It operates from 15 VDC. The price is $139.95 plus $4 shipping. An optional 115-VAC adapter is available at $9.95.

For further information, contact Palomar Engineers, Box 455, Escondido, California 92025.

Circle #302 on Reader Service Card.

COR module

Hamtronics, Inc., has announced the COR-3, a new version of its popular COR module. Like the COR-2, the COR-3 has all the circuitry needed to control a transmitter and receiver to make a repeater, including an electronic relay to switch the transmitter on and off as a function of the receiver squelch, a tail timer, a time-out timer, an audio mixer, and a local speaker amplifier. The COR-3 also has a "courtesy beep" function, and an additional timer that allows the beep to be adjusted up to five seconds after the receiver squelch drops. Whenever a station using the repeater releases its microphone, a beep tone is heard after a short delay period. The beep indicates that the party has finished talking and the time-out timer is reset.

The price of the COR-3 kit is $58. For more information on this module and other transmitter, receiver, and control modules for building repeaters, contact Hamtronics, Inc., 65F Moul Road, Hilton, New York 14468-9535.

Circle #104 on Reader Service Card.

NEW products

signal generator

A programmable, general-purpose signal generator base-priced at $4500, is said to meet or exceed the quality and performance of units costing over $6000. The Fluke 6060A Synthesized Signal Generator accurately tests a wide variety of RF receivers, filters, amplifiers, and mixers. It covers a frequency range of 0.1 to 1050 MHz, (selectable with 10 Hz resolution) and has a switching speed less than 100 ms typical. Non-harmonic spurious products are less than −60 dBc, and harmonics are less than −30 dBc across the entire frequency range. Amplitude levels are selectable from −137 dBm to +13 dBm with 0.1 dB resolution.

For further information, contact John Fluke Manufacturing Co., Inc., P. O. Box C9090, Everett, Washington 98206.

1.3 GHz frequency counter

Digital Instruments Inc. (formerly David Electronics) of Tonawanda, New York, has announced its frequency counter (#7216). The new counter has a range of 10 Hz to 1.3 GHz and a gate time of 100 MHz 0.1 and 1.0 second as well as 1.3 GHz 0.16 and 1.6 seconds. Its display consists of eight 0.04-inch LEDs with an automatic

decimal point. The prescaler and built-in gate light all fit neatly into the small 5-1/2 x 6 x 2-inch all-metal case. Its power requirements are 105-125 volts 50/60 MHz at 3 watts with a safe input of 120 volts RMS to 10 MHz and 2 volts RMS above 50 MHz. The price is $249.95.

For additional information contact Digital Instruments, 636 Sheridan Drive, Tonawanda, New York 14450.

Circle #136 on Reader Service Card.

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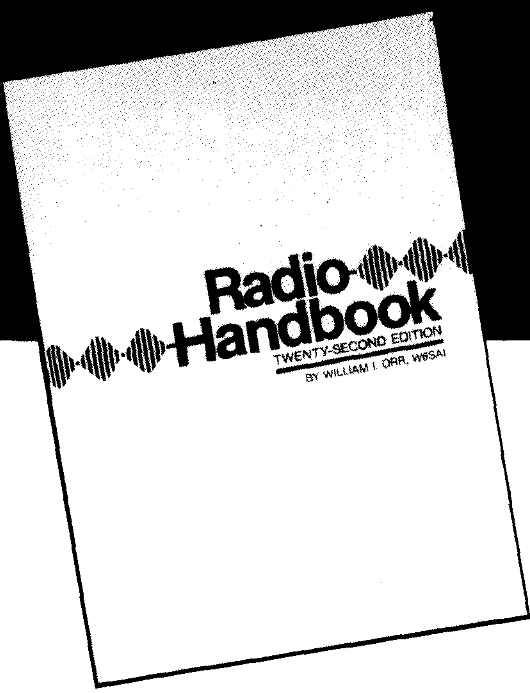

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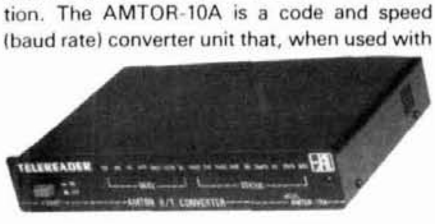
NEW products

AMTOR code converter

The AMTOR-10A unit from HAL allows transmission and reception of radio teleprinter signals with the added feature of error correction. The AMTOR-10A is a code and speed (baud rate) converter unit that, when used with



the CWR6850, allows 7-unit AMTOR code communications at 100 baud while retaining communications capability with the standard 5-unit BAUDOT code from 20 to 100 baud and the 8-unit ASCII code from 75 to 200 baud. The AMTOR-10A is designed to interface easily with the CWR6850 while retaining all the standard features of the unit, including CW operational capability. The price of the AMTOR-10A code converter is $395.00.

