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• WPM readout
• active mixers
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Ever notice how many of the really dominating signals you hear originate from ALPHAs? You know that a great amplifier alone doesn't guarantee a standout signal. Still, it obviously must be a big step in the right direction.

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April 1982
The ultimate team... the new Drake "Twins"

The TR7A and R7A offer performance and versatility for those who demand the ultimate!

TR7A Transceiver

- CONTINUOUS FREQUENCY COVERAGE — 1.5 to 30 MHz full receive coverage. The optional AUX7 provides 0 to 1.5 MHz receive plus transmit coverage of 1.8 to 30 MHz, for future Amateur bands, MARS, Embassy, Government or Commercial frequencies (proper authorization required).
- Full Passband Tuning (PBT) enhances use of high rejection 8-pole crystal filters.
  New! Both 2.3 kHz ssb and 500 Hz cw crystal filters, and 9 kHz a-m selectivity are standard, plus provisions for two additional filters. These 8-pole crystal filters in conjunction with careful mechanical/electrical design result in realizable ultimate rejection in excess of 100 dB.
  New! The very effective NB7 Noise Blanker is now standard.
  New! Built in lightning protection avoids damage to solid-state components from lightning induced transients.
  New! Mic audio available on rear panel to facilitate phone patch connection.
- State-of-the-art design combining solid-state PA, up-conversion, high-level double balanced 1st mixer and frequency synthesis provided a no tune-up, broadband, high dynamic range transceiver.

R7A Receiver

- CONTINUOUS NO COMPROMISE 0 to 30 MHz frequency coverage.
- Full passband tuning (PBT).
  New! NB7A Noise Blanker supplied as standard.
- State-of-the-Art features of the TR7A, plus added flexibility with a low noise 10 dB rf amplifier.
  New! Standard ultimate selectivity choices include the supplied 2.3 kHz ssb and 500 Hz cw crystal filters, and 9 kHz a-m selectivity. Capability for three accessory crystal filters plus the two supplied, including 300 Hz, 1.8 kHz, 4 kHz, and 6 kHz. The 4 kHz filter, when used with the R7A's Synchro-Phase a-m detector, provides a-m reception with greater frequency response within a narrower bandwidth than conventional a-m detection, and sideband selection to minimize interference potential.
  Front panel pushbutton control of rf preamp, a-m/ssb detector, speaker ON/OFF switch, i-f notch filter, reference-derived calibrator signal, three agc release times (plus AGC OFF), integral 150 MHz frequency counter/digital readout for external use, and Receiver Incremental Tuning (RIT).

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- FREQUENCY FLEXIBILITY. The TR7A/R7A combination offers the operator, particularly the DX'er or Contestor, frequency control agility not available in any other system. The "Twins" offer the only system capable of no-compromise DSR (Dual Simultaneous Receive). Most transceivers allow some external receiver control, but the "Twins" provide instant transfer of transmit frequency control to the R7A VFO. The operator can listen to either or both receiver's audio, and instantly determine his transmitting frequency by appropriate use of the TR7A's RCT control (Receiver Controlled Transmit). DSR is implemented by mixing the two audio signals in the R7A.
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Tell 'em you saw it in HAM RADIO!
During my ham career, I have always been an enthusiastic member of the DX fraternity. I’ve been through more sunspot cycles than I care to remember, always trying to increase my DXCC country total, and I have managed to accumulate a fairly respectable number of confirmed contacts. All this took place when I was in California. There I had all the trappings considered necessary to compete with the big guns in kilowatt alley and was able to hold my own pretty well in the pile-ups. However, things have changed — and for the better, I might add.

My move from California to New Hampshire required that I put all my ham gear, some 35 years’ accumulation, in storage, where it still remains. But strangely enough, I don’t really miss it. I’ve rediscovered a new and far more challenging pursuit, one that I followed only casually in California — the world of QRP, or low-power operation.

With the kind help of Skip Tenney, I was on the air with borrowed equipment and a longwire antenna shortly after I arrived in New Hampshire. The gear was nothing exotic — just a good high-frequency transceiver with adjustable power output and a few accessories. My experience with this borrowed rig, operating from a good location (for a change), has been most rewarding. It’s altogether a different game chasing DX stations when you don’t have a big antenna and a kilowatt amplifier.

The term QRP is from the international list of Q-code signals and means reduce power. QRP operation in the Amateur bands has several definitions, depending on whom you talk to. To some, anything below 100 watts is QRP. However, true QRP is generally recognized as an input power of 10 watts or less, and 1 watt or less is called QRPp. With a good antenna and only a few milliwatts, tremendous distances can be covered in the high-frequency bands. But this kind of achievement is possible only with skillful operating and, above all, patience and perseverance.

A certain etiquette and some unwritten rules prevail among the QRP brethren. These have evolved over the years and are generally accepted by serious operators. For example, it’s not considered good form to make contact with high power then ask the station to listen for your QRP signals. Another practice frowned upon is making prior arrangements with a station for a QRP contact. It is also considered unethical to have a friend ask the desired station to “stand by for W6BLANK using QRP.” I tried a few of these tactics during my early days of QRP operating before I learned how to behave. True, I made some contacts, but they didn’t seem as rewarding as those made according to the rules.

I’ve heard some operators say that it isn’t very productive for a QRP station to call CQ. This may be true during the early evening hours when the bands are crowded; however, I’ve found that if I get on the air late at night, find a clear spot, and call “QRP” several times followed by my callsign, it is possible to obtain replies. Furthermore, many foreign stations enjoy working U.S. QRP stations, as I have discovered on 40 meters, especially near 7025 and 7060 kHz. Overseas operators also do a lot of listening, and I was surprised one evening to find that stations were actually calling me after I finished a QSO with a Belgian station. What an experience! I’d actually created a mini pile-up.

It doesn’t take much equipment to get a QRP station on the air. A simple but stable VFO will serve as a transmitter. A selective receiver is important, and the antenna should be as efficient as you can possibly make it. Although a gain antenna is an asset, a simple, well-matched dipole will give good results in QRP work.

If you want to experience a real sense of achievement resulting from some rigid operating discipline, I highly recommend QRP work. Now, I wonder who will be the first U.S. ham to earn the Worked All Continents award with 500 milliwatts on the new WARC bands?

Alf Wilson, W6NIF editor
Pendergast’s return

Dear HR:

The article on traps in the October issue is excellent stuff. Very seldom do I get so interested in the first article that I get stuck there to the temporary exclusion of the rest of the magazine. I normally scan the whole issue first and start with the one that interests me most. The return of the magazine to an emphasis on technical excellence is most encouraging. Congratulations on a good job.

A couple of small points. In the July issue, page 52, 0.707 of the power has somehow become half the power; 0.707 of the field strength in volts would be half the power.

In the August issue I couldn’t help feeling a foreboding that Pendergast had snuck back into Bill Orr’s material, and sure enough when I arrived at work there was a note from a colleague saying, “If you are writing to ham radio for Pete’s sake complain about Pendergast’s return.” I am hereby passing on that message, with a fervent plea from me that Pendergast stay dead. Bill Orr is often interesting, but the dialogue sometimes gets in the way. I know many readers that feel this way, and can honestly say that I don’t know anyone who likes Pendergast. It was most encouraging to see no mention of P. in October!

To come back to the Gary O’Neil piece on traps, I think it would interest all readers if you could print a note from Gary on the kind of power his RG8/U traps can handle, and how far off frequency you can operate before the trap heats up because of circulating current. And here is a question. When feeding a trap dipole directly with coax (no balun), I find the trap in the side connected to the braid gets hot enough to melt the encapsulant, while the other one stays cool. Why? (Swapping them confirms that it is the position, not the trap, which causes the heating.)

Bob Eldridge, VE7BS
Pemberton, B. C.

lifeline SAR

Dear HR:

We are a volunteer, non-profit public service organization headquartered in Madison, Indiana, and we’re looking for members from all over the world. Our prime objective is to form local chapter units for those areas that don’t already have a qualified search and rescue (SAR) organization. There are, unfortunately, all too many such areas.

Our organization currently possesses members in 27 divisions (25 states plus Canada and the Republic of Singapore). We rely on Amateur Radio as a back-up communications network and for any long-range traffic we may have. Our primary means of communications is our own VHF-fm commercial radio system.

In addition to looking for members, we are also looking for donations of radio equipment (both VHF-fm commercial and Amateur) to help us build our organization and its chapters up to an operational level. All donations (cash and otherwise) are 100 percent deductible for federal income tax purposes.

Everyone is eligible to apply for membership. Those who inquire are asked to enclose an SASE to help with the rising costs of mailing. All inquiries will be answered promptly. For further information, please write Lifeline Search and Rescue, P.O. Box 237, Milton, Kentucky 40045.

Jeff E. Howell, CEMT (WBSPFZ)
Executive Director
Lifeline SAR

a card from Frenchy

Dear HR:

Bill Orr’s article in the August issue brought back fond memories of DX for a then-young Amateur. After I got my ticket in mid 1938, it took me almost a year to scrounge up the parts for my homemade superhet and 6A6-PP6L6 transmitter.

By that time World War II had drawn the curtain over much of the DX, but nevertheless, my spare and weekend hours were spent listening for something “rare.”

One early morning in December of 1939, there was the same KF6ROV mentioned in Bill’s article, calling CQ; with trembling hand I called him, and “Frenchy” Paquette in Pago Pago came back to my call. This is one of my most prized QSLs.

Thanks again to Bill and to you for a fine magazine.

F. V. Sprick, W2LPV
Clifton, New Jersey

kilo foxtrot Charlie?

Dear HR:

In regard to the comments by Mr. H.B. Mouatt, W6BQD, in the January, 1982, issue, I am in full agreement. The use of Q signals on phone has long been a source of improper operation, insofar as they are intended for use on CW. If one is going to stop transmitting, simply say so, not QRT. I am sure there are others, but this is one I hear a lot. This practice seems to be common among newcomers and old-timers as well.

The use of one’s own phonetics for one’s own call, however, seems to be perfectly proper to me. I have heard both U.S. and foreign Amateurs use their own personal phonetics many times. In fact, one very good friend, Vic Clark, W4KFC (past ARRL VP and Roanoke Division Director), would probably be unrecognizable to me on the air if he were to say Whiskey Four Kilo Foxtrot Charlie instead of W4 Kentucky Fried Chicken.

Ed Redington, Jr., N4AGS
Ashland, Virginia
MFJ Super Keyboards

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Get text buffer, programmable and automatic message memories, error deletion, buffer preload, buffer hold, plus much more.

MODE 1: CW

The 256 character (50 for 494) text buffer makes sending perfect CW effortless even if you “hunt and peck.”

You can preload a message into the buffer and transmit when ready. For break-in, you can stop the buffer, send comments on key paddles and then resume sending the buffer content.

Delete errors by backspacing.

A meter gives buffer remaining or speed. Two characters before buffer full the meter lights up and the sidetone changes pitch.

Four programmable message memories (2 for 494) give a total of 256 characters (30 for 494). Each message starts after one ends for no wasted memory. Delete errors by backspacing.

To use the automatic messages, type your call into message A. Then by pressing the CO button you send CO CO DE (message A).

The other automatic messages work the same way: CO TEST DE, DE, ORZ.

Special keys for KN, SK, BT, AS, AA and AR.

A lot of thought has gone into human engineering these MFJ Super Keyboards.

For example, you press only a one or two key sequence to execute any command.

All controls and keys are positioned logically and labeled clearly for instant recognition.

Pets are used for speed, volume, tone, and weight because they are more human oriented than keystroke sequences and they remember your settings when power is off.

Weight control makes your signal distinctive to penetrate QRM.

MODE 2 & 3 (RTTY): BAUDOT & ASCII

5 level Baudot is transmitted at 60 WPM. Both RTTY and CW ID are provided.

Carriage return, line feed, and “LTRS” are sent automatically on the first space after 63 characters on a line. This gives unbroken words at the receiving end and frees you from sending the carriage return. After 70 characters the function is initiated without a space.

All up and down shift is done automatically. A downshift occurs on every space to quickly clear garbled reception.

The buffer, programmable and automatic messages, message memory, PET control (keys your rig) are included.

The ASCII mode includes all the features of Baudot. Transmission speed is 110 baud. Both upper and lower case are generated.

MODE 4: MEMORY KEYER

Plug in a paddle to use it as a deluxe full feature memory keyer with automatic and programmable memories, iambic operation, dot-dash memories, and all the features of the CW mode.

MODE 5: MORSE CODE PRACTICE

There are two Morse code practice modes. Mode 1: random length groups of random characters.

Mode 2: pseudo random 5 character groups in 6 separate repeatable lists (with answers).

Insert space between characters and groups to form high speed characters at slower speed for easy character recognition.

Select alphanumeric or alphanumeric plus punctuation. You can even pause and then resume.

MORE FEATURES

Automatic incrementing serial number from 0 to 999 can be inserted into buffer or message memory for contests.

Repeat function allows repetition of any message with 1 to 99 seconds delay. Lets you call CO and repeat until answered.

Two key lockout operation prevents lost characters during typing speed bursts.

Clock option (496 only) send time in CW, Baudot, ASCII. 24 hour format.

Set CW sending speed before or while sending.

Tune switch with LED keys transmitter for tuning. Tune key provides continuous dots to save finals. Built-in sidetone and speaker.

PTT (push-to-talk) output keys transmitter for Baudot and ASCII modes.

Reliable solid state keying for CW: grid block, cathode, solid state transmitters (.300V, 10 ma Max., +.300V, 100 ma Max). TTL and open collector outputs for RTTY and ASCII.

Fully shielded. RF proof. All aluminum cabinet. Black bottom, eggshell white top. 12”x7”x1”/4” (front) x3/8” (back). Red LED indicates on. 9-12 VDC or 110 VAC with optional adapter.

MFJ-494 is like MFJ-496 less sequential numbering; repeat/delay functions. Has 50 character buffer, 30 character message memory. Clock option not available for MFJ-494.

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MFJ-53 AFSK Plug-In Module. 170 and 850 Hz shift. Output plugs into mic or phone patch jack for SSB rigs and AFSK with FM or AM rigs. $39.95 (+ $3).

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MFJ-61 Clock Module (MFJ-496 only). Press key to send time in CW, Baudot or ASCII. 24 hour format. $29.95 (+ $3).

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April 1982
THE FCC HAS ACTED IN THE JAMMING cases involving the Maritime Mobile Services Net. In a release dated January 28, 1982, the Commission revoked the license of Leonard M. Boucher, K4MME, of Cantonment, Florida, and at the same time suspended for one year the license of Gerard Morin, W1GM, of Sanford, Maine. In their "Order of Revocation and Suspension" the Commission found that, between August, 1980, and June, 1981, Boucher and Morin operated in a split frequency scheme designed to deliberately and maliciously interfere with radio nets operating on the frequency of 14.313 MHz.

Boucher received the greatest amount of criticism and criticism from the FCC. They noted that on February 17, 1981, they monitored Boucher requesting a station to move down with him to the frequency on which net participants were operating, and on February 18th he was again heard to criticize the operator of a station for not moving directly on top of the frequency occupied by the net. Other instances were cited as well.

A5 Magazine Has Proposed SSTV operating frequencies based on the recent approval of FCC Docket 80-252 which expands SSTV/FAX operations in the General Class phone bands. A5 proposes the following frequencies: 3.990, 7.290, 14.340, and 21.440 MHz with all movement upward only utilizing the top 10 kHz of band edges. Any comments should go to Mike Stone, N8WCD, A5 Magazine, P.O. Box H, Lowden, Iowa 52253. Mike will also be on hand at the 1982 Dayton Hamvention's ATV Forum at 2:00 PM on Saturday, April 24th.

Arnold Verdon, W6LJ, a familiar sight at many hamfests as he manned the Collins booth, passed away February 1. His easy going, likable manner, along with his extensive first-hand knowledge of post-war Collins products, combined to make him a favorite in the Amateur community. He will be missed as a professional but even more as a good friend.

The United States Will Soon Have A "Woodpecker" operating in the 5 to 35 MHz spectrum. Like its Russian counterpart, it also will be an over-the-horizon radar system. It will be located near the city of Moscow, Maine. Built by General Electric Corporation under contract to the U.S. Air Force, it will have an ERP of 1.2 megawatts. The system is in its final testing stage, with full-time operation scheduled to begin sometime in 1982. What effect our woodpecker will have on Amateur Radio HF communications is unknown, but the military has promised to work with the Amateur community to reduce its impact.

Dr. Norm Chalpin, K6PGX, reports that Wednesdays (UTC) are Experimental Days on the Soviet R/S Satellites. During that time, the birds are reserved for previously arranged uplink transmissions. Chalpin also informs us that all passbands on the R/S transponders are linear; that is, USB in on 2 meters produces USB out on 10.

Satellite Telecasters Are Going to Great Lengths to try to end the pirating of their subscription services. Premier Communications Network of San Jose, California, placed an advertisement in a recent edition of the San Jose Mercury newspaper appealing to area residents to trade in their "illegal" pirate antennas on a "legal" installation from their company for Home Box Office service. The newspaper ad contains a reprint of findings regarding the secrecy of communications provisions of Section 605 of the Communications Act, and it then offers amnesty along with guaranteed service to anyone answering the ad and becoming a subscriber.

Swift Reaction Has Come From The Area's Local Amateurs. Writing in the Bay Area 220 Group newsletter, Pres. Jerry Lahtinen, W6TTU, stated that his organization considers HBO to be the real pirates since that service utilizes publicly owned spectrum, and he has called for the abolition of HBO entirely.

An On-Site Alert at the R.E. Ginna Nuclear Plant January 25th saw hams from Rochester, Monroe County, and Wayne County (New York) put themselves on alert as well. Fortunately, the problem did not require evacuation of the area and the alert was downgraded just a few hours later. Ironically, this "real alert" followed only four days after a mock Nuclear Evacuation Drill had been held in the same area.