For further information contact HAL Communications Corporation, P.O. Box 365, Urbana, Illinois 61801.

Circle #303 on Reader Service Card.

flea market

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SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

WANTED for my hobby. Old or new foreign paper money. Will accept for free or will buy or trade. Hincke (WA6LFJ), 1854 East Bay Drive, North Bend, OR 97459.

IBM-PC RTTY ASCII/BAUDOT/CW send and receive. Split screen, buffers and features beaucoup. SASE to: E. Alline, 773 Rosa, Metairie, LA 70005.

CX7 REPAIRS. 415-549-9210.

ELECTRONICS Technology Course 34 credit, excellent for FCC license prep. $50. Johnson 6&2 $50, RCA volt ohmyst $30, RF signal Gen. $20, AF signal gen. $20. K6KZT, 2255 Alexander, Los Osos, CA 93402.

TRAVEL-PAK QSL KIT — Converts post cards, photos to QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

WANT old antenna books, handbooks, CQ, pre-1940 QST. Tad Cook, KT7H, 5519 — 12th N.E., Seattle, WA 98105.

CUSTOM MADE embroidered patches. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept. 301, 4202 N. Drake, Chicago, IL 60618.

REALISTIC 5-band solid state communications receiver, short wave and domestic, with speaker, efficient plastic dipole room antenna, with calibration charts and manual. Four years old. Used very little and has excellent world wide reception. $70 or best offer. Paul S. Bassett, Jr., RFD #2, Sutton Center, MA 01527.

HEATHKIT VF7401 — 2 meter FM digital scanning transceiver. Checked out by Heath. Asking $225. Call (217) 344-4265. Tim Lazaro, W9OEN, 203 S. Birch St., Urbana, IL 61801

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RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

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REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN. (207) 495-2215.

ATLAS 350XL owners group. Send QSL card with s/n your rig. Know anyone who repairs them? Have any technical information to share? Any questions? Rod, N5NM, Box 2169, Santa Fe, NM 87504.

CHASSIS and cabinet kits. SASE K3IWK.

SCHEMATICS: Radio receivers 1920/60's. Send name brand, model, SASE, Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001. (602) 897-2534.

AMATEUR RADIO's newspaper — Worldradio. Latest info. One year subscription (12 issues) only $10. Worldradio, 2120-B 28th St., Sacramento, CA 95818.

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WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

WANTED: Used microwave equipment. D. Mitchell, K8URL, 1 Cider Mill Lane, Upton, MA 01568. (617) 529-4638.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

VHF, UHF, ATC kits high quality professional kits from Wood & Douglas, England. Designs for RX's, TX's, Xcvr's, preamps. Catalog. Lance Lyman, K8IXZ, Tactical Electronics Corp., PO Box 1743, Melbourne, FL 32902.

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VERY in-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

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COMING EVENTS

Activities — "Places to go . . ."

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.

AFCEA Amateur Radio Luncheon sponsored by the Armed Forces Communications and Electronics Association at their 6th Western Conference held at the Disneyland Hotel in Anaheim, California, 29 through 31 January 1985. Admission free to all registrants. There is a nominal charge for luncheon tickets. Following the meal, a panel of distinguished leaders will participate in a forum on "Amateur Radio Support for the National Communications System". For additional information: John W. Browning, W6SP, 6202 Lochvale, Palos Verdes, CA 90274. (213) 544-2543.

ARIZONA: The Eastern AZ Amateur Radio Society will hold Amateur Radio license exams, Saturday, January 26, 9 AM to 3 PM, National Guard Armory, 4001 First Avenue, Safford. Prior registration deadline January 21. Send completed form 610, copy of license and $4.00 registration fee to EAARS, PO Box 402, Thatcher, AZ 85552. For information: Richard, N7DZH (602) 428-6560 7 AM to 3:30 PM.