Re-Balloting in Two Disputed ARRL Director's Elections show that the incumbents, William Stevens, W6ZN (Pacific Division), and Leonard Nathanson, W8RC, (Great Lakes Division), have been re-elected. According to League headquarters, in each case the winner's margin was almost 2:1 over his opponent.

On January 4th the FCC began issuing the new General Radiotelephone license to replace the 1st and 2nd class phone licenses that were discontinued last year. The order initiating the change came last August but could not be implemented until new license forms had been printed. Current holders of 1st or 2nd class tickets should note that they are valid until expiration, but when renewed they will become General Radiotelephone licenses. All new licenses will be of that class as well.
The right design — for all the right reasons. In setting forth design parameters for ARGOSY, Ten-Tec engineers pursued the goal of giving amateurs a rig with the right features at a price that stops the amateur radio price spiral.

The result is a unique new transceiver with selectable power levels (convertible from 10 watts to 100 watts at the flick of a switch), a rig with the right bands (80 through 10 meters including the new 30 meter band), a rig with the right operational features plus the right options, and the right price for today’s economy—just $549.

Low power or high power, ARGOSY has it. Now you can enjoy the sport and challenge of QRPP operating, and, when you need it, the power to stand up to the crowds in QRM and poor band conditions. Just flip a switch to move from true QRPP power with the correct bias voltages to a full 100 watt input.

**New analog readout design.** Fast, easy, reliable, and efficient. The modern new readout on the ARGOSY is a mechanical design that instantly gives you all significant figures of any frequency. Right down to five figures (± 2 kHz). The band switch indicates the first two figures (MHz), the linear scale with lighted red bar-pointer indicates the third figure (hundreds) and the tuning knob skirt gives you the fourth and fifth figures (tens and units). Easy. And efficient—so battery operation is easily achieved.

The right receiver features. **Sensitivity** of 0.3 µV for 10 dB S+N/N. **Selectivity:** the standard 4-pole crystal filter has 2.5 kHz bandwidth and a 2.7:1 shape factor at 6/50 dB. Other cw and ssb filters are available as options, see below. **I-f frequency** is 9 MHz, i-f rejection 60 dB. **Offset tuning** is ± 3 kHz with a detent zero position in the center. **Built-in notch filter** has a better than 50 dB rejection notch, tunable from 200 Hz to 3.5 kHz. An optional noise blanker of utes on all bands. **3-function meter** shows forward peak power on transmit, SWR, and received signal strength. **PTT** on ssb, **full break-in on cw. PIN diode antenna switch. Built-in cw sidetone with variable pitch and volume. ALC control** on “high” power only where needed, with LED indicator.

**Automatic normal sideband selection plus reverse. Normal 12-14V dc operation plus ac operation with optional power supply.**

The right styling, the right size. Easy-to-use controls, fast-action push buttons, all located on raised front panel sections. New meter with lighted, easy-to-read scales. Rigid steel chassis, molded front panel with matching aluminum top, bottom and back. Stainless steel tilt-up bail. And it’s only 4” high by 9½” wide by 12” deep (bail not extended) to go anywhere, fit anywhere at home, in the field, car, plane or boat.

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**Model 525 ARGOSY — $549.** Make the right choice, ARGOSY— for the right reasons and low price. See your TEN-TEC dealer or write.
Updated version of
a design published
previously in \textit{ham radio}

Front and rear views of the 220-MHz stripline kilowatt amplifier, which is enclosed in two mated chassis. Two optional designs are available, one for triode and one for tetrode tubes.

The 144-MHz stripline kilowatt amplifier described in the October, 1977, issue of \textit{ham radio} has been constructed and operated by a number of Amateurs, both from the information in the article and from parts kits available from ARCOS. A 220-MHz version of this amplifier was built and tested during 1977 and has been reproduced several times. One of these models has been in service on a 220-MHz repeater for over two years. This same amplifier was borrowed from the repeater and used during VHF/UHF tests over the past two years by the W2SZ/1 (Mount)

*ARCOS, Amateur Radio Component Service, Box 546, East Greenbush, New York 12061. All parts for the 220 amplifier and power supply are available.

By F. J. Merry, W2GN, 35 Highland Drive, East Greenbush, New York 12061
Top view of the stripline amplifier with cover removed showing the load and tuning flappers.

Bottom view of the triode amplifier with cover removed.

Socket assembly for the 8874 triodes, which may be installed onto the amplifier chassis in place of the tetrode sockets.

Front view of the power supply with cover removed.

Bottom view of the tetrode amplifier with cover removed.

Front view of the power supply for the stripline amplifier with cover installed.
Greylock) group. Experience has been favorable in all respects.

The experienced builder, especially if he has built the 144-MHz model, will find no difficulty in building and operating the 220-MHz version. With EME and tropo activity on 220 MHz on the increase, this amplifier is a good candidate where there is a need for reliable high-power operation.

Similar to the 144-MHz amplifier in chassis dimensions and other respects, the 220-MHz version uses a quarter-wave plate line and a coil-simulated half-wave grid line. Except for the plate-line mounting screws, the location of the high-voltage feedthrough capacitor and a hole for a bushing for the plate tuning flapper drive string, chassis punching is identical to the drawings for the 144-MHz amplifier as described in reference 1. Originally described by K2R1W (reference 2) for a 432-MHz stripline amplifier, this type of chassis construction has proven adaptable to not only 432, 220, and 144 MHz but also to an equally successful 50-MHz version using a pi-network output circuit with inductive tuning. The 50-MHz model will be described in a subsequent article.

The amplifiers mentioned above can use any of the ceramic tetrodes such as the 4CX250R, 4CX250B, or 8930, further illustrating the flexibility of this type of chassis construction. The 8874 triodes can also be used in a grounded-grid application by installing a mounting plate for the 8874 sockets in place of the individual EIMAC 630A sockets. The triode amplifiers have the advantage of not requiring the critical screen supply voltage and the disadvantages of higher tube cost and higher drive power requirement.

design

Referring to the schematic, (fig. 1), the quarter-wave plate line is tuned and loaded by flapper capacitors. The plate blocking capacitor is a Teflon sandwich at the cold (rf) end of the line. The grids are connected by a strap to which is connected the grid coil. A 1k, 2-watt resistor is used for a grid choke (class AB1 operation — no grid current). A 1k, 2-watt resistor in series with a 1000-pF capacitor is connected to ground from each grid. These two loading resistors increase the stability margin of the amplifier and desensitize it so that it can be driven to 500 watts output with about 10 watts of drive. Additional resistors may be added depending on the drive power available. If the grid load resistors are omitted, the amplifier is stable but it becomes extremely power sensitive.

A further examination of the schematic reveals an optional rf output indicator circuit and 660-MHz and 680-MHz and
440-MHz harmonic traps in the rf output, surge voltage protectors (SVPs) on the screen leads, and the usual lead filtering to keep the rf locked into the grid box and plate compartment. Metering is with a single 0-1 mA meter and a six-position switch with appropriate metering resistors for plate current, plate voltage, individual screen currents, grid current, and relative rf output. Thus, all important operating parameters are monitored.

A nonlocking reversing switch facilitates reading negative values of screen and grid currents, which are normally experienced with tetrode amplifiers. I recommend that an rf output wattmeter be used instead of the relative rf output indicator circuit. Proper adjustment of the plate tuning controls is difficult unless both plate current and the rf power output can be observed simultaneously. Other dc metering arrangements may, of course, be employed, including enclosing the meter circuit in a separate box at the operating position with the amplifier remotely located or by locating the metering circuit in the power-supply chassis.

The surge-voltage protectors in the screen circuit will ground the screens if the screen voltage rises above 470 volts. This protection is important and should be provided on all tetrode amplifiers. Emission effects in ceramic tetrodes will cause the screen to go negative under certain operating conditions. When this condition occurs, the screen voltage rises, causing higher plate current. The tubes can go into a runaway condition unless the amplifier is shut down. The surge voltage protectors prevent the runaway condition by automatically reducing the screen voltage to a very low value. Once one of the SVPs has fired, it’s usually necessary to shut off the power. Power may be restored after a pause to let the capacitors discharge to a point below the sustaining voltage of the SVP that fired. A small saving can be realized by using only one SVP connected to the screen-supply lead.

Two other factors are involved in foolproof operation of the screen circuit:

1. The screen power supply must be provided with a current-limiting resistor so that the current doesn’t exceed about 100 mA when one of the SVPs fires.

2. The power supply must have a resistor, from screens to negative, which is of a value low enough to provide a sink current of at least 40 mA. This feature provides a path for the negative screen current so that the screen voltage will hold at the regulated value.

With the above features provided in the screen circuit, the tetrode amplifier will perform as reliably and smoothly as a triode amplifier.

construction*

Chassis punching and drilling follows the same pattern shown by the drawings and as discussed in the 2-meter amplifier article. Variations are listed below.

Plate-line mounting holes. The five holes in the right end (facing front) of the upper chassis (used to mount the 2-meter plate line) are not required. The 3/16 inch (5 mm) holes that fasten the 1-1/4 meter plate line to the chassis are located by setting the line in place with the tubes installed.

With the above features provided in the screen circuit, the tetrode amplifier will perform as reliably and smoothly as a triode amplifier.

*See the appendix before proceeding further.
Mounting the high-voltage feedthrough capacitor. The 7/16 inch (11 mm) hole for the high-voltage feedthrough capacitors is located 2 inches (5 cm) to the right (facing the front of the amplifier) of the cold end of the plate line and in line with the rear edge of the plate line. The rf choke is mounted between a lug on the plate line and another lug on the high-voltage feedthrough capacitor.

Plate-tuning flapper. The plate-tuning flapper is mounted on an aluminum block as in the 2-meter amplifier. It is shorter than the 2-meter flapper. The drive cord is connected to the tuning control through a pulley inside the grid box, fig. 2. A knob shaft lock and a steel shaft provide the plate tuning adjustment, fig. 3.

An alternative method of installing and controlling the plate tuning flapper is to secure it to the front of the upper chassis wall and adjust it with a 1/4 inch (6.4 mm) threaded rod from the top of the chassis — a simple and positive method of control, the same as that used to control the plate load flapper.

Plate line. The construction of the plate line is shown in figs. 4 and 5. The parts of the line must be free of burrs to avoid puncturing the Teflon insula-
tion. Assembly of the line must be accurate so that the screws holding the clamping bars are centered in the holes in the plate line and the underside of the plate line is 1-1/2 inches (38 mm) from the floor of the upper chassis.

Grid circuit. Details of the grid circuit are shown in fig. 6. Miniature capacitors (20-pF) may be substituted for the butterfly caps. The advantage of the costly butterfly cap is that no bearings are included in the rf path. The 2-11 pF butterfly caps will require a10-pF padder mounted across the two stators.

triode amplifier

If the 8874 triodes are the tube choice, the sockets can be mounted on a brass plate, which is then installed onto the amplifier chassis in place of the EIMAC 630A sockets used for the tetrodes, fig. 7. There’s no need to change dimensions of the plate line provided it’s for the 4CX250 tubes. The grid line becomes the cathode line, and an rf choke is substituted for the 1k, 2-watt grid resistor. The grid load resistors are also omitted. See the schematic of the triode version, fig. 8.

Note that the triode amplifier has two meters mounted on the chassis. The right-hand meter reads plate current; the left-hand meter reads grid current. By operating a nonlocking meter switch, plate voltage is read on the grid-current meter. (I assume that an rf wattmeter will be used to monitor power output.)

Metering and cathode resistors are mounted on the right-hand end of the lower chassis. The zener bias diode and the control connector are mounted on the rear of the lower chassis. An ungrounded contact closure on the control connector is required to establish operating bias.

harmonic traps and rf indicator pickup

The harmonic traps (440 MHz and 660 MHz) in the rf output are installed in a small box (Pomona 2428) mounted on the rear of the amplifier, fig. 9. Alternatively, these traps can be installed in a box with coaxial connectors on each end (Pomona 2411). This box can be inserted in the output line at the rf output connector or immediately following the output wattmeter. The adjustment of these traps is best done

---

fig. 6. Grid-circuit details. The butterfly caps will require a 10-pF padder mounted across the two stators.
while observing the level of the harmonic output on a spectrum analyzer.

The rf pickup assembly is also illustrated in fig. 9. The amount of rf pickup is obtained by adjusting the position of the lug, which is mounted on top of the output flapper, with respect to the lug on the stand-off insulator.

The rf pickup assembly may be omitted if you plan to have an rf wattmeter in the output circuit. As I mentioned previously, it's desirable to have an rf wattmeter to monitor output and to achieve the best adjustment of plate tune and load controls for maximum efficiency.

**assembly**

The following sequence is suggested for assembly:

1. Mount and wire all parts on the lower chassis.

2. Fasten the grid box to the upper chassis and install the sockets, feedthrough capacitors, SVPs and BNC input connector. For the tetrode amplifier, orient the sockets so that terminals 1 and 3 are opposite their respective feedthrough caps. The sockets for the triode amplifier are mounted so that the heater terminals are positioned between their respective feedthrough capacitors on the cathode box.
3. Install the high-voltage feedthrough capacitor and the rf output pickup assembly.

4. Fasten the upper and lower chassis together and complete the wiring interconnections.

5. Mount the butterfly caps in the grid box (fig. 6) and install the tune and load controls.

6. Install the grid line, grid coil and resistors as in fig. 6.

7. Install the plate-tuning flapper, pulley, dial, shaft and bearing. Tie the nylon line to the flapper before mounting the flapper.

8. Install the tubes temporarily and put the plate line (previously assembled) in place. Work the finger stock over the tubes very carefully. Make sure everything lines up. Mark the mounting holes at the end of the plate line, then remove the plate line and tubes and drill the mounting holes for the plate line. Reinstall tubes and plate line.

9. Connect the plate rf choke.

10. Install the output flapper, rf grounding choke and RFO assembly.

11. Assemble the top plate screen and vent plate and the threaded bushing for flapper adjustment. Put the tubes in place and put the Teflon chimneys in position in the vent plate. Fasten the top plate in place.

12. Recheck wiring and fasten the bottom plate.

The amplifier is now ready for test.

blower

The blower may be mounted onto the air-intake plate on the rear of the amplifier or it may be hose connected. The Dayton model 4C012A specified in the 2-meter construction article is satisfactory for normal operation. Figs. 10 and 11 give the dimensions for air-intake plates for higher-power blowers;

fig. 10. Air intake cover-plate details, Dayton 4C443 blower mount.
that is, the Dayton 4C443 (100 CFM) and the Rotron V537A2R4 (160 CFM). Generally speaking, the more air flow the better, so the choice of blowers is usually a compromise based on noise level and price. The noise can be reduced appreciably by control circuits that reduce blower speed during standby periods. To make the blower operation foolproof (the amplifier will fail in less than a minute without air), an air switch can be mounted in the output air stream of the blower, or a pressure switch can be mounted on the upper chassis. These switches can be connected to shut down the power supply or bias the amplifier to cutoff (see next section).

power supply

The power supply was given a rather brief treatment in the previous article. There's a tendency to consider the power-supply design and construction as less significant than the amplifier. This can be a mistake, especially for the tetrode screen supply which, as discussed previously, has critical requirements for successful operation of tetrodes. This time, to provide background on its operation, the power supply is described in some detail (see the schematic, fig. 12).

Features of modern high-voltage power supply

fig. 11. Air intake cover-plate (Rotron V537A2R4).

fig. 12. Power-supply schematic for the tetrode stripline kilowatt amplifier.
design include compactness: 12 inches wide by 7 inches deep by 10 inches high (30.5 by 17.8 by 25.4 cm) and light weight (37 pounds, or 17 kg). All output voltages and other features to operate the amplifiers discussed above are provided. No-load output is 2300 volts dc. Outputs under loads of 1 ampere, 500 mA, and 100 mA are respectively 1850, 2000, and 2200 volts dc. Screen voltage is 300 volts dc regulated to 40 mA sink current. Cutoff and operating bias voltages are respectively –120 volts dc and –56 to –90 volts. (Operating bias is regulated at –56 volts dc.) Also provided are 7.6 volts ac, which is adjustable from 5.5 volts to 6 volts ac at 6 amperes. One-hundred-and-twenty volts ac are provided at the blower receptacle.

The transformer (custom manufactured by H. E. Johnson and Associates, Clearwater, Florida for ARCOS) has input provisions for either 120 Vac or 240 Vac, 50 to 400 Hz. Assembled around a 1540-watt hipersil core, the transformer is vacuum impregnated with insulating varnish then coated with a single-part thermostating epoxy for mechanical protection. In ambient air of 25 degrees C with convection-radiation cooling only, continuous operation at 1000 watts dc results in a temperature rise of less than 30 degrees C.

The primary power circuit consists of a three-wire power cord, double-pole power switch, power relay (optional) and top and bottom cover interlock switches (optional). A test switch (also optional) simulates the ground for operating the power relay, which comes from the associated amplifier over pin 7 of the low-voltage connector. The power cord must be sensed correctly for this feature to work with the 120-Vac connection.

A blower receptacle is provided. The power cord from the power supply blower, which is mounted on top of the power supply, plugs into the blower receptacle. The blower cord has a bridged receptacle to which the amplifier blower can be connected. Alternatively, 120 Vac may be connected to pins 2 and 4 of the power supply’s low-voltage connector for operating the amplifier from a receptacle on the amplifier chassis. If this option is chosen, metering resistors for plate current and plate voltage must be located in the amplifier instead of in the power supply.

A review of the schematic will show that a voltage-doubling circuit is used with over 30 μF of electrolytic capacitors in each leg of the rectifier circuit. Six 1000-volt PIV/2.5-amp diodes are used in series in each rectifier leg. The diodes are protected by a secondary fuse (2 amps/1000 Vdc) and a 10-ohm, 50-watt series resistor. The short-circuit current is limited in the high voltage lead by a 25-ohm, 50-watt resistor.

The voltage doubling circuit provides a 1000-Vdc output for developing regulated 300-Vdc screen...
voltage (pin 1 of the low-voltage power connector) by using a series resistor and zener diodes. The zeners are protected from high-voltage transients on the screen leads from the amplifier by series diodes in the screen lead. The screen terminating resistor may be located either in the power supply or in the amplifier.

The screen series resistor and screen terminating resistor values were chosen to maintain regulation with a 40-mA sink current in the screen terminating resistor. Adequate sink current for the screen circuit is essential to the proper operation of tetrode amplifiers. The sink current path through the terminating resistor provides a bleed path for the reverse screen current, which is normal for tetrodes. As I mentioned previously, if an adequate bleed path is not provided, the screen voltage will attempt to rise to the plate voltage and the tube will go into a runaway condition unless protected by surge voltage protectors at the screen-socket terminals. The 40 mA sink current provided for the screen terminating resistor in this power supply meets the tube manufacturer’s recommendation for at least 15 mA per tube.

The series diodes in the screen lead block any high-voltage transients, which can occur in the time interval it takes for the surge voltage protection to operate, from destroying the zeners used for regulation.

The blower, mounted on top of the power supply over the vent slots near the screen-series resistors, provides cooling for approximately 100 watts of heat dissipation from the screen circuit components.

The grid-bias voltage (pin 3 of the low voltage connector) is regulated by a zener and is adjustable by a potentiometer. A delay tube is used to delay operating bias on the amplifier during warm-up. The bias changes from cutoff (−120 Vdc) to operate (−56 to −90 Vdc) when ground is placed on the control jack. Provision is made for metering plate voltage and plate current over pins 2 and 4 of the low-voltage connector (when these pins are not used, as described above, for 120 Vac to the amplifier chassis). Pin 5 of the low-voltage connector provides adjustable filament voltage to the amplifier. Pin 6 provides ground and pin 8 connects to the negative lead of the power supply. The high voltage is connected to the amplifier by RG-59/U coaxial cable using Amphenol MHV connectors.

The conductors between the amplifier and the power supply, other than the high voltage, are connected over an eight-conductor cable (Belden 8448) having one pair of No. 16 AWG conductors used for the filament voltage and ground. The other six conductors are No. 22 AWG.

Construction of the power supply is shown in the photos. The amplifier was assembled on a pre-punched foundation chassis available from ARCOS. Diagrams of the terminal strips used to mount the high-voltage diodes, screen and bias circuitry, and so forth, are shown in fig. 13. Other construction details are included in fig. 14. There’s nothing critical about parts location, so any convenient chassis arrangement may be used. Be sure to provide adequate ventilation for the screen resistors and zeners.

For use with the triode amplifier, the screen regulation and grid bias components are omitted (fig. 15). Note that the metering resistors for the triode amplifier are located in the amplifier.

**test and operation**

Connect the amplifier to the power supply and make the usual checks of blower operation, filament voltage, bias voltage, screen voltage and plate voltage. Set the bias for an idling current of about 100 mA for initial tests. Apply a watt or so of excitation and adjust the grid controls for maximum plate

---

**fig. 14.** Construction details for the zener mounting, capacitor board, and resistor board in the power supply. The transformer is custom made for ARCOS by H.E. Johnson and Associates, Clearwater, Florida. Secondary ac voltages shown at a dc load of 2000 volts at 0.5 amp.
current. Then resonate the plate circuit by observing SWR toward the driving source. Next, adjust the plate load and tune controls for a compromise between maximum output and minimum plate current at an output level of 500 to 600 watts. See table 1 for a typical set of readings taken during the test of one of these amplifiers.

**fig. 15. Schematic of the power supply for the triode stripline kilowatt amplifier.**

table 1. Typical test results for the 220-MHz amplifier.

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<th>E_{grid}</th>
<th>E_{scr}</th>
<th>E_{plate}</th>
<th>I_{grid}</th>
<th>I_{scr 1}</th>
<th>I_{scr 2}</th>
<th>I_{plate}</th>
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<th>drive</th>
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<td></td>
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<td>.540</td>
<td></td>
<td>12</td>
<td>600</td>
<td>1134</td>
</tr>
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</table>

**tubes:** 4CX250R  
**dummy load:** 150 ft (46 meters) of RG-8/U terminated by a Heath Cantenna™  
**operation:** Class AB1  

**Notes:**  
1. Temperature 1 and temperature 2 refer to the temperatures at the two exhaust ports.  
2. Efficiencies are 46 percent on line 2 and 53 percent on line 3.  
3. RFO readings are the relative power output on the multimeter of the amplifier.  
4. The negative grid and screen currents are normal for this type of operation.  
5. Observe the excellent screen voltage regulation.  
6. Plate voltage regulation is 11 percent.  
7. At 1-kW input the power output would be about 540 watts on CW for this operation condition.  
8. SSB inputs can exceed the above figures somewhat if desired.  
9. Key down on CW at 1 kW input, the tubes are within their dissipation rating (500 watts).  
10. Power readings were taken with Bird 43 wattmeters in both drive and output.
A relative indication of the effectiveness of cooling, as well as the relative dissipation shared by the two tubes, can be obtained, as suggested by K2R1W2 by mounting a candy thermometer over the air outlets. Temperatures read in this manner should not exceed 200 degrees F (93.5 C) under any condition. The thermometer is equipped with a pair of stiff wire legs, which facilitate setting it on top of the amplifier over the exhaust holes.

A failure of the air supply, if undetected, will result in a very rapid and disastrous temperature rise inside the plate compartment. Usually the solder on the plate line melts and the finger stock springs out to touch the chassis — grounding the high voltage and operating the high-voltage fuse. In cases observed so far, the tubes have survived. To prevent damage, an air switch in the blower or a pressure switch for the plate compartment may be used to shut down the amplifier to cutoff when air pressure fails.

A suggested setup for switching the antenna when using a transceiver for drive power is shown in fig. 16. Some transceivers don’t furnish ground on transmit, providing either 12 Vdc or some other voltage. There’s not enough current available in some cases to operate a relay. A transistor and relay may be con-
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IC4AT

ICOM
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IC-25A, IC-251A, IC-2KL, IC-451A

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D-1010,
B-1016

FM-2025

FT-208R
FT-708R

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“Kitty Says

“Kitty Says

“Kitty Says

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More Details? CHECK — OFF Page 92
I was young once and knew everything. One of my beliefs was in the honest-to-gosh reading of an SWR meter. You merely put it in the line to the antenna and this magical instrument would tell you just what was happening inside the coaxial line. It was all very simple. Fortunately, I learned rapidly, and in the process amassed six or seven SWR meters. It was always amusing to make SWR measurements on an antenna with one meter and then to repeat the measurements with another meter. It was almost possible to hand-pick the SWR curve I wanted by the correct choice of instrument, since they provided widely different readings.

I'm not the only Amateur who has been led down the daisy path by the SWR meter. The complications associated with this interesting device are more important today than ever before because of the advent of the solid-state, high-frequency transceiver.

the more you have the less you get

Most high-frequency transmitters that have a solid-state power amplifier incorporate a protective circuit that will gradually turn off the amplifier as the SWR on the antenna system rises. Since most antennas are single-frequency devices (that is, adjusted at one spot in a particular band), a low value of SWR is obtainable at only one frequency. Operating the antenna off frequency causes the SWR on the feed system to rise, even though the antenna may work in excellent fashion across the whole band (fig. 1).

Tube-type amplifiers with their adjustable output controls (TUNE and LOAD) can adapt themselves to wide variations in the SWR of the antenna system. The solid-state wide-band amplifier, on the other hand, requires protection against SWR excesses. Hence the fail-safe shut-down design. When the SWR starts to rise, the amplifier transistors are electrical-ly derated for protection.

All well and good, I say, but the user of such equipment must pay attention to the SWR across his band of operation, or he will find that he can get full power output from his rig only over a small portion of the band. This vexing problem is particularly true on 160, 80, and 10 meters, where the width of the band is large in proportion to the center frequency.

enter the SWR meter

To determine just what is going on with regard to a particular antenna, an SWR meter is commonly used to gain a picture of SWR vs frequency, as shown in fig. 1. But is this a true picture of what is happening? Possibly not. A knowledge of the workings of the SWR meter and its use are of utmost importance.

Most modern SWR meters are composed of two directional couplers built into one case. A single indicating meter is switched between the couplers and the meter is usually calibrated directly in terms of SWR (standing wave ratio).

A directional coupler is a device which samples power flowing in one direction in a transmission line but is insensitive to power flowing in the reverse direction.

If the antenna exactly matches the characteristic impedance of the transmission line and also matches the line...
with respect to balance, it will absorb all power transmitted down the line. If a mismatch exists at the antenna, a certain portion of rf power will be reflected back down the line toward the transmitter.

The circuit in the directional coupler picks up energy from the line by means of both inductive and capacitive coupling. The inductive current in the line flows according to the direction of the traveling wave producing it. Thus there can be direct and reflected waves passing through the coupler. The capacitive pickup, however, is independent of the direction of the traveling waves, and the sum of coupled currents in the device produced from the waves of one direction will add in phase. Those produced from waves of the opposite direction will subtract in phase.

The electrical balance of a coupler is such that the current induced from the reverse-traveling wave will cancel the other completely, or nearly so, resulting in a directivity factor in the coupler. This means the coupler is highly insensitive (nullled) to a wave traveling in the reverse direction. Thus the device is sensitive to only one of the traveling waves which produces standing waves by interference. To determine SWR it is necessary to read forward and reverse (incident) power flowing in the line. Two couplers, reverse connected, can do the job (fig. 2). In order to obtain accurate readings, both couplers must be identical. And each coupler should be insensitive to power passing through it in the unwanted direction.

The important characteristic of a coupler is the ratio of the measurement in the forward direction to that in the reverse direction. If the coupler is sensitive to the unwanted reading, the accuracy of the coupler is seriously affected. When two couplers are used to make up an SWR meter, the problem is compounded.

A good laboratory-type coupler will have a directivity of better 25 dB, indicating that the coupler provides 25 dB of discrimination between opposite directions of power flow. An SWR meter made up of two such couplers provides an indicated value of SWR differing from the true value, as shown in fig. 3. As an example, a true value of SWR of 1.5-to-1 on a transmission line can provide an indicated value on the SWR meter which can vary between the extremes of 1.23-to-1 and 1.8-to-1. And most cheap CB-type SWR instruments are not this accurate.

Added to the directivity limitation, most inexpensive SWR meters have a built-in error because of the nonlinearity of the diode used to provide voltage for the indicating meter. At low voltage levels where the SWR reading is of the greatest importance, diode linearity is poorest.

Finally, all directional couplers are sensitive to second harmonic voltage that may exist in the antenna circuit. Since the antenna is mismatched at harmonics, it is possible for high SWR to exist at a harmonic frequency and if the coupler is accidentally placed at a point in the line having high harmonic current, pickup of this current will adversely affect the reverse reading of the coupler.

You may scoff at this notion, and say that the second harmonic level of your transmitter or exciter is "down 35 dB," or some such number. Well and good, but just remember that with a high value of antenna mismatch reflection at a harmonic, the harmonic voltage passing through the coupler may be many times higher than you suppose. And don't forget that when a coupler is measuring the reflected wave in a line, it may be measuring as high as 40 to 50 dB below the fundamental signal level. That is to say, the unwanted harmonic voltage can easily be of the same order of magnitude as the measured reflected wave.

For best results, therefore, I suggest you buy the best SWR meter you
PERFECT DIRECTIVITY AND CORRECT INDICATED SWR

fig. 3. Extreme limits of indicated SWR versus actual SWR for a coupler having 25 dB of directivity. (Adapted from “Possible Errors in VSWR Measurement,” Breetz, QST, November, 1959.)

fig. 4. “Antenna currents” can be induced into shield of transmission line by inductive coupling (M) between line and antenna and also by lack of balun at the antenna.

 table 1. Recommended non-resonant transmission line lengths (L) for the high-frequency Amateur bands. Lengths indicated include distance between one tip of driven element and feed point, plus coaxial line length.

<table>
<thead>
<tr>
<th>Band</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>20 - 30 MHz</td>
<td>23 - 30 ft.</td>
</tr>
<tr>
<td>35 - 40 MHz</td>
<td>35 - 44 ft.</td>
</tr>
<tr>
<td>40 - 45 MHz</td>
<td>46 - 47 ft.</td>
</tr>
<tr>
<td>47 - 50 MHz</td>
<td>52 - 63 ft.</td>
</tr>
<tr>
<td>52 - 55 MHz</td>
<td>71 - 81 ft.</td>
</tr>
<tr>
<td>55 - 60 MHz</td>
<td>86 - 90 ft.</td>
</tr>
<tr>
<td>60 - 65 MHz</td>
<td>93 - 97 ft.</td>
</tr>
<tr>
<td>65 - 70 MHz</td>
<td>106 - 112 ft.</td>
</tr>
<tr>
<td>70 - 75 MHz</td>
<td>110 - 137 ft.</td>
</tr>
<tr>
<td>75 - 80 MHz</td>
<td>114 - 141 ft.</td>
</tr>
</tbody>
</table>

(Adapted from The ARRL Antenna Book)

can afford. Some SWR meters are made up of two directional couplers, back to back. Others have a single coupler with a reversible element. I prefer the latter type. One coupler made in the U. S. A. has plug-in heads for various frequencies and power levels. It is useful for both hf and VHF antenna measurements. While I don’t believe in “plugging” name brands in this column, be assured this high-flying instrument is really a Bird!

pitfalls in making SWR measurements

So now you have a good SWR meter! Congratulations. If you use it properly, you’ll get meaningful information. But you just can’t jam it into a coaxial line and expect the instrument to do its job. It is up to you to make sure that the meter reads what you are looking for (true SWR of the antenna) and not a jumble of information resulting from unwanted coupling between the transmission line and the antenna. Let me explain.

Any conductor in the field of an antenna is coupled to it inductively. The degree of coupling depends upon the position of the conductor with respect to the antenna and the distance between antenna and conductor. A good example of such a conductor is a parasitic element in a beam antenna. It is closely coupled to the antenna and tuned closely to its frequency.

Other conductors coupled to your transmitting antenna are overhead power lines, telephone lines, and your transmission line.

Yes! The outer shield of a coaxial line can be inductively coupled to the antenna if it runs parallel, or nearly so, to the antenna and is elevated above ground level. An example of this is shown in fig. 4, a typical antenna installation.

Antenna current induced on the outer shield of a transmission line will influence the SWR reading of currents within the line, as the outer shield is no longer at ground potential, even though the SWR meter and transmitter are supposed to be near ground potential. The outside of the coax line has become part of the antenna system (due to inductive coupling to the antenna) and thus becomes part of the load on the line, in addition to the antenna load. The SWR on the line is now determined by the composite load of the antenna and the outside of the line.

This is one reason why changing the length of the transmission line changes the SWR reading. The portion of the load caused by unwanted line coupling is being changed!

how to reduce unwanted line currents

To achieve an accurate SWR reading on your transmission line it is necessary to detune and decouple the line from the antenna. Certain lengths of transmission line, as measured between one tip of the antenna and the SWR meter (L) are not resonant in the Amateur bands (table 1).

Cutting the transmission line to rec-
decoupling the transmission line

Let's assume that your situation is this: You have a tri-band beam for 20-15-10 meters atop a 40 foot tower. The coax feedline runs down the tower to the 10 foot level and then runs along your roof for about 20 feet to the station, then drops down to near ground level, entering a window near the transmitter. You make SWR measurements across each band and get a reassuring set of curves that bear a little resemblance to the "typical" curves supplied by the manufacturer. How can you determine your curves are valid?

The easiest and quickest check is to add four or five feet of coaxial line between the SWR meter and the antenna and rerun the SWR curves. If the shape or amplitude of the SWR curves change, it is a good bet that you have unwanted coupling between the antenna and transmission line. Of course, to make this experiment, it is understood that the unbalanced, coaxial line is properly terminated at the antenna in a balun or other device which provides a match between the unbalanced line and the balanced driven element of the beam. (Note: Such a test is valid only if you have a good SWR meter.)

If you find that interaction between line and antenna exists and it is impractical to move or otherwise change position of either the line or the antenna, what is to be done?

One helpful and easy thing to do is to coil the line into a simple rf choke at the foot of the tower. Four or five turns about a foot in diameter, taped together with electrical tape, will help to "cool off" the line at the tower.

At the station end of the line, a second, similar coil may help solve the problem. The coil can be made by splicing an extra length of line into the system with coaxial adapters. After the coils are in place new SWR runs are made, with and without the extra spliced-in line section. Now, do

![fig. 5.](Image)

fig. 5. (A) Measured SWR curve with unwanted antenna currents flowing in outer conductor of coaxial line. Curve indicates antenna system resonance near 14.15 MHz with very low SWR. The more accurate curve run with antenna currents on line suppressed shows a flatter response, with higher value of minimum SWR at 14.225 MHz. Classic curve of measurement A may lull Amateur into thinking he has a perfect match, whereas the correct curve (B) shows match is not quite so good.

ommended lengths helps, but is not a total cure to the problem.

Of equal importance is the fact that the transmission line should not run parallel to the antenna elements. And it should be positioned close to the ground and not suspended above the ground. This is a large order when a rotary beam is used because at some beam heading the antenna elements will probably run parallel with the transmission line. The best solution is to run the coaxial line along the ground from the antenna tower to the station, or bury it beneath the ground in a section of water hose. The worst thing is to run the coaxial line a long distance above the ground from tower to station (along the roof top, for example). This places the coax line up in the air and closer to the active antenna.

But what do you do when it is impossible to cut line length to a recommended value and the line must run along the rooftop in the vicinity of the antenna? Obviously, induced antenna currents are going to flow in the outer conductor of the line. How can SWR measurements made under these conditions be trusted?

![fig. 6.](Image)

fig. 6. Bird coupler (top) is machined from solid brass casting. The coupling element is plugged into hole at center of the structure. Center conductor of coaxial line can be seen running through the center of the coupler. Voltage pickup is taken from fitting on side of the casting. Inexpensive "CB-type" SWR meter (bottom) has two directional couplers made up of wires placed parallel to center conductor of open coaxial line section. Open trough with one side exposed permits antenna currents flowing on outside of the coaxial line to enter the measuring section of the line where the coupling elements are located.