MASSACHUSETTS: The Mount Tom Amateur Repeater Association will host its first annual indoor Flea Market, March 3, 9 AM to 4 PM, Knights of Columbus Hall, Elder Council 69, Granby Road, Chicopee. General admission $1.00. Kids and spouse free. Tables $8.00 door; $7.00 advance. Set up 8-9 AM. Food and drink. Contact Mickey Yale, N1CDR, 6 Laurel Terrace, Westfield, MA 01085. (413) 562-1027.

NEW YORK: Yonkers Electronics Auction, Sunday, January 27, 9 AM to 3 PM, sponsored by the Yonkers Amateur Radio Club. All indoors at Lemko Hall, 556 Yonkers Avenue. Admission $3.00. Children under 8 free. Inspection from 9-10 AM. Club commission on successful sales only 10% first $100; 5% remainder. Hams computer and electronic wizards, CBers bring equipment (new and used) you want to auction off. Unlimited free coffee all day. Talk in: 146.265T/146.865R, 52 direct. 440.150T/445.150R. For information: YARC, 53 Hayward Street, Yonkers, NY 10704. (914) 969-1053.

LOUISIANA: The Southeastern LA University ARC (SLUARC) and the Southeast LA ARC (SELARC) are jointly sponsoring a Hamfest, Saturday, January 19, 9 AM to 3 PM at the old men's gym on the Southeastern LA University Campus. Free admission.

THE VOLUNTEER EXAMINERS of the Grand Rapids Amateur Radio Association in cooperation with ARRL will conduct Amateur Radio exams in Grand Rapids, Michigan on the following dates: Friday, February 15, Friday, June 21, Friday, October 18, 1985 and Friday, February 21, 1986. Mail FCC Form 610, check/MO for $4.00 made out to ARRL/VEC to: ARRL/FCC Amateur Testing, c/o Mike Bottema, K8EX, 930 — 92nd Street, SE, Byron Center, MI 49315.

Operating Events — "Things to do . . ."

JANUARY 26: West Virginia QSO Party, 1700Z Jan. 26 to 1700Z Jan. 27. Single operator only. Exchange signal report, serial number and QTH (county for WV stations; state or country for others). Mail logs by Feb. 11 (include large SASE for results) to K8DS, PO Box 1694, Charleston, WV 25326

YL-OM CONTEST. Phone: Start St. February 9, 1800 UTC ends Sun. Feb. 10 1800 UTC. CW start Sat. Feb. 23 at 1800 UTC ends Sun. Feb. 24 at 1800 UTC. All licensed men and women operators throughout the world are invited to participate. Exchange station worked, QSO number, RS or RST, ARRL section or country. Entries in log must also show time, band, date and transmitter power. Logs must show claimed score and be postmarked by March 15, 1985 and received no later than April 5, 1985. Please send logs to: Marty Silver, NY4H, 3118 Eton Road, Raleigh, NC 27608, USA.

VT QSO PARTY 1985 0001Z February 2 to 2400Z February 3. Exchange VT stations send RS(T) and county (CW) two-letter county designators. Other stations send RS(T) and state, province or ARRL country. Send SASE now for official score and log sheets. Send logs/facsimiles, name, address, county (VT), n/t March 1, 1985 to: D. Nevin, KK1U, W. Hill, Northfield, VT 05663.

1985 NEW HAMPSHIRE QST PARTY, 1900Z February 2 to 0700Z February 3 and 1400Z February 3 to 0200Z February 4. Work stations once per band and mode. NH to NH QSO's allowed. Exchange signal report and QTH (county for NH stations, state, VE province or DXCC country for others). Logs must be postmarked by March 15. Include large SASE for results. Mail logs to: Great Bay Radio Assoc., PO Box 911, Dover, NH 03820

WEST COAST 160 BULLETIN SSB CONTEST, February 9 0000 GMT to February 10 2359 GMT. Single operator only. Exchange RST, QTH. Subscribers/non subscribers. Score 10 points per QSO. Multipliers: states, VE Prov. Country. Log info: date, time, rst, QTH. Send logs to: R. Koziomkowski, 5 Watson Drive, Portsmouth, RI 02871. Logs must be postmarked before March 31, 1985.