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2716 EPROM programmer

Easy method for burning EPROMs with your microprocessor

Programming read-only memories (ROMs) is a virtual necessity if you’re developing microprocessor-controlled ham gear. Many microprocessor systems, such as the KIM, SYM, and AIM, are capable of developing and debugging simple control programs, but none have ROM programming capability. As a first step to integrating microprocessor control into my projects, I found that I needed something to go with my KIM that would program and read erasable programmable read-only memories (EPROMs). The circuits presented here will work with the KIM, SYM, and AIM and are adaptable to other microprocessor systems.

What is an EPROM? Why program one?

Solid-state memory for microprocessors comes in two forms, random-access memory (RAM) and read-only memory (ROM). The former is a general-purpose memory that can be used to store both program and data material. Information stored in it may be changed at will.

ROM is used for permanent programs and their constants. As the name implies, it can be read only. If scratch pad space beyond that contained in the microprocessor is required, it must be supplemented with RAM.

The shortcoming of RAM is that it is volatile; that is, if the power is removed the RAM forgets. Many schemes for making RAM more permanent are available, such as battery or large-capacitor power supply backup, but in the long run permanent programs should be stored in some sort of nonvolatile ROM. This places them beyond the reach of power failure and programming errors, which would leave a RAM filled with garbage.

An additional advantage of ROM is that microprocessor chips, when powered up, awaken seeking the address of their first instruction at a particular location. If the program stored in the ROM has the correct addresses, pointing the way to the beginning of the program, the routine will be self-initiating when powered up. In other words, the operator doesn’t even have to know that a microprocessor is there.

ROMs fall into different categories depending on their application and the volume of their manufacture. The particular chip considered here, a 2716, is selected based on its popularity and price. The 2716 is an erasable programmable read-only memory (EPROM) with 2K (2048 bytes) of memory. It may be programmed with simple circuitry and straightforward techniques.

To correct mistakes, or when you wish to replace an old program with a new one, the chip can be erased by exposure to ultraviolet light.* Programming

*Commercial UV erasers are available from computer hobbyist supply houses for $65 to $75. A simple home-built eraser is described by Golter, "Build a Low-Cost EPROM Eraser." BYTE, April, 1980.

By C.A. Eubanks, N3CA, P.O. Box 127, Valencia, Pennsylvania 16059
fig. 1. Schematic diagram of the Z716 EPROM programmer-reader.
Programming can be accomplished a byte at a time, or the entire EPROM may be programmed at once. Unfortunately, during erasure, everything gets erased together.

**How it works**

The basic programming circuit is built around an 8255 programmable peripheral interface chip (PPI). The 8255, designed for compatibility with the 80XX line of microprocessors, is easy to interface and has sold in sufficient quantity to be economical for Amateur use. It can be addressed by the controlling microprocessor bus and directed to either read from or write to its input/output lines, providing communication with the EPROM being programmed.

The microprocessor board supplies the byte address of the location in the EPROM being programmed, the data to be recorded and the enabling signal. Programming all of the 2048 EPROM locations at one time requires about 103 seconds, or 50 milliseconds per location.

Programming a 2716 EPROM requires a 25-volt positive signal. This voltage is derived from a 24-volt LM340T monolithic regulator chip, as shown in fig. 1. A diode in the regulator's reference leg increases the output to approximately 25 volts. The chip is powered by three 9-volt transistor radio batteries connected in series. Though not elegant, the technique is quite effective for small volume use.

The microprocessor chips used in the KIM, AIM and SYM circuits have a pull-down-to-reset signal, which is not directly compatible with the 8255 chip (pull-up to reset). Rather than add an additional IC to provide the extra gate needed for hardware reset, I decided to do it with software. This is accomplished by converting all of its ports to inputs with a control instruction before connecting the batteries to the 24-volt regulator.

The programming algorithm includes a testing operation before programming to ensure that the EPROM area to be used is clean (all bits set to logical ones — the erased condition). The algorithm also contains a second comparison routine after programming to ensure that data transcription is correct. Fig. 2 is a flow chart showing the initial verification and programming. The final comparison uses similar technique.

I selected the addressing of the PPI and 2716 EPROM read socket to be compatible with the SYM, AIM, and KIM microprocessor boards. On the first two these addresses are mapped into the user expansion space. On the KIM they fall above the monitor.

Both the PPI and the EPROM read socket must be enabled; that is, told when to respond to address and data bus inputs. This is accomplished by applying a logical zero (low level) signal to the NOT chip enable (CE) line of the selected chip. A 74LS156 three-to-eight line open collector decoder performs this selection by address decoding. It breaks the 64K addressing capability of the microprocessor into eight 8K segments as shown in fig. 3. The NOT 8K segment 3 line (8K3) is assigned to the PPI and the 8K2 line is assigned to the EPROM read socket.

Beyond this point the KIM differs from the SYM and the AIM. The latter two fully decode the portions of memory that they use internally, whereas the KIM does not have this feature. Without outside help it can’t tell one 8K segment from another. The KIM’s NOT decode enable (DE) must be brought low only when its on-board devices are to respond to an address on the address bus. To permit this action the 8K0 and the 8K7 lines from the 74LS156 decoder are
wire-OR tied (paralleled — see fig. 3) together. The first of these picks up on-board KIM devices during normal addressing, and the latter allows interrupt responses.

This occurrence gives rise to an interesting opportunity for the KIM user. If the 8K7 line is instead wire-OR tied to the EPROM read socket, the resident EPROM will respond to interrupts. To permit this, a DIP switch is included on the board to control which device gets the 8K7 enabling signal. As mentioned above, this applies only to the KIM.

Note that if circuitry is set up for use with the KIM, it may be used on AIM and SYM systems with the following provisions:

1. The EPROM read socket must not be selected as the interrupt source. If it is, bus contention will result when both the host board and the EPROM try to respond to 8K7 addresses on read cycles.

2. Line 20 on the expansion connector must be left open with the AIM.

The AIM 65 uses this connection for other purposes. Alternatively, AIM and SYM users may omit the connections to the 8K0 and 8K7 lines, the associated resistors and the DIP switch.

You will notice that this is a fairly wasteful method of addressing. The PPI uses only four addresses out of the 2K available to it. The EPROM read socket uses only 2K of its 8K. I found no use for the additional possible PPI addresses, but the EPROM addresses are a different story. The price of 2732 EPROMs (4K byte EPROMs) are today about half of what 2716s cost two years ago. I expect that before long I'll modify the programming board to handle 2732s and possibly 2764s. They are virtually identical to the 2716 and require minimal circuit changes.

architecture

The unit is assembled on a Radio Shack two-volt edge-card board, part no. 276-154. I selected this board because it has an edge connector matching that found on the KIM, SYM, and AIM systems. Those with a different type of microprocessor system may choose alternative boards better suited to this application.

All ICs except the voltage regulator are socketed. The sockets make modifications easier and let you remove static-sensitive MOS chips when the board must be handled. For the EPROMs, zero insertion force (ZIF) sockets were selected to minimize wear and tear on the 2716 chips. The ZIF sockets are wide enough so that they overhang the outermost connections for each pin on the programmer board, which means that the wiring must be completed before installing the ZIF sockets. Advance planning in the wiring layout is necessary to keep it from obstructing ZIF socket installation.

A four-unit DIP switch was selected to control decoding for top-end-of-memory addressing. A conventional toggle switch could be substituted, but the DIP switch fits more neatly into the board layout. These switches are needed only if KIM compatibility is sought.

All wiring and components are located on the blank (non-foil) side of the board. Install short jumpers for \( V_{cc} \) and ground connections before wiring the longer runs.

Jumper cables with suitable connectors are available to connect the boards together but they are considerably longer than my layout required and they're expensive. I obtained wire-wrap connectors at a hamfest and hand wired them together instead. I mounted the programmer board quite close to the microprocessor board to minimize reactance and crosstalk. The wiring between the connectors is 1-1/2 inches (4 cm) long. After wire wrapping the connectors I applied a liberal coat of silicone rubber sealer to keep repeated installations and removals from loosening the wraps.

The KIM expansion connector contains all the necessary lines except the \( DE \) line. Fortunately, there are unused lines on the expansion connector, termi-

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*Available from the Computerist, Inc., Post Office Box 3, South Chelmsford, Massachusetts 01824. Program documentation and programmed EPROMs are available from the author. Send an SASE for information.
nals 18, 19 and 20. I wired the DE line from the application connector’s terminal K to terminal 20 on the expansion connector with an on-board jumper. No board changes are needed on the SYM or AIM.

The regulator chip probably doesn’t need a heat-sink at the currents drawn during programming. If one is handy in the junk box, however, there’s certainly nothing lost in installing it.

The batteries are connected using 9-volt battery connector leads. When not programming EPROMs the batteries should be disconnected and may be relegated to the refrigerator to prolong their life.

Both the jumper lead on the microprocessor board and the battery connector leads on the programmer board should be anchored with spots of silicone rubber sealant, which prevents placing mechanical stress on their solder joints.

notes on components

The basic cost of the parts for this board is about $34, which includes the two ZIF sockets but does not include connectors or 2716 EPROMs. Comparable EPROM programming systems available commercially start around $80. Many run in excess of $200 wired, although the higher end boards usually have multiple EPROM read sockets and may perform other functions such as handling memory expansion as well.

The more exotic chips used here (the 2716s and the 8255) were picked up from mail-order suppliers. Most of these accept phone orders for credit-card billing and feature 48-hour order turn-around via UPS. A sample of five suppliers of 8255s from the July, 1981, issue of BYTE magazine shows a unit price range from $5.40 to $9.95.

initial operation

Check the wiring to make sure its correct and that no shorts or solder bridges exist before applying power. My original wiring was all right, but apparently I’m losing my color vision: one resistor was the wrong value.

With both EPROM sockets empty, connect the board to the microprocessor system’s expansion connector (with the system de-energized, please!). Fire up the microprocessor and direct the 8255 PPI to output logical ones to each of the programming socket’s pins, one at a time. Check operation with a voltmeter. Note that KIM and similar systems can’t write to the PPI manually because the monitor tries to treat the PPI as a read/write memory. Things get snarled up with the monitor when it tries to read back what it’s just written. To write to the PPI, a store absolute instruction (or its equivalent) must be used.

I’ve encountered only one hardware problem since startup (my software is always full of bugs). During an EPROM read, one or more bit at one or more addresses would occasionally be wrong. To the best I’ve been able to determine, this was due to low power-supply voltage (about 4.75 volts on the EPROM board). I corrected this by installing a 0.001 µF bypass at the read socket’s Vcc pin and by cranking the power supply up to 4.98 volts under load.

Table 1 lists the rules for using the PPI to talk to the programming socket as they apply to this circuit.

<table>
<thead>
<tr>
<th>function</th>
<th>PGM/CE port line PA3 (2716 Pin 18)</th>
<th>OE port line PA4 (2716 Pin 20)</th>
<th>2716 response</th>
</tr>
</thead>
<tbody>
<tr>
<td>program</td>
<td></td>
<td></td>
<td>program data in</td>
</tr>
<tr>
<td>program inhibit</td>
<td>0</td>
<td>1</td>
<td>high Z outputs</td>
</tr>
<tr>
<td>program verify</td>
<td>0</td>
<td>0</td>
<td>data out</td>
</tr>
</tbody>
</table>

8255 PPI control for this circuit

control addresses:
|$6000 = port A$
|$6001 = port B$
|$6002 = port C$
|$6003 = control word register$

control words:
|$80 = make all ports outputs (for programming)$
|$9B = make all ports inputs (resets PPI and applies no signals to EPROM)$
|$89 = make ports A and B outputs; port C an input (for verify)$

port assignments:
| port A = EPROM address lines A10 through A8 correspond to port lines PA2 through PA0, EPROM CE to PA3, EPROM OE to PA4 |
| port B = least-significant address byte |
| port C = data bus |

Once everything checks out, you’re ready to go. Remember in use that the 8255 PPI and the 2716 are static-sensitive devices and may be ruined by improper handling. Follow manufacturer’s recommendations and use an insulated insertion/extraction tool to handle the devices.

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performance capability of active mixers

Part two: practical circuits and testing

In the first part of this article, I introduced the basic characteristics that distinguish active double-balanced mixers. The equations that define the significant interfering effects during signal handling were also developed. Now we will look at some practical mixer circuits and examine some testing techniques.

practical circuits

Because we are interested in the problems of balancing harmonics and other unwanted products, we will not consider simple, single-stage mixers but instead will look at several active mixer circuits that pretty much represent the state of the art. Depending on the bias to cutoff ratio of the devices selected, fairly high intercept points (up to +30 dBm) can be obtained.

Fig. 9 is the schematic of an active mixer using two VHF transistors in push-pull. Because of the number of spurious frequencies generated, substantial filtering at the output is recommended, as can be seen from the highpass-lowpass filter section. The two rf chokes at the base of each transistor prevent unwanted oscillation at fairly high frequencies.

Fig. 10 shows a balanced mixer using the 3N200 field-effect transistor in push-pull. This is a multiplicative mixer in which the rf input signal is fed in parallel.

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Ed Oxner of Siliconix has designed several mixers based on the U257 transistor. The schematic is shown in fig. 11. Siliconix also made a version with four field-effect transistors in one package. Fig. 12 shows the schematic of such a configuration using the U350.

VMOS transistors have become popular, and a push-pull version with the VMP4 power FET, as described by Doug DeMaw, was shown in QST.* (See fig. 13.)

Most loads into which the mixers operate do not present precisely 50 ohms to the mixer output. Also, most designers of active mixers have carefully avoided indicating the effects that occur in active mixers as the termination changes from a purely resistive 50-ohm load into some other value. In general, any change in resistive load that does not introduce reactive components does not affect the mixer substantially. A VSWR of 1:2 from 50 ohms, or change of load from 25 to 100 ohms resistive, does not have too many adverse effects. If, however, the mixer is terminated by an LC filter or crystal filter, the impedance changes and becomes reactive, and the intercept point changes drastically. In some cases with active mixers, I have observed instabilities even to the point where the mixer turned into a low frequency oscillator.

Three basic circuits are known to prevent change of impedance. Fig. 14 shows a recommended ar-
fig. 10. Push-pull mixer with 3N200 FETs.

fig. 11. Double-balanced mixer with Siliconix U257.
arrangement whereby the mixer, in this case a passive double-balanced mixer, is terminated by the input impedance of a grounded-gate field-effect transistor. It must be remembered that grounded-gate field-effect transistors properly biased exhibit purely resistive input over an extremely wide frequency range.

This holds true in most cases basically from dc to several hundred MHz. The CP643 or CP640 made by Teledyne Crystalonics is a good choice.

Another alternative is a feedback amplifier that uses noiseless feedback as described in the literature\(^5\) based on patent 3891934 of 1975. The third alternative is the use of a diplexer whereby the image at the output of the mixer is terminated with a 50-ohm resistor.

Probably the best solution is a combination of two transistors with a diplexer as shown in fig. 15. Again, for convenience, the circuit is shown with a passive balanced mixer together with this particular termination circuit.

Let's now take a look at some systems calculations that will yield a surprising result.

**Active mixer with perfect termination.** Consider an active mixer such as the Plessey SL6440 in any of the previously shown schematics. The noise figure under large-signal operation is around 11 dB for the Plessey device and 8 dB for a U257 mixer. Relative to the typical loss of 6 dB in a passive mixer, the zero-dB gain of an active mixer already represents gain; to be specific, 6-dB gain over the passive device. Let us assume further that the amplifier following uses noiseless feedback and its noise figure is 2 dB. As the mixer has unity gain, the noise figure at the input is equal to the noise figure of the second stage plus the noise figure of the mixer, and in the case of the Plessey mixer the resulting noise is 11.2 dB.
If we use a U257 stage, we get a 10-dB noise figure if we allow the same amount of gain. The intercept point is determined by the mixer and the second amplifier, and because of the special rf feedback applied in the second amplifier, we will, for the moment, assume that the second amplifier does not contribute any intermodulation distortion products. The very moment we operate the mixer with gain, we must take distortion of the second stage into consideration.

**Passive double-balanced mixer with termination stage.** Let’s use the same example with a high-performance double-balanced mixer. The double-balanced mixer has an insertion loss of 6 dB, and the noise figure also is 6 dB.

The noise figure of the termination stage again is 2 dB, which results in a total systems noise figure of 8 dB, or 2 dB better than the previous example with the U257. Because of the 6-dB losses of the double-balanced mixer, the intermodulation distortion of the double-balanced mixer can be neglected, and the designer can concentrate on the mixer itself. This, I am sure, is a surprising result for most design engineers.

It is important to understand that the termination stage, when using the noiseless-feedback system, must also operate into a stable load. Any changes of the output load of such an rf feedback amplifier will be reflected into the input. A recommended way to reduce this change is to operate this stage at a higher than necessary gain; 3 or 4 dB is sufficient. A resistive pad with 3-dB attenuation will then prevent dramatic changes at the output.

In the case of the grounded-gate field-effect transistor as a termination, this circuit works reliably only if the drain-to-source feedback capacitor is kept extremely small and the capacitance is basically determined by the transistor itself.

If the output stage must operate into a crystal filter, we will find that most crystal filters outside the passband characteristic exhibit high impedance, which is either inductive or capacitive. This effect can be reduced by using a highpass filter at the output that incorporates the crystal filter. If the crystal filter’s impedance increases, the highpass filter is mistuned, and the voltage at the drain or collector remains low. As a result, the third-order intermodulation distortion products remain low. In a conventional circuit, it is sometimes found that the sudden increase of impedance at the output of the transistor makes the intermodulation distortion deteriorate.

**Passive mixer with active devices**

Recent developmental work in the field of mixers
indicates that the best way of achieving high intercept points in mixers is the use of:

1. Bipolar transistors as switches and with feedback.
2. Field-effect transistors as switches.

In ordinary applications, mixers using active devices operate on the nonlinear transfer characteristic, as explained earlier. Diode mixers are substantially better, because here the device is only switched on and off, and if the on-off resistance of the device has a high enough ratio, the device will be fast enough to follow the local-oscillator drive waveform. Enough local-oscillator drive power is available, theoretically, so that no harmonic distortion should occur. The losses would be 3 dB, as would the noise figure; thus we would not observe any intermodulation distortion products at all. Using active devices, we depend upon the nonlinearities of the input and output ports and, as with the field-effect transistor, possible distortion of the gate source diode and the potential nonlinearities of the channel resistance.

The state of the art for mixers using field-effect transistors without operating voltage and, therefore, only as switches, indicates approximately a +42 dBm input intercept point, 5.5 dB noise figure and insertion loss, and local-oscillator requirements of about +23 dBm. The local-oscillator drive requirement really results from the fact that a certain voltage has to be available at the gate electrode. In a quad configuration, this voltage can be as high as 50-volts PEP into the input capacitance of the transistor. The step-up transformer helps reduce the required power.

Fig. 16 shows the schematic of such a recommended mixer which, for test purposes, has a tuned input. This circuit is based on a patent issued to Mr. William Squires in 1968, number 3383601. It can be reported that for 1-volt input signal or +13 dBm, the third-order intermodulation distortion products are at -83 dBm, or 100 dBm down. This would increase to an intercept point of +70 dBm but can only be achieved in a narrowband circuit. In a wideband configuration, only 40 to 42 dBm is obtainable. The isolation between oscillator and signal port is about 60 dB and provides about 40-dB isolation to the i-f signal.

The area of using passive mixers with active devices is fairly new. The only company that seems to have a commercial product is Lorch in New Jersey, and the latest prices I have seen for their mixer were $600 or $700.

I had built an active mixer based on feedback and switching, which was published in Ham Radio. This mixer with similar performance was used in the Rhode & Schwarz 400-watt transceiver in the Tornado warplane.

testing

To make proper tests on the mixers using signal generators, a hybrid coupler with at least 40 dB isolation between the two input ports and an attenuator are required. The test setup provided by DeMaw in QST and shown in fig. 17 is ideal for this. He used
two signal generators with outputs around 14 MHz and combined them. An attenuator drives the mixer under test (MUT), the local-oscillator signal is supplied by a VFO, and the output is then analyzed.

The 2N5109 amplifier shown may not be sufficient for extremely high intercept points, as this stage may no longer be transparent. For stability tests when using active mixers, it is recommended one have a reactive network at the output of the mixer for the sole purpose of checking mixer instability.

Rather than use expensive signal generators, two oscillators with extremely low harmonic content and

![fig. 16. Schematic of a passive double-balanced mixer using FETs in a quad arrangement. This circuit represents the state-of-the-art that is possible today. While the narrowband version can have input intercept points of +70 dBm, a wideband version achieves about +42 dBm.](image)

fig. 17. Recommended test setup for measuring mixer intermodulation distortion. (Original circuit appeared in QST, January, 1981, page 20.)

![fig. 18. Low-noise crystal oscillator with 60-dB harmonic suppression.](image)

fig. 18. Low-noise crystal oscillator with 60-dB harmonic suppression.

![fig. 19. Ultra-low-noise crystal oscillator with ultimate noise floor of 168 dB/Hz.](image)

fig. 19. Ultra-low-noise crystal oscillator with ultimate noise floor of 168 dB/Hz.
very low noise sideband performance can be used. A convenient circuit to provide the required harmonic suppression and low noise is shown in fig. 18, based on an earlier paper of mine. For those interested in obtaining an additional 20-dB improvement in noise sidebands and need a test oscillator with this performance, the circuit in fig. 19 is recommended. This oscillator shows an ultimate noise floor of 168 dB/Hz at 1 kHz off the carrier. As can be seen, this oscillator is a derivative of the earlier one. The input impedance of the grounded-base stage is about 2 ohms and, therefore, does not really deteriorate the Q of the crystal.

**summary**

I have explained that ordinary active mixers based on the inherent nonlinearities of their transfer characteristic by definition will show a lower intercept point than is possible with passive devices. Passive devices are already used with great success, and the termination-insensitive mixers, although they are not yet offered below 1 MHz, are currently the state of the art in diode mixers. By using feedback techniques together with switching-type active stages and bipolar transistors (or better yet, using modern power junction field-effect transistors), I have measured third-order and higher input intercept points to 40 dBm. In selective cases and narrowband frequency operation, +70 dBm intercepts have been reported. It is not likely that these figures will be useful, as the termination stages of following crystal filters or other devices will become the limiting factor.

I have just learned that Mini-Circuit Laboratories has introduced a new mixer, type VAY1, that claims a 38-dB intercept point at the input, which results in a 32-dBm intercept point at the output. However, these drive requirements are much higher than for the passive FET mixer.

**references**


**bibliography**

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words-per-minute readout for the deluxe memory keyer

After having constructed several "Accu-Keyers,"¹ and the "Deluxe Memory Ke~er,"² I found that the one thing missing in both was the ability to tell exactly how fast I was sending. I thought there must be a simple way to add this convenience to these superb keyers. Most of the hardware to accomplish this is already mounted in the keyers: the clock pulses, counters, and the drivers with the readouts. The only missing part for the frequency counter are the timing gates, a time base and the readout latches.

words per minute versus frequency

The handbook states that 24 WPM equals 10 dots per second times 2.4. To put it in another way, 24 WPM equals 24 dots per 2.4 seconds. It takes two clock pulses to equal one dot (1 dit and 1 space), so 24 WPM equals 48 clock bits per 2.4 seconds, or 24 clock bits in 1.2 seconds. All we have to do now is to let the counters tally the clock bits for 1.2 seconds, put in two 7475 latches to hold this count and display it on the readouts, and we have our WPM counter. On my first attempt at building a speed readout, I used a 3.58-MHz TV crystal and a 5369 oscillator chip for the 60-Hz time base and a multipole rotary switch to change modes. But then I decided a less expensive approach would be to use the 60 Hz from the secondary of the power transformer and two more 7400 chips to do the switching instead of the more expensive rotary switch. Thus evolved the circuit described here.

theory of operation

Refer to fig.1. In the WPM mode, the DPDT switch

View of the speed adapter board mounted on top of the driver board. The existing keyer memory board is under the driver board. The bit count, WPM mode switch is mounted in the upper left corner, rear, next to the speaker. (Photo courtesy of Bill Mansfield, W1MLT.)

By Vernon W. Smith, WA1OEH, Box 75, E. Thompson Rd., Thompson, Connecticut 06277
except as indicated, decimal values of capacitance are in micro-
farads \( (\mu F) \); others are in picofarads \( (pF) \); resistances are in ohms.

\[ k = 1,000 \quad M = 1,000,000 \]

fig. 1. Schematic diagram of the WPM readout adapter. The circles indicate wire connections.

turns on the 5-volt supply for IC U1, U2, U3, and U6. The switch also grounds pin 2 of U4C and U5A, blocking out the count and reset pulses from the memory chip. Sixty Hz is fed into U6 from the power-
transformer secondary. The 60-Hz waveform is shaped and conditioned by U6 and is fed into U3 (divide by 12). The output of U3, pin 8 (5 Hz) is fed into a second divide-by-12 IC, U1. The output of pin 8 of U1 is 2400 milliseconds, or 2.4 seconds long. This pulse is fed to pin 5 of the U2D NAND gate, which is high for 1.2 seconds and low for 1.2 sec-
onds. During the high on U2D pin 5, the gate is held open, which allows the clock input pulse on pin 4 to pass through for a 1.2-second period. This count is in words per minute; that is WPM equals the number of clock pulses per 1.2 seconds. When pin 8, U1 goes low after the count time is concluded, pin 1 of U2A is pulsed, causing a high strobe pulse at its output and at pins 4 and 13 of U7 and U8. This action moment-
arily opens the 7475s, which latch onto the count and display it through the readouts. A pulse at pin 9, U2C resets the counters to zero and a new count begins.

In the NORMAL mode of operation, pins 9 of U4B and U5B are grounded, blocking the WPM counter and reset pulses, which allows the normal memory bit count and reset pulses to pass through to the counters. The other half of the DPDT switch removes the 5 volts from U1, U2, U3, and U6. This disables them and places 5 volts on the 7475 latches, holding them open to allow a steady bit count to be dis-
played.

construction

The speed adapter board uses eight inexpensive ICs, about $3 worth. The whole project shouldn't cost over $10 if all the parts have to be purchased.
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fig. 2. Foil side, A, and component side, B, of the speed adapter board. Dotted lines are wire jumpers.

fig. 3. Foil side of Deluxe Memory Keyer driver board. Cut foil at 12 points marked with X. Connect wires from speed adapter board to identical letters on driver board. Add two jumper wires as indicated by dotted lines.
Components are mounted on the non-foil side of the board, as shown in fig. 2. All resistors can be ¼ watt. Capacitors are disc, except the 50 µF, which can be either electrolytic or tantalum at 10 volts. Dotted lines indicate jumper wires. Be sure the ICs are oriented correctly. I used sockets for the ICs, which makes them easier to change in case I zap one. Care must be taken when soldering in the sockets to prevent solder bridges across the pins. (I can supply a limited number of drilled and etched circuit boards for $6 each.)

The hookup requires 30 wires, so ribbon cable is the best route to follow. The jumper wire along the edge of the board in fig. 2 is used to anchor the ribbon cable to the board. Mount the cable here first and leave about 5 inches (13 cm) for a hookup, then peel back each wire to the jumper anchor before routing to the designated letter. Measure and route each wire separately to its terminal. Cut, insert and solder each wire by number as shown in table 1 to avoid a mixup.

driver board
To prepare the driver board, the foil must be cut in twelve places as shown in fig. 3. Also add two jumper wires as shown by the dotted lines, from pin 11 of U7D to pin 14 of U6D, and from pin 11 of U6D to pin 14 of U5D.

wiring the driver board
The ribbon cable can now be installed on the driver board. Leave about 1½ inches (4 cm) of the cable between the boards for a hinge, and anchor the ribbon on the foil side of the driver board with a jumper wire across it. Leave about 7 inches (18 cm) of the ribbon on the free end. Peel back the first twenty-four wires to the anchor point. The last six wires may stay together. Run the last six wires to the DPDT switch and solder them on as indicated in table 1. All the connections to the driver board will be made on the foil side. Tack each of the first sixteen wires to the IC pin numbers as indicated in table 1. The next six wires go to the indicated letters. The twenty-third wire goes to either side of the secondary of the power transformer. The twenty-fourth wire is soldered to point L, in the keyer section of the driver board. The speed adapter board is mounted directly over the driver board with spacers in between.

operation
With the switch in the WPM position, a string of dots about 3 seconds long will give an exact WPM count on the readout. Set the speed control to the speed at which you would like to send. The readout will receive a count update every 2.4 seconds. Any hesitations during sending will show a corresponding slower speed on the readout. With the switch in the normal position the keyer will operate as it did before the adapter board was installed. The bit count from or to the memories will be displayed on the readout.

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More Details? CHECK — OFF Page 92
The sixth part in a continuing series designed to help you upgrade your ticket

This series of articles is being presented to help you pass a higher grade Amateur license exam, to give you the basic radio theory needed to pass a Novice, Technician/General, or Advanced class license test. After these basics are presented in as simple a form as possible, there will be articles covering Extra class license subjects.

This month we will examine the three basic forms in which active devices can be connected in amplifier circuits. Also, feedback and decibels will be outlined briefly.

classes of amplifiers

When active devices are used in amplifying circuits, the portion of the time that current flows in them determines the class of operation. The class of operation also helps to determine the efficiency of the amplifying stage. The three basic classes of operation are class A, class B, and class C. Most amplifiers are operated in class A, so we will discuss this class first.

Class A operation is the least efficient, but also distorts the amplified signal the least. Let's consider the solid-line curve of fig. 1 first. Represented vertically, at the left, is the output circuit current (drain, collector, plate). Represented horizontally is negative, zero, and positive bias voltage for N-channel FETs or for VTs (vacuum tubes). You will note that the class A bias value is about halfway between output current cutoff and zero bias (or output current maximum). The shaded area illustrates the output-circuit current variations that occur with both a small input ac signal ($e_i$), and then a higher amplitude input signal. If the large input signal were made much greater it would drive the device into output current cutoff, or into the bent part of the curve at the top, or both.

Under the maximum undistorted output signal condition shown, the efficiency of the amplifier stage (its ac power output compared with its output circuit dc power input) would be about 45 percent (the theoretical maximum is 50 percent). With a small input signal, the dc power input to the output circuit averages the same but the ac power output will be much less because the output dc current varies less. In this case the efficiency of the stage might be only 5 or 10 percent. Normally, the efficiency of an operating class A audio amplifier averages about 25 percent. Class A operation has the advantage of requiring only a single active device to produce relatively undistorted amplification. This class of amplifier can be used for both audio and radio frequency amplification when distortion must be kept to a minimum.

Can you see that if the operating curve of the active device used happened to be curved or nonlinear (which it usually is), as is shown dashed, that even if

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the input signal voltage had equal positive and negative peaks, the output circuit dc variations would be smaller for the negative half cycle of input ac, and larger for the positive half? This would produce a noticeably distorted output audio signal from such a stage. It would produce harmonics of all frequencies it amplified.

Class B operation is more efficient than class A but usually distorts more than class A; it requires two push-pull devices (first one operates and then the other) in a single stage for audio amplification. In the two-stage amplifier shown in fig. 2a the first stage is a JFET class A single-ended af amplifier (one device used and one end of its load grounded or bypassed to ground) which is transformer-coupled to a push-pull class A or B power amplifier second stage, transformer-coupled to a loudspeaker. Whether the output stage is biased to class A or B is determined by the values of bias developed by $R_1$ and $R_2$. The curve shown in fig. 2b indicates that with no bias at all on the BJT ($R_I = \text{zero ohms}$) there is no collector current with no input signal. The bases of the two NPN transistors must be forward biased a little to produce class B operation ($R_I = \pm 5 \text{ ohms}$). The BJT bases must be forward biased considerably more to put them in class A (dashed line). With class B bias, only the positive half of the input signal produces $I_C$ in the transistor whose curve is shown.

Assume this curve is for the top BJT. Then during the next half cycle of the input signal, the top BJT will have no $I_C$ flowing, but the lower BJT is now being driven by the second half of the input ac signal. The voltages at the top and bottom of the center-tapped input transformer secondary will be $180^\circ$ out of phase with each other. As the top BJT is being fed the signal as a negative (reverse) voltage, the lower BJT is being fed the same signal but as a positive (forward) voltage, and the lower BJT now has $I_C$ flowing in it. In this way the output transformer primary has current flowing first in its top half and then in its bottom half. Both primary currents induce voltages into the secondary to feed power to the loudspeaker. Can you see that each BJT is working only about half of the time? For example, during much of the second half cycle of the input signal, the top BJT rests (cools). As a result, a pair of class B biased active devices will produce much more power output than two parallel-connected (B to B, E to E, and C to C) class A devices, both of which would have current flowing in them at all times. Practical class B stages can operate with maximum efficiencies in the range of 55 percent. If a single BJT in class A is capable of 1 watt output, two connected in parallel would be capable of 2 watts output, but two in push-pull could be driven to produce about 8 watts output if in class B, partly because of their ability to rest almost half of the time, plus the fact that a greater input signal could be accommodated with class B bias. When producing power output all active devices must have effective heatsinks attached to their collectors, drains, or plates.
It should be mentioned that you will normally find BJT curves plotting emitter-collector voltage versus collector current with a series of more or less parallel horizontal lines indicating various base (input circuit) current values.

Class AB operation means biasing to some point between class A and class B. For audio amplifiers this also requires that two push-pull devices be used. As you might expect, class AB is in between classes A and B as far as power output and distortion are concerned. When using push-pull class A and class AB there is one other advantage. The two devices are operating $180^\circ$ out of phase with each other so that any bend in the operating curves, which normally produces distortion in the output of a single ended stage, will be cancelled by the reversed curve (effectively) of the other device. This results in nearly distortion-free amplification in such push-pull stages. When using vacuum tubes (and N-channel FETs) class AB implies that the input signal should never be great enough to drive up into the positive bias area and produce grid (gate) current. Class AB2 operation implies some grid current and that the input circuit resistance (impedance) is engineered low enough so that when the signal does drive into an input current condition, there will be no unwanted voltage drops developed across any resistance in the input circuits to produce distortion.

Class C operation is not used with audio amplification, but usually only in transmitter rf amplifiers or in oscillators. Class C stages are normally biased from 1.5 to perhaps four times the cutoff, or pinchoff, bias value. (Technically, a BJT with zero bias is in class C.) As a result, for a major proportion of the $360^\circ$ of any input cycle (perhaps $240^\circ$) the device has no output circuit current; it’s resting. It can produce maximum output power at perhaps 85 percent efficiency. This means that the output circuit current comes in relatively narrow ($120^\circ$), high-amplitude pulses. If these powerful pulses are forced through an output LC circuit tuned to some radio frequency, such as 7 MHz (as in fig. 3), the pulses shock-excite the LC circuit into flywheel oscillation and produce nearly sinusoidal 7-MHz rf ac output from it. Of course, the input rf ac to such a class C amplifier should also be at 7 MHz, although if it is at 3.5 MHz the circuit will operate as an rf frequency doubler. Since class C amplifiers depend on the flywheel effect of relatively high-Q LC circuits to produce the output ac, the stages may use either single-ended one-device circuits, or may use push-pull circuits. Push-pull rf ac circuits produce less even-order (second, fourth, sixth, and so forth) harmonic ac output, and may be desirable to suppress such frequencies. If the coupling to a load on a single-ended class C rf amplifier ($R_L$ in fig. 3) is too tight it will lower the Q of the LC circuit involved and drain the flywheel energy on the undriven half cycle too much. This will reduce the amplitude of the second flywheel half cycle and distort the sinewave shape of the ac wave. This develops harmonics and unwanted radiations. Push-pull stages do not have this difficulty as much as single-ended stages.