THE GUERRI REPORT

Ernie Guerri W6MGI

spectrum utilization: a challenge to technology

In almost every segment of the radio spectrum we're faced with increasing demands for frequency assignments, greater bandwidths, and increased radiated power (range/reliability). Some specific segments groan under the burden: the AM and FM broadcast bands, the 40 and 20-meters Amateur bands, VHF and UHF business radio, and the 2 GHz microwave area. Spectrum conservation and improved utilization techniques have become a high priority for practically every user of the airwaves.

Until now the methods used to squeeze more signals into the same tight space have been relatively simple: use SSB, reduce FM deviation, improve antenna directivity, and so on. But more sophisticated techniques will be needed if we're to make more efficient use of the spectrum.

Fortunately, in 1948 a thoughtful scientist — Claude Shannon at Bell Laboratories — developed a theory of information transmission that showed the relationship between speed, bandwidth, and time in a usable mathematical form. Dr. Shannon's work made possible the coding concepts that permit more effective utilization of spectrum space. There are several ways in which RF signals can share the same "space":

- time sharing
- frequency sharing
- different antenna polarizations
- coding

The goal of each of these techniques is to yield a signal-to-noise ratio that conveys information (a change of data) usable to the data sink — which is frequently a person, but could also be an unattended data terminal. Indeed, time, frequency, and polarity diversity are all forms of coding.

Modern computer technology gives us options with respect to code complexity, efficiency, and speed that were simply not available when our present modes of communications were being developed. The basic objective of data coding for spectrum efficiency is to omit as much data as possible while still conveying relevant information.

One of the more successful techniques for bandwidth reduction is being used by some computer manufacturers to permit very high resolution graphics on conventional RGB displays. This process is called bit-plane encoding. In this process a signal of 2^n possible amplitudes is transformed into n signals, each of which has only two amplitudes. The 2^n amplitudes subsequently consist of an n-bit binary word at the output of an appropriate quantizer. This process has demonstrated that it can reduce by six times the bandwidth needed to transmit high quality, full motion TV images. Manufacturers are now developing dedicated digital signal processing chips to perform the necessary bandwidth compression and S/N ratio enhancement functions.

If we remember that information requires a *change* of data, then even more bandwidth reduction is possible by further reducing, or eliminating, redundant data. AT&T adopted this technique, — called "conditional replenishment" — to make possible the "picture-phone." In this approach, a TV frame is stored in a memory and compared against a subsequent frame. Only the parts that are different are transmitted. It was found that a bandwidth of less than 100 kHz could convey an acceptable moving picture using this method. If variable persistence and digital background refresh are available at the receiving end, the data needs to be sent even less frequently. Think of how much redundant data is conveyed in the Amateur bands — background noise, non-linear distortion, excess power when band conditions are good, and so on. If Amateur SSB/FM rigs just had some digital storage and "variable persistence" audio output stages . . .!

Some TV receivers utilizing these techniques may be available by next year, but the real challenge remains a general commitment by the electronic industry to more frequent implementation of modern techniques.

Since much of the processing needed to effect significant bandwidth reduction is very complex, most Amateurs will have to wait until the chips are readily available before they'll be able to actually use these techniques in hardware.

Even more efficient than these techniques, but still years away from Amateur implementation, are mutually adaptive data links. This approach enables both ends of the link to regularly adjust their own performance to accommodate the predetermined acceptable data quality.

Although this column is reserved for discussions of technological trends with implications for Amateur Radio, the need for better spectrum management is so urgent that I can't help but offer a few comments about ways in which each of us can help assure better use of our present bands:

- Use the minimum power necessary. If conditions are good, settle for S9 on the other end. Remember that the mike gain is as useful as the volume control.
- Use filters — keep out-of-band harmonics and spurious responses to a minimum. Upgrade the internal filters (crystal/mechanical) in your rig to units with steeper skirts if they're available. Audio filters and response shaping can be useful in both the microphone and speaker circuits at your station.
- When possible, use directive antennas and keep the main lobe on the station you're working. Never mind proving that you can work Lonely Island off the back of your beam.
- Operate your rig within its limits. If you run a rig rated at 1 kW, keep it at 1 kW. Pushing it to 1500 watts will only generate distortion, won't help you at the receiving end, and will probably cause the stations on either side of your frequency to miss the opportunity altogether.

Someday the electronic capabilities I've discussed will compensate for the effects of individual operating habits on the spectrum. Until that time comes, solving the problem of effective spectrum utilization will be up to us.

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