**common-emitter type amplifiers**

There are three basic methods of connecting and feeding three-element amplification devices such as FETs, BJTs, and VTs. Let’s examine the basic theory as it applies to bipolar junction transistors. The most common amplifier circuit configuration is that of the common-emitter (common-source for FETs, common-cathode for VTs), also known as the grounded-emitter (-source, -cathode). The circuit shown in fig. 4 illustrates a single-ended NPN BJT class A amplifier. $R_1$ and $R_2$ provide the forward bias for the transistor. $R_3$ produces the stabilizing bias to keep the transistor from thermal runaway in power stages. $R_L$ is selected to produce a voltage-drop across itself with no signal input, of perhaps one third to a half of the $V_{CC}$ value (depending on the $R_3$ value). $R_L$ will probably be in the 1 to 5 kilohm range, depending on the beta of the transistor, the bias value, and the $V_{CC}$ value. For af amplification the input and output coupling capaci-
tances would be in the 5 to 15 \( \mu F \) range. The lower the lowest frequency to be amplified, the larger the capacitance values required. For high-frequency rf amplification the capacitors might be in the 0.01 to 0.1 \( \mu F \) range, but such an untuned rf amplifier is not often used. In rf amplifiers the \( R_L \) would usually be supplanted by a tuned LC circuit, shown dashed.

Generally, regardless of the \( V_{CC} \) value, the load resistor and bias values are selected so that with no signal input the voltage-drop across \( R_L \) will be equal to the EC voltage-drop for good class A operation.

Common-emitter-type circuits represent a relatively low impedance load for the stages ahead of them, perhaps 50 to 200 ohms, and look like a medium impedance to the stages that follow. With FETs, fig. 5, the common-source circuit usually uses a resistor between source and ground to provide bias. The input impedance of this stage in class A operation is essentially as high as the resistance of the gate-to-ground resistor used in it. The output circuit impedance is perhaps one half of the resistance of the drain load resistor, \( R_L \). If using tuned LC circuits for rf amplification, the output impedance may be several hundred to several thousand ohms. Vacuum tubes are similar to the FETs in impedance and component values. The phase shift of a signal amplified by all of these common-emitter type circuits is usually considered to be 180°.

common-base type amplifiers

A common-base, or grounded-base (-gate, -grid) amplifier is shown in fig. 6. \( R_1 \) and \( R_2 \) form a voltage divider across the power supply \((+V_{CC} \) to ground\). Since the base is connected up the voltage divider from ground it is more positive than the emitter, which returns to ground through the input resistor. The amount of forward bias required to produce class A operation is that which will produce a static (no signal) \( I_C \) value somewhat less than half the maximum rated \( I_C \) value. \( C_1 \) is a filter capacitor whose function is to prevent any signal voltage variations from developing across \( R_1 \) as the input signal forces the \( I_B \) to change at the signal frequency and amplitude. Input-signal currents through the 200-ohm input resistor develop varying voltage-drops across it. Since these varying voltages are between base and emitter they force the collector current to vary, producing an amplified voltage-drop variation across the relatively high resistance load resistor. Can you see that the emitter current is the sum of both base and collector currents (arrows)? Therefore, the ratio of \( I_C \) to \( I_E \) will always have to be something less than 1, usually in the range of 0.95 to 0.98. This ratio or value is known as the alpha \((\alpha) \) of the transistor. The alpha cutoff frequency is the frequency where the gain of the transistor drops to 0.707 of its gain at 1 kHz.

Although a common-base circuit cannot have a current gain, it can have a voltage gain and therefore a power gain. If the signal voltage applied to the 200-ohm resistor is going positive, the base, being held constant by \( C_1 \), can be considered to be going relatively more negative. Such a reverse bias to the base reduces the \( I_C \) and the voltage drop across \( R_L \) decreases. This allows the top of \( R_L \) to become more nearly the \(+V_{CC} \) value, or more positive. Therefore, as the input goes more positive so does the output, representing a phase shift of zero. With 0° phase shift there is little likelihood that this circuit will become regenerative and produce oscillations on its own. This is one of the advantages of such circuits. We say they are inherently stable, meaning that they are not likely to break into unwanted oscillations. When using vacuum tubes and an LC circuit in place of \( R_L \), the equivalent circuit is known as a grounded-grid amplifier and is used in many of the present-day high-power Amateur rf linear amplifiers.

common-collector-type amplifiers

The third basic form of amplifier circuitry is the
common-collector, or grounded-collector (-drain, -plate) circuit, fig. 7. At first glance it may look like a common-emitter circuit, until you note that there is no load resistor in the collector circuit. Instead, the load resistor, $R_L$, is also the emitter resistor, $R_E$. Note that the collector is held at ac-ground potential by the relatively large value bypass capacitor, $C_{bp}$. An input signal applied base-to-ground causes $I_C$ and $I_E$ variations, which in turn produce variations of the voltage drop across $R_E$. The output voltage variation across $R_E$ will always be a little less than the input voltage being fed to the base. However, this circuit has both current gain and power gain. Because a positive going signal to the gate produces an increase in $I_C$ and therefore a more positive emitter, the common-base circuit produces zero phase shift of the amplified signal current. With $0^\circ$ phase shift the circuit produces no regeneration and cannot break into oscillation, even at high frequencies. It is used mostly as an impedance-changing circuit. The input impedance is usually about half the value of $R_L$. With a high value of $R_L$ the grounded-collector circuit acts as an impedance reducer, somewhat like a step-down transformer, except that a transformer cannot produce any power gain. This circuit is also known as an emitter-follower (source-follower, cathode-follower).

![fig. 7. Common-collector or grounded-collector amplifier, also known as an emitter-follower circuit.](image)

feedback

The term feedback deserves a little more explaining, since there is some of it in most circuits in which amplification occurs. The term itself indicates that a signal voltage (or current) being fed into an amplifier stage is amplified, and then part of the amplified output voltage (or current) is fed back into the input circuit. Assuming no time difference (phase shift) between the input signal voltage and the output, if the ac feedback voltage is out of phase $180^\circ$, or degenerative, the net result is a lessened effective input signal voltage, and therefore a lessened total gain of the stage. If there had been any distortion of the signal voltage (any change in the signal’s waveshape) produced by the amplifying process of this particular stage (due to nonlinearity of its operating curve, for example) this distortion is also fed back $180^\circ$ out of phase. This can almost completely cancel out any such distortion produced by the stage. Of course, if the input signal voltage were distorted to begin with, the degenerative feedback could not correct that. Degenerative, or inverse, feedback is used in many linear af amplifier systems, and may be developed as a degenerative loop feeding back distortion-cancelling voltages from a third stage back to a first stage, for instance.

The actual gain of a stage with inverse feedback can be determined by the formula

$$A_f = \frac{A}{1 + (-\beta)A}$$

where $A_f$ is the amplification or voltage gain with feedback, $A$ is the amplification without feedback, and $-\beta$ is a decimal fraction of output voltage (or current) fed back to the input. As an example, an amplifier without feedback has a gain of $20$. What is its gain if the feedback factor is $5$ percent ($5$ percent $= 0.05$)? Try working out this problem. Your answer should be an undistorted gain of only $10$. With such inverse feedback, signal gain may be reduced, but when undistorted output is paramount this loss can be made up by merely adding another stage with inverse feedback. Furthermore, without feedback the gain of an amplifier stage might drop to a 50 percent output voltage at some lower frequency (due to the $R$ and $C$ values used in it). With the 5 percent feedback, at the original half-voltage frequency the output will now drop only about $10$ percent. Thus, inverse feedback widens the bandwidth (spread of frequencies amplified almost equally) of an amplifier. In addition, instead of developing a $45^\circ$ phase shift at the original half-voltage frequency, the resultant phase shift will be reduced to about $27^\circ$. So inverse feedback also reduces any undesirable phase shift (another form of distortion) in an amplifier.

If the feedback is close to $0^\circ$ it has the tendency to make the stage around which it is introduced increase its gain (particularly at one frequency) if the feedback percentage is small. If such a forward or regenerative feedback is increased, it will usually cause the gain to rise so much at the frequency at which the greatest gain is being produced that the stage will break into oscillation at this frequency. As you may have deduced, both positive (regenerative) and negative (degenerative or inverse) feedback can be either an advantage or a disadvantage in electronic circuits.

decibels

You have probably noted that all of the amplifier
circuits have power gains, although the common-collector has no voltage gain, and the common-base has no current gain. We often talk about amplifier gain in terms of bels (after Alexander Graham Bell), or more commonly in tenths of bels (deci = 1/10), called decibels, or dB. Basically, a 3 dB gain means an approximately two-times output-over-input power gain. Another 3 dB, or a total of 6 dB gain represents an output of approximately twice plus twice more, or a four times gain. A gain of 10 dB is a power gain of exactly ten times. Thus, if an amplifier has 10 dB gain it amplifies the power fed to its input ten times. (We are speaking of power amplifiers here, not the usual class A FET or VT voltage amplifier which has essentially no power input but may have considerable power output.) If a power amplifier has 13 dB gain, it amplifies ten times (10 dB) plus 3 dB more, or twice ten, for a gain of twenty times the input power. What would be the power gain of an amplifier that has 16 dB gain? Can you see that the output would be ten times, plus being doubled again, then doubled again (10 dB + 3 dB + 3 dB), for a total gain of forty times the input power?

Since power is proportional to either the $E^2$ or the $I^2$ (from $P = E^2/R$, and $P = I^2R$), the voltage gain of an amplifier (assuming input and output impedances to be equal) in dB will be the square root of the power gain. A power gain of 6 dB is the equivalent of a four times power gain. The square root of 4 is 2. Therefore, 6 dB is the equivalent of a voltage (or current) gain of two times. If receivers have signal-strength-indicating meters (S-meters) and each S-unit is calibrated to be 6 dB greater than the one lower, one S-unit is equivalent to doubling (or halving) the signal voltage applied to the input of the receiver.

For accurate computations of decibels the formula to use is:

$$dB = 10 \log \frac{P_1}{P_2}$$

where $P_1$ is the larger power value being considered.

Variations of this formula to determine dB gain or loss using voltage or current values, again assuming input and output impedances to be equal, are:

$$dB = 20 \log \frac{E_1}{E_2}$$

$$dB = 20 \log \frac{I_1}{I_2}$$

If you use a table of logarithms you can solve dB problems using these formulas. For example, what is the dB gain if the input is 10 watts and the output is 20 watts? We have already given you the information to obtain an approximate answer of 3 dB. But let’s try this problem using the first dB formula:

$$dB = 10 \log \frac{P_1}{P_2}$$

$$dB = 10 \log \frac{20}{10}$$

$$dB = 10 \log 2$$

From a log table, pocket calculator, or slide rule, if we can determine that the log of the number 2 is 0.301. So, ten times 0.301 means 3.01 dB is exactly twice the power. If the power input were 20 watts and the output 10 watts, the ratio is still equal to 2:1, or 3.01 dB, but now it is a loss of 3.01 dB, which may be expressed as $-3.01 \text{ dB}$. You may not be given dB problems to work in an Amateur test, but you might! In any case you might want to memorize these ballpark dB values:

1 dB = 1.25 times the power
(a 25 percent power increase)

3 dB = 2 times the power
(a 100 percent increase)

6 dB = 4 times the power

9 dB = 8 times the power

10 dB = 10 times the power (exact)

20 dB = 100 times the power (exact)

30 dB = 1000 times the power (exact)

If you have a beam antenna with a 20 dB gain and you feed it 100 watts of rf ac, how much more power does it put out in its direction of maximum radiation? ______________. Wow! Where can I get an antenna like that?

**FCC test topics**

The following Technician/General test topic is discussed in this article, but should be understood by Advanced applicants also:

- **decibels**

The following Advanced class test topics are discussed in this article:

- amplifiers, classes A, AB, B, C, characteristics
- common-emitter class A transistor amplifiers, bias network, signal gain, input and output impedances
- common-collector class A transistor amplifier, bias network, signal gain, input and output impedances.

For more information on these subjects it is recommended that you refer to a textbook such as *Electronic Communication*, by Robert L. Shrader, McGraw-Hill Book Company, available through Ham Radio’s Bookstore, and to radio handbooks.
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**SPECIAL EVENTS**

**SUN 1 APRIL:**
- **WIAW Qualifying Run:** Sun 1 Apr. WIAW Qualifying Run. Contact W4HIM - 10.

**MON 2 APRIL:**

**TUES 3 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**WED 4 APRIL:**

**THUR 5 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**FRI 6 APRIL:**
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**SAT 7 APRIL:**
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**MON 9 APRIL:**
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**TUES 10 APRIL:**
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**WED 11 APRIL:**
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**THUR 12 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**FRI 13 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**SAT 14 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**SUN 15 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**MON 16 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**TUES 17 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**WED 18 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**THUR 19 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**FRI 20 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**SAT 21 APRIL:**
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**SUN 22 APRIL:**
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**MON 23 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**TUES 24 APRIL:**
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**WED 25 APRIL:**
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**THUR 26 APRIL:**
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**FRI 27 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**SAT 28 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**SUN 29 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**MON 30 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**TUES 31 APRIL:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**WED 1 MAY:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**THUR 2 MAY:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**FRI 3 MAY:**
- **WIAW Qualifying Run:** WIAW Qualifying Run. Contact W4HIM - 10.

**SAT 4 MAY:**
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<table>
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<tr>
<th>Model</th>
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<td>Model CB converter board</td>
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Cushcraft

<table>
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<td>Ham IV</td>
<td>$169</td>
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<tr>
<td>Tailtwister</td>
<td>$228</td>
</tr>
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</table>

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Tell ’em you saw it in HAM RADIO!!
binary coded decimal addition

In the course of typical Amateur Radio do-it-yourself projects, you may seldom encounter a requirement for a binary adder. But if you should, the CD4008 is a logical choice if you prefer CMOS. However, the 4008 operates in pure binary format and is not intended to be used for BCD arithmetic.

I recently encountered the need to add one binary coded decimal number to another. The requirement resulted from the desire to drive a seven-segment LED display from a BCD thumbwheel and also a divide-by-

n IC in a frequency synthesizer. Sure enough, in true Murphy’s Law fashion, these two requirements were not compatible — that is, unless a fixed BCD number was first added to the thumbwheel number for the divide-by-

n circuit.

A search through all available handbooks failed to identify an integrated circuit dedicated to adding in BCD. Rather than resorting to use of more exotic means to resolve this dilemma, such as by the use of a PROM for example, a combination of five readily available and inexpensive integrated circuits was assembled. What emerged is shown schematically in fig. 1. Despite the fact that the circuit shown is a hybrid combination of CMOS and TTL (they were available in the “goodie” box), from the first time power was applied, performance was flawless.

The operation of the BCD “adder” is as follows:

Assume that initially the \( Q \) outputs of the 4029 counter agree bit-for-bit with the thumbwheel output bits at the comparator input terminals. This sets the \( 74LS85 \) \( A = B \) output high, which in turn prevents clock pulses from reaching the 4029 counter. This is a static situation that will persist

fig. 1. Circuit diagram of the BCD adder.
until such time as the thumbwheel switch setting is changed.

When the thumbwheel setting is changed, the falling edge of \( A = B \) activates the 74121 monostable, which in turn generates a single LOAD or PRESET-ENABLE pulse. This pulse presets the 4029 counter so that its \( Q \) outputs agree with the JAM inputs. In other words, instead of being reset to zero, the counter is preset to some binary number equal to or greater than zero, according to the settings of the JAM switches.

At the same time, the comparator \( A = B \) output is LOW, and the NAND gate is open, allowing clock pulses to reach the 4029 input terminal. Thereafter, the 4029 \( Q \) outputs change, starting from their preset values, until they again agree with the thumbwheel bits. At this time, the comparator’s \( A = B \) output goes high, thus interrupting the train of clock pulses and freezing the contents of the counter.

The number of clock pulses necessary to re-establish this static \( A = B \) state depends upon the difference between the thumbwheel number and the JAM number. If, for example, the number nine (1001) is dialed, and the JAM number selected is three (0011), the number of clock pulses gated to the 4029 counter will be \( 1001 - 0011 = 0110 \). (9 - 3 = 6).

The gated pulses are also fed to an auxiliary counter, a 74LS192, which has its DATA inputs hard-wired to zero. Each time it receives a LOAD pulse from the monostable, it will reset to zero and accumulate a count numerically equal to the number of the clock pulses in the train. In the example above, the 74LS192 \( Q \) outputs will read 0110, or six, after being reset to zero, which is exactly the desired result.

The 74LS192 provides a difference output equal to the thumbwheel number minus the JAM number. Would it not therefore be more appropriate to call this circuit a SUBTRACTOR? After all, it’s the difference that counts.

Norman J. Foot, WA9HUV
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last-minute forecast

April brings a continuation of equinoctial propagation conditions and so could be a ball — remember last year’s March and April? One cannot expect a duplication of last year, but be prepared to take advantage of the situation because there is good probability that something will happen. The sun/earth alignment makes for maximum ionospheric effect from solar wind particles released from the sun. The periods in April that look good for geomagnetic disturbances and erratic ionospheric movements are around the 10th, 19th, and 29th. These last two are associated with periods of solar flares and a 27-day solar activity maximum.

For the first week and a half of the forecast period the lower bands will probably be the best for DX. The higher bands are expected to improve the last weeks of the month both for trans-equatorial paths and north/south paths. North/south paths are enhanced the day or two before a disturbance; then the trans-equatorial paths are enhanced during the disturbance. The former is hard to recognize except by the presence of very low A and K geomagnetic indexes and by the use of beacons (see last month’s DX Forecaster).

Other geophysical phenomena which may be of interest to DXers this month are the moon and meteor showers. The perigee of the moon’s orbit (for moon-bounce DX) is on the 25th at 2100 hours; the moon will be at full phase on the 8th at 1018 hours. There will be short meteor shower, the Lyrid, on April 20-22. The rate is five per hour, hardly a real help for meteor-scatter DX. But a bigger shower, the Aquarid, starts before the end of the month, peaks on May 5, and ends by mid May. Its rate is 10 to 30 per hour.

Let’s take a look at last year’s spring equinox DX season. Using 6-meter contacts reported to HR Report as an indicator of good DX conditions, let’s check on the openings and accompanying solar and geomagnetic conditions. With the solar flux above 240 and a mild geomagnetic disturbance in progress on February 28, a dip in the flux increased the disturbance so that a southern Africa to northern Europe trans-equatorial path opened up. At the same time, Liberia to the southwestern U.S.A. in a low-latitude east/west path was also good. On March 1st and 2nd other trans-equatorial paths from southern Africa to Europe and South America to the U.S. were open. After a quiet 3rd and 4th, a large disturbance with aurora arrived on the 5th and 6th with Columbia/Caribbean stations working into the U.S. On March 7th, the South America to Japan path was a big trans-equatorial opening.

The next disturbance, on March 13-15, found Ascension Island making over one hundred Japanese contacts; the flux was still above 200. Argentina and Chile had trans-equatorial openings all three days. There were good DX openings between Australia/New Zealand and the U.S. on the 14th and 16th. By the 21st and 22nd, the mid-latitude ionosphere was enhanced just before the commencement of the March 24-26 disturbance at low solar flux; east/west paths from the U.S. to Liberia, from Southern Africa to Hawaii, and from Canada/the Caribbean/French Guiana to Japan were open. The disturbance was very effective on the 26th between Southern Africa and Europe and between Australia/New Zealand and the Caribbean as the flux came back up above 200 again. LU3EX worked 550 JAs during the first half of March.

On the 6th of April the southern Africa/Hawaii path came alive again as the solar flux peaked on up to 276. From April 11 to 14 there was such a big storm most DX was wiped out, but a lesser disturbance on April 19-21 had French Guiana working Australia and southern Africa at the same time. Mexico worked Australia and New Zealand. The solar flux was about 200 during this disturbance. You can see how our solar-ionosphere connection really gives our DX hobby its interesting openings. Some geophysical rules of thumb, WWV flux and geomagnetic values, and beacons to use as tools can enhance the fun. Good DXing this month.

band-by-band forecast

Six meters should provide frequent band openings with a peak during the early afternoon hours on many days.
<table>
<thead>
<tr>
<th>April</th>
<th>00h</th>
<th>03h</th>
<th>06h</th>
<th>09h</th>
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*Look at next higher band for possible openings.*
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The dismantling of some towers should be done with the use of a crane in order to minimize the possibility of member, guy wire, anchor, or base failures. Used towers in many cases are not as inexpensive as you may think if you are injured or killed.

Get professional, experienced help and read your Rohm catalog or other tower manufacturers’ catalogs before erecting or dismantling any tower. A consultation with your local, professional tower erector would be very inexpensive insurance.

---

Trans-equatorial north/south paths will be the best. Your guide to possible openings will be strong openings on 10 meters and high values of solar flux.

Ten and fifteen meters will be loaded with good DX signals from morning until early evening hours almost every day. Times of geomagnetic disturbance will limit the number of signals heard, but listen carefully — they can be from very unusual places. Fifteen meters should be open later in the day than 10 meters. So hit 10 first and finish off with 15. The lengthening of the daylight will be noticed as these bands open up a little sooner and stay open longer in the day.

Twenty meters will be the main daytime DX band, as it is almost always open to some part of the world. It opens to the east as the sun rises and extends into the late evening hours to the west. Geomagnetic disturbances do not affect this band as much as the higher ones, but still look for unusual trans-equatorial DX locations to be coming through once in a while. One-hop trans-equatorial DX of 5,000 to 7,000 miles (8,000 to 11,200 km) may be possible in the late evening hours during some of these unusual conditions.

Forty and eighty meters will have short skip during daylight hours and turn to DX after dark. The bands will open to the east soon after sundown, swing more to the south to Latin America about midnight, and end up to the Pacific areas during the hour or so before dawn. On some nights these bands will be as good as during the winter DX season. The coastal regions usually have the edge for working rare DX on these bands.

One-sixty meters will probably have many nights that will remind you of last summer’s noise. However many good nights are left for working DX before this summer’s noise comes to stay. Many stateside stations are fair game as DX on this band during this season.
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More Details? CHECK-OFF Page 92

State of the Art

K.V.G.

9 MHz CRYSTAL FILTERS

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10.7 MHz CRYSTAL FILTERS

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</table>

Write for Data Sheets.

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Say you saw it in ham radio!
A useful instrument that does a good job on ham-band inductances

easy-to-build inductance meter

This handy instrument will permit you to measure inductance from 0.5 µH to about 10 µH by reading a capacitor dial. I came up with this idea for my own work bench after searching for a good circuit for measuring inductance. This device does the job for the ham bands, and I hope you will like it as much as I do.

description

This inductance meter uses a crystal oscillator on a frequency of 8100 kHz (see fig. 1). The oscillator drives an amplifier. The unknown inductance is placed in the collector lead, and a capacitor across it is tuned for resonance. The dial is calibrated in µH so that the inductance can be read directly.

Another amplifier is used to build up the r-f voltage to operate the 0 to 1.5 mA meter used for tuning to a peak for resonance. A more sensitive meter (such as a 0-50 µA) could be used by shunting a 120-ohm resistor across it, if that is what you happen to have in the junk box.

I suppose the first question that will be asked is, "Why use 8100 kHz for the crystal oscillator?" Well, a surplus 8100-kHz crystal is cheaper to buy than one

By Ed Marriner, W6XM, 528 Colima Street, La Jolla, California 92037
fig. 2. Printed-circuit board and parts placement for the inductance meter.
for the ham bands. Also, if you examine the chart in table 1, you will see that the choice of a 10-400 pF capacitor covers a good range of inductances to be measured. Of course, it's your choice which span you want to cover. If another crystal frequency is used, calibrate the variable capacitor by using the formula in table 1. I was surprised to learn that this formula is not shown in most handbooks.

The accuracy of the dial reading in µH will only be as good as the accuracy with which you can measure the variable capacitor. A better way would be to check or calibrate the dial against known inductance values. Some suggestions on where to obtain inductances of known value are shown under the subheading calibration. (I found that Air-Dux coils did not correspond to the values shown in their listing, and I wouldn't use them for calibrating the meter unless they were measured first.)

construction

There is nothing difficult about building this tester once you've located the parts. Some sources for parts are listed at the end of the article. The printed-circuit layout I used is shown in fig. 2. I cut the board by gripping it between two pieces of angle iron and using a hack saw. Then I filed the edges, sanded the copper with steel wool, and used paper draping tape for the masking. The dots were Avery self-adhesive color-coding labels that I bought at a stationery store. The wider tape I used for the border is Bishop precision-type 201-250-11, 1/4 inch x 20 yards (6.35 mm x 18.2 m). It's expensive, and yellow shelf-paper with stickum on the back can be used, if you prefer, by cutting it on a paper cutter.

The board was etched with ferric chloride, which takes about an hour, but the process can be speeded up by placing a lamp over the solution and juggling it around once in a while. Once the board has been etched and cleaned, by again scrubbing it with steel wool, it should be tinned with a hot soldering iron. Be careful not to apply to much heat, or the copper will come off! (There are other ways of tinning, using a solution.)

The holes can be drilled out using a No. 60 (1-mm) drill. The crystal socket was mounted using a spacer and a long 4-40 (M3) machine screw.

Before putting a unit in a box, test it out by determining if the crystal oscillator is working. Using FT-243 crystals, I found my values of C-1 and C-2 to be correct as shown. Handbooks call for other values, but they did not make the oscillator work for me. You can listen for the oscillator at 8100 kHz with a receiver or put an rf probe at the emitter output. You should have 1.5 volts of rf, enough to drive the following amplifier. The amplifier can also be checked by placing a coil in the collector, resonating it, and measuring with the probe at the collector. A coil with a value of 3-10 µH should be sufficient for this test.

I used germanium diode rectifiers to operate the meter because they provide more dc voltage than the silicon type. The 1N34A, 1N38, or 1N64 work well.

I used a Radio Shack cabinet, 5-1/4 x 3 x 5-5/8 inches (13.3 x 7.6 x 14.3 cm), to house all the parts including the power supply. Before installing the parts mount the capacitor on a piece of plastic to insulate it from the cabinet; use an insulated coupler and plastic shaft through a panel bushing for the dial knob. I used GR terminals with long shanks that went right up to the capacitor terminals. Make the leads as short as possible, because they become part of the inductance.

Once the amplifier has been finished it can be checked by advancing the sensitivity control and resonating a coil placed on the terminals. I put a 100-

---

**table 1. Capacitor calibration setting in µH.**

<table>
<thead>
<tr>
<th>pF</th>
<th>3500 kHz</th>
<th>7000 kHz</th>
<th>8100 kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>200</td>
<td>51.6</td>
<td>38.5</td>
</tr>
<tr>
<td>20</td>
<td>165</td>
<td>25.7</td>
<td>19.2</td>
</tr>
<tr>
<td>30</td>
<td>70</td>
<td>17.2</td>
<td>12.8</td>
</tr>
<tr>
<td>40</td>
<td>52</td>
<td>12.9</td>
<td>9.6</td>
</tr>
<tr>
<td>50</td>
<td>42</td>
<td>10.3</td>
<td>7.7</td>
</tr>
<tr>
<td>60</td>
<td>34</td>
<td>8.5</td>
<td>6.4</td>
</tr>
<tr>
<td>70</td>
<td>30</td>
<td>7.3</td>
<td>5.5</td>
</tr>
<tr>
<td>80</td>
<td>26</td>
<td>6.4</td>
<td>4.8</td>
</tr>
<tr>
<td>90</td>
<td>23</td>
<td>5.7</td>
<td>4.2</td>
</tr>
<tr>
<td>100</td>
<td>21</td>
<td>5.17</td>
<td>3.8</td>
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<tr>
<td>125</td>
<td>17</td>
<td>4.5</td>
<td>3.0</td>
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<tr>
<td>150</td>
<td>14</td>
<td>3.4</td>
<td>2.5</td>
</tr>
<tr>
<td>175</td>
<td>12</td>
<td>2.9</td>
<td>2.2</td>
</tr>
<tr>
<td>200</td>
<td>16.5</td>
<td>2.5</td>
<td>1.92</td>
</tr>
<tr>
<td>225</td>
<td>8.6</td>
<td>2.3</td>
<td>1.6</td>
</tr>
<tr>
<td>250</td>
<td>8.4</td>
<td>2.0</td>
<td>1.5</td>
</tr>
<tr>
<td>300</td>
<td>7.6</td>
<td>1.7</td>
<td>1.2</td>
</tr>
<tr>
<td>325</td>
<td>6.0</td>
<td>1.5</td>
<td>1.1</td>
</tr>
<tr>
<td>350</td>
<td>6.5</td>
<td>1.4</td>
<td>1.1</td>
</tr>
<tr>
<td>375</td>
<td>5.7</td>
<td>1.3</td>
<td>1.0</td>
</tr>
<tr>
<td>400</td>
<td>5.5</td>
<td>1.2</td>
<td>0.9</td>
</tr>
</tbody>
</table>

Note: The 8100-kHz crystal was used because of its range and availability. Calibrate variable capacitor for whichever crystal you have from this formula:

\[
\frac{1}{L} = \frac{4\pi^2 f^2 C}{f^2 + \frac{4\pi^2 f^2 C}{(39.5)(7 \times 10^8)^2 (10 \times 10^{-12})}}
\]

Example: 7000 kHz = \( \frac{1}{39.5 \times 49 \times 10} = 0.0000516 = 51.6 \mu H \)
table 2. Some values of Air-Dux coils which might be used to calibrate or check the dial. Values must be checked first.

<table>
<thead>
<tr>
<th>Air Dux number</th>
<th>total inductance of coil (µH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>404</td>
<td>0.18</td>
</tr>
<tr>
<td>406</td>
<td>0.39</td>
</tr>
<tr>
<td>408</td>
<td>0.71</td>
</tr>
<tr>
<td>410</td>
<td>1.10</td>
</tr>
<tr>
<td>416</td>
<td>2.87</td>
</tr>
<tr>
<td>432</td>
<td>11.30</td>
</tr>
<tr>
<td>504</td>
<td>0.27</td>
</tr>
<tr>
<td>506</td>
<td>0.61</td>
</tr>
<tr>
<td>508</td>
<td>1.10</td>
</tr>
<tr>
<td>510</td>
<td>1.60</td>
</tr>
<tr>
<td>516</td>
<td>4.30</td>
</tr>
<tr>
<td>632</td>
<td>17.30</td>
</tr>
<tr>
<td>604</td>
<td>0.38</td>
</tr>
<tr>
<td>606</td>
<td>0.86</td>
</tr>
<tr>
<td>608</td>
<td>1.52</td>
</tr>
<tr>
<td>610</td>
<td>2.38</td>
</tr>
<tr>
<td>616</td>
<td>6.08</td>
</tr>
<tr>
<td>632</td>
<td>24.20</td>
</tr>
<tr>
<td>804</td>
<td>1.02</td>
</tr>
<tr>
<td>806</td>
<td>2.33</td>
</tr>
<tr>
<td>810</td>
<td>4.67</td>
</tr>
<tr>
<td>816</td>
<td>16.30</td>
</tr>
<tr>
<td>832</td>
<td>66.30</td>
</tr>
</tbody>
</table>

Other values of coils can be obtained by writing Illumitronic Engineering Corporation, 680 East Taylor Avenue, Sunnyvale, California.

The meter dial plate is shown in fig. 3.

**calibration**

Several suggestions on calibrating the dial are given. I used a 0-400 pF capacitor, a BDC type out of Swan equipment, that I found at a flea market. The 0-365 pF units are more common and are available from J.W. Miller and Radio Shack.

One of the best ways of calibrating or checking your dial readings is to use the rf chokes listed below. They seem to be very accurate:

- Miller rfc 420 0.22 µH
- Miller rfc 220 0.82 µH
- Miller rfc 144 1.72 µH
- Miller rfc 50 8.20 µH
- Ohmite Z-50 10.00 µH

Other values of the type 4500 and 70F106AI chokes can be obtained from the J.W. Miller catalog (J.W. Miller, 19070 Royes Avenue, Compton, California 70224). If you know someone who can measure inductance, the Air-Dux line in the 400, 500, 600 and 800 series are also good to use. Some of the approximate values of Air-Dux are given in table 2.

**table 3. Reference chart for other J.W. Miller coils that could be used for calibration.**

<table>
<thead>
<tr>
<th>J.W. Miller stock No.</th>
<th>inductance µH</th>
<th>Q</th>
</tr>
</thead>
<tbody>
<tr>
<td>4580</td>
<td>0.10</td>
<td>70</td>
</tr>
<tr>
<td>4582</td>
<td>0.15</td>
<td>80</td>
</tr>
<tr>
<td>4584</td>
<td>0.22</td>
<td>95</td>
</tr>
<tr>
<td>4586</td>
<td>0.33</td>
<td>100</td>
</tr>
<tr>
<td>4588</td>
<td>0.47</td>
<td>105</td>
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<td>4590</td>
<td>0.68</td>
<td>113</td>
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<tr>
<td>4592</td>
<td>0.75</td>
<td>115</td>
</tr>
<tr>
<td>4594</td>
<td>0.82</td>
<td>112</td>
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<tr>
<td>4602</td>
<td>1.00</td>
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<td>4604</td>
<td>1.50</td>
<td>62</td>
</tr>
<tr>
<td>4606</td>
<td>2.40</td>
<td>63</td>
</tr>
<tr>
<td>4608</td>
<td>3.90</td>
<td>70</td>
</tr>
<tr>
<td>4609</td>
<td>5.50</td>
<td>70</td>
</tr>
<tr>
<td>4610</td>
<td>6.20</td>
<td>67</td>
</tr>
<tr>
<td>4611</td>
<td>8.20</td>
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<td></td>
</tr>
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<td>70F336AI</td>
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<td></td>
</tr>
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<td>70F436AI</td>
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<td></td>
</tr>
<tr>
<td>70F476AI</td>
<td>4.7</td>
<td></td>
</tr>
<tr>
<td>70F686AI</td>
<td>6.8</td>
<td></td>
</tr>
<tr>
<td>70F826AI</td>
<td>8.2</td>
<td></td>
</tr>
<tr>
<td>70F105AI</td>
<td>10.0</td>
<td></td>
</tr>
</tbody>
</table>

See the catalog for a complete list of values in between. Contact J.W. Miller Co., Sales, Mr. Bill Courtney, 19070 Reyes Avenue, Compton, California 90224.
Table 3 gives values for some J.W. Miller coils, and table 4 shows some typical toroid values. All these components are available from many sources throughout the country.

**Table 3**

<table>
<thead>
<tr>
<th>band</th>
<th>core</th>
<th>inductance (µH)</th>
<th>no. turns</th>
<th>enamel wire size (AWG)</th>
</tr>
</thead>
<tbody>
<tr>
<td>40 meters</td>
<td>T-50-2</td>
<td>13</td>
<td>50</td>
<td>28 (0.3)</td>
</tr>
<tr>
<td>20 meters</td>
<td>T-50-2</td>
<td>8</td>
<td>44</td>
<td>28 (0.3)</td>
</tr>
<tr>
<td>15 meters</td>
<td>T-50-2</td>
<td>4</td>
<td>25</td>
<td>28 (0.3)</td>
</tr>
<tr>
<td>10 meters</td>
<td>T-50-2</td>
<td>0.6</td>
<td>12</td>
<td>24 (0.5)</td>
</tr>
</tbody>
</table>

Table 4. Some typical toroid values collected from magazine articles.

**Table 4**

<table>
<thead>
<tr>
<th>core</th>
<th>inductance (µH)</th>
<th>no. turns</th>
<th>enamel wire size (AWG)</th>
</tr>
</thead>
<tbody>
<tr>
<td>T-68-1</td>
<td>0.5</td>
<td>10</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-68-6</td>
<td>0.5</td>
<td>8</td>
<td>20 (0.8)</td>
</tr>
<tr>
<td>T-68-6</td>
<td>0.7</td>
<td>12</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-68-6</td>
<td>0.8</td>
<td>13</td>
<td>26 (0.4)</td>
</tr>
<tr>
<td>T-68-6</td>
<td>1.0</td>
<td>14</td>
<td>20 (0.8)</td>
</tr>
<tr>
<td>T-68-2</td>
<td>1.2</td>
<td>13</td>
<td>20 (0.8)</td>
</tr>
<tr>
<td>T-68-2</td>
<td>1.8</td>
<td>19</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-68-2</td>
<td>2.0</td>
<td>20</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-68-2</td>
<td>2.1</td>
<td>22</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-68-2</td>
<td>6.0</td>
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<td>24 (0.5)</td>
</tr>
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<td>T-68-2</td>
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<td>T-68-2</td>
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<td>T-68-2</td>
<td>24.0</td>
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<td>28 (0.3)</td>
</tr>
<tr>
<td>T-50-2</td>
<td>0.57</td>
<td>7</td>
<td>26 (0.4)</td>
</tr>
<tr>
<td>T-50-6</td>
<td>0.60</td>
<td>12</td>
<td>24 (0.5)</td>
</tr>
<tr>
<td>T-50-2</td>
<td>1.70</td>
<td>17</td>
<td>26 (0.4)</td>
</tr>
<tr>
<td>T-50-2</td>
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<td>11</td>
<td>20 (0.8)</td>
</tr>
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<td>T-50-6</td>
<td>2.4</td>
<td>25</td>
<td>26 (0.4)</td>
</tr>
<tr>
<td>T-32-2</td>
<td>0.28</td>
<td>8</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-32-2</td>
<td>0.37</td>
<td>9</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-32-2</td>
<td>0.8</td>
<td>14</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-32-2</td>
<td>1.0</td>
<td>22</td>
<td>22 (0.6)</td>
</tr>
<tr>
<td>T-32-2</td>
<td>2.6</td>
<td>25</td>
<td>26 (0.4)</td>
</tr>
<tr>
<td>T-32-2</td>
<td>3.0</td>
<td>27</td>
<td>26 (0.4)</td>
</tr>
<tr>
<td>T-32-2</td>
<td>6.0</td>
<td>37</td>
<td>28 (0.3)</td>
</tr>
</tbody>
</table>

Note: T-94 coil is 0.94 inch dia. (24 mm)
T-80 core is 0.79 inch dia. (20 mm)
T-68 core is 0.69 inch dia. (17.5 mm)
T-50 core is 0.50 inch dia. (12.7 mm)

Perhaps this little gadget is not the best way to measure inductance, but I find it does the job for most of my construction projects, especially for ham-band inductances.

Another reference for a final check of your dial: ARRL L/C/F Calculator Type A.
CADDILLAC OF GSL CARDS, 3 to 4 colors, send #1 for cards (Refundable). Mac's Shack, P.O. Box 43175, Seven Points, TX 75153.

MOBILE OPERATORS: Anteck's Mobile Antennas cover 3.2 to 30 MHz inclusive, with no coil changing, 50 Ohms input. Two models, the MT-1 MANUAL, MT-1R REMOTE-TUNED from the operators position. Uses two Hyd. Pumps and Motors. MT-1 $192.95, MT-1RT $240.00 plus UPS postage. Check with your local dealer or write for Dealer List and Brochure. ANTECK, INC., Route One, Box 415, Hansen, Idaho 83334. 208-423-4100.

WANTED: Micor and Mstr II Base Stations 406-420 and 450-470 MHz. Also 2 and 6 GHz solid state microwave equipment. AKT-7, A 4 Ajax Place, Berkeley, CA 94704.

AMP-LETTER: Dedicated to designing, building, and operating Amateur Radio Amplifiers. Sample #2.00, AMP-LETTER, RR Box 39A, Thompsonville, IL 62090.

TUBES, TUBES wanted for cash or trade. 304TL, 4CX1000A, 4PR56C, 7F7, 7F7N, 53, 6L6M. Any power or special purpose tubes of your choice. OCC, 111 Scuyler Avenue, No. Arlington, NJ 07003. (201) 526-1270.

HAM RADIO FANATICS! You need the W5YI Report - Twice monthly award-winning insider newsletter. 24 issues - $16.00. Sample Issue SASE (2 stamps). W5YI, Box #10101-H, Dallas, TX 75207.

KENWOOD TS-820S, two years old. Excellent condition. Call after 6 PM. (203) 348-3663.

ATLAS D6C Digital Dial $120.00 plus $4.00 UPS. NEW, while they last. Mical Devices, P.O. Box 343, Vista, CA 92083.


IMPROVE MORSE INTERPRETATION — automatically. Fully integrated microelectronics hardware, software. Unusual features. $29.00. Telecraft Laboratories, Box 1185, E. Dennis, Mass. 02641.

RTTY JOURNAL — EXCLUSIVELY AMATEUR RADIO TELETYPE, one year subscription $7.00. Beginners RTTY Handbook $5.00. RTTY Index $1.50. P.O. Box RY, Cardiff, CA 92027.

CW FILTER, 250 Hz, for all Yaesu FT-101A through FT-847 models. Excellent condition, $10. Also SSB filter, $10. John Szkubik, KJS1, 711-106 Avenue, Naples, FL 33940.


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WANTED: Pre-1950 TV sets and old TV Guides. Jeff Kedr. W3CRX, Box 90-HR, Rockville, MD 20850. (301) 654-1876.

RTTY FOR SALE: Several machines remaining. Model 15, Model 30. $200, model 100, $300. Also 500H, 1400, 1600, 44C. $500, 1600, 44C. $500, 1600, 44C $500. Model 28 demodulator, $250. For $250. Telecraft Laboratories, Box 1185, E. Dennis, Mass. 02641.


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MICROVERTER VUC-36 Converts Mid and Superband Signals to UHF Channels 43 to 83. Allowing more channels to be tuned without costly separate selector boxes. Rated #1. Accessory kit available for $2.00, consisting of matching transformer plus 2 jumper cables. Call or write for Free Catalog

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Indoor Matching XFRM $1.25
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April 1982
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pression accessories. Literature. Estes Engineering, 300
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I WOULD LIKE TO GET a copy of the manual, circuit dia-
gram, crystal information for the Standard SRU146, 2
meter handheld transceiver. I am willing to pay any rea-
sorable compensation. Any information where this
documentation can be obtained, like club, library, etc.,
will be appreciated. Dennis Stadan, VE1BJZ.

QSL ECONOMY: 1000 for $13. SASE for samples. WATG,
Box F, Gray, GA 30112.

TEKTRONIX 661 SCOPE: 451 s.t. sampling, 511 time
base plugs in, dc - 1 GHz $250. 12 - scope camera $175.
A $600 spectrum analyzer w/ manual "as is." $125.
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SEND 9% SASE for surplus parts and equipment catalog.
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MEISSNER SIGNAL SHIFTER, manual, coils, best offer
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Drive, Port Richley, FL 33568.

OVERPRINTED — 1981 Fox-Tango Club Newsletters.
Sixty loose-leaf pages packed with modifications and in-
formation on Yaesu rigs. Only $6 while they last. Also a few
1980 sets at $5. (Overseas add $3 deposit, airmail) N4ML,
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SALE — HW-16 1125 (w/crystals) $150; 2 Johnson match-
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meter mobile amp $30. 10.40 vertical $25. You pay ship-
ing K4AEW, Jim Howell, 15 Dan Street, Salisbury, NC
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minute buying your earth station. Six pages of what's needed.
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Tract sure to please. Clyde Stanfield, W4HEG, 1570 N.
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to build your own receiver. For more information write:
Ron Coleman, RT, 3 Box 58-AHR, Travelers Rest, SC
29690.

WANTED: Early Hallicrafter "Skyriders" and "Super Sky-
rider" with silver panels, also "Skyrider Commercial",
early transmitters such as HT-1, HT-3, HT-19, and other
Hallicrafter gear, parts, accessories, manuals. Chuck
Dachs, WD5EUG, The Hallicrafter Collector, 4500 Rus-
sell Drive, Austin, Texas 78745.

WANTED: Surprise 1-3 KW Transmitter type FRT-17,
Collins TDI or equivalent, including parts. G.W. Plummer, 2 Lake Ave.,
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April 1982

WANTED: Early Hallicrafter "Skyriders" and "Super Sky-
rider" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-3, HT-19, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachs, WD5EUG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

WANTED: Surprise 1-3 KW Transmitter type FRT-17, Collins TDI or equivalent, including parts. G.W. Plummer, 2 Lake Ave., Danbury, CT 06810, WA1LDU.

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GET OFF YOUR EARPHONES and start laughing. "Slices of Ham" is a light-hearted look at Amateur Radio. N1ADX has collected more than a dozen short stories to brighten your shack. For witty, technical humor, this must be read. Add $1 armlace overseas. Massachusetts residents add sales tax. Pless Press (H), Box 94 Turnpike Station, Shrewsbury, MA 01545 USA.

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April 1982
MASSACHUSETTS: A general Amateur Radio outdoor event was held at Flea Market, May 2, sponsored by the New England Amateur Radio TV Group, Freeport Hall, Dorchester, just off S.E. expressway, Rain or shine. Admission: $10.00. Safety information is enclosed. Help is requested by April 25 to NET, PO Box 406, Boston, MA 02102. Start at 7 PM. Talk in on 145.250 Repeater or 52 direct.

MASSACHUSETTS: The Wellesley Amateur Radio Society's annual auction sale, Saturday, April 17, Wellesley High School Cafeteria, Rice Street, Wellesley. Talk in on 63.03 and 52 direct. Check or money order only to Wellesley Amateur Radio Club c/o Ann M. Carro KATDIN, 652 Old Colony Terrace, Tiverton, RI 02878.

MINNESOTA: The Arrowhead Radio Club's annual swapfest, Saturday, May 8, First United Methodist Church, 230 East Skyline Parkway, Duluth. Admission: $2.00. Door prizes include a license, $100.00 in cash, and 350000.00 in merchandise by the Blue Ridge Amateur Radio Society. Admission: $1.00; $2.50 for parking. Door prizes include a license, $100.00 in cash, and 350000.00 in merchandise.

NEW HAMPSHIRE: The Great Bay Radio Association's annual Hamfest-Flea Market, Saturday, April 17, Portsmouth Armory, Portsmouth. Admission: $2.00 advance; $2.50 door. Door prizes include a license, $100.00 in cash, and 350000.00 in merchandise.

NEW JERSEY: The Tri-County Radio Association's annual Indoor Hamfest-Flea Market, Saturday, May 2, 9 AM to 4 PM, Parsippany Trade Center, Parsippany. Admission: $2.00 advance; $2.50 door. Door prizes include a license, $100.00 in cash, and 350000.00 in merchandise.

NEW YORK: The Southern Tier Amateur Radio Club's 23rd annual Hamfest, Saturday, May 1, Owego Town Hall, Owego. Admission: $1.00. Door prizes include a license, $100.00 in cash, and 350000.00 in merchandise.

SOUTH CAROLINA: The Greenville Hamfest sponsored by the Blue Ridge Amateur Radio Society will be held at the American Legion Fairgrounds, White House Road, 1/2 mile north of I-85 in Greenville, May 1. Admission: $5.00. Safety information is enclosed. Help is requested by April 25 to NET, PO Box 263, Dayton, OH 45401.

TENNESSEE: The Memphis Mini-Fest, Saturday, April 3, 8 AM to 5 PM, Pipkin Bldg, Mid South Fairgrounds. Admission: $10.00. Flea Market space $5.00 or 2 spaces $8.00. Bring your own tables/chairs. Hospitality party.
OPERATING EVENTS

"Things to do..."

APRIL 10: The Missouri Valley Amateur Radio Club's third annual Pony Express Day, from 1000 CST to 1900 CST commemorating the original running of the Pony Express from St. Joseph, Missouri to Sacramento, California. In addition, the Club is offering a "wanted" poster of outlaw Jesse James as this will be the 100th anniversary of his death. Anyone communicating with W9HNN is eligible to receive both certificates. Phone frequencies: 10 Kc's from bottom of general phone bands on 15, 20, 40 and 75 meters. 10 meters - 28.575, CW: 28 156.75, 120, and 170. 40. To receive certificates send two first-class stamps and a QSL card to Missouri Valley Amateur Radio Club, W9HNN, 401 N. 12th Street, St. Joseph, Missouri 64501.

APRIL 17 & 18: A Novice Mini-Expedition. The North Texas High Frequency Association will be operating the Novice bands from Novice, Texas, from 1800Z Saturday to 1900Z Sunday. Call signs will be on the air on 80 meter CW and 75 meter SSB. For a certificate for confirming contact or reception (free to each station or short wave listener sending a QSL) send to W9HNN, 401 N. 12th Street, St. Joseph, Missouri 64501.

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The conventional air-gap lightning protector design affords virtually no protection to solid-state components, since its breakdown voltage and response time are unpredictable because of the arc-forming characteristics of the molecules of air between electrodes in an air-gap protector.

By comparison, the Alpha Delta Transi-Trap gas tube protector design encloses the gap in a hermetically sealed ceramic tube filled with an isotope of known breakdown characteristics and response time. As a result, the variable and unpredictable nature of the ionized particles is minimized. This yields a protector design with a known response time of 100 nanoseconds and a predictable breakdown voltage with a tolerance of ±15 percent, compared with the approximate breakdown tolerance of 100 percent for the air-gap. Also, breakdown voltage and response time are not affected by humidity, temperature, altitude, and pressure changes; they are with the air-gap.

The Transi-Trap protector design allows devices to be set to fire at the lowest possible lightning pulse level for maximum protection of solid-state receivers and transceivers (Model R-T protector), or at a higher voltage level for protection of amplifiers, both tube-type and solid-state (Model HV protector). By using special constant-impedance brass tubing design for the in-line circuitry, excellent performance through 500 MHz is realized (typically 0.1 dB loss at 500 MHz).

For more information, contact Alpha Delta Communications, 116A North Main Street, Centerville, Ohio 45459.

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Mobile radio operators who use their equipment in the base station will find their job made easier by the new Big Ben Mike Stand by Valor Enterprises, Inc. The new mike stand allows the operator to convert his mobile microphone to a base station microphone.

Valor manufactures Big Ben microphone stands in both black (Model 221) or polished chrome (Model 221C). The stand is part of a complete line of personal and Amateur communications products and accessories offered by Valor Enterprises, Inc., West Milton, Ohio 45383. A complete catalog is available by calling 513-698-4195.

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The Commsoft Codem, a universal CW interface for personal computers, is now available for Radio Amateurs. The Codem provides an easy way to get your Morse code software on the air. Converting received CW audio to RS232 or TTL signal levels and RS232 or TTL signal levels to transmitter keying, the Codem doubles as a code practice oscillator and CW regenerator.
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**THE RADIO AMATEUR ANTENNA HANDBOOK**
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Contains lots of well illustrated construction projects for vertical, long wire, and HF/VHF beam antennas. There is an honest judgment of antenna gain figures, information on the best and worst antenna locations and heights, a long look at the quad vs. the yagi antenna, information on baluns and how to use them, and new information on the popular Stoper and Delta Loop antennas. The text is based on proven data plus practical, on-the-air experience. The Radio Amateur Antenna Handbook will make a valuable and often consulted reference. 195 pages. ©1978.
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A sharp 800-Hz bandpass filter, a.m detector and low pass filter are designed into the Codem to provide outstanding noise and QRM rejection. CW can be monitored using an internal 2-inch speaker or with an external high impedance earphone. Front panel sensitivity, tone and volume controls are provided. The Codem comes with a comprehensive manual that includes operating details and connection instructions. The Codem requires an external 9-Vdc power supply.

The price of the Codem is $124.95. The 9-Vdc supply is $9.95. Add $5.00 for shipping and handling. California residents add applicable sales tax. VISA and MasterCard orders accepted.

For more information, contact Comsoft, 665 Maybell Avenue, Palo Alto, California 94306; telephone 415-493-2184.

**Hy-Gain tribander**

Hy-Gain introduces the TH7DX, a broadband tribander based on the excellent front-to-back characteristics of the older TH6DXX plus the superior VSWR characteristics of a dual-driven element system. The combination produces an amazingly efficient broadband tribander without compromises.

During the development of the TH7DX, the company's engineering tests and research indicated that a higher average front-to-back ratio could be maintained on each band by employing a combination of trapped and monoband reflectors and directors rather than fully trapped parasitics. Also, the gain bandwidth was broader and average half-power beam width was smaller. Research also showed that other tribanders sacrificed gain and high front-to-back ratio to maintain a low VSWR across each band. And finally, none of the tested antennas covered all of the 10-meter band; most stopped at 29.2 or 29.4 MHz.

The new TH7DX features a dual driven element system that maintains a VSWR of less than 2:1 on all bands, including the entire 10-meter band. Both elements use Hy-Gain's efficient Hy-Q traps capable of handling power levels well in excess of the legal limits with a 2:1 safety margin. These traps permit element lengths of 0.225 wavelength on 10 meters, 0.203 wavelength on 15 meters, and 0.186 wavelength on 20 meters. The dual driven elements are fed directly with Hy-Gain's 50-ohm BN-86 balun. Hy-Gain's Beta Match provides a dc ground and matches each band to a

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<th>ISOTRON 40</th>
<th>ISOTRON 20</th>
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<td>54 IN HIGH</td>
<td>31 IN. HIGH</td>
<td>17 IN. HIGH</td>
</tr>
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April 1982
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new products

VSWR of less than 1.5:1 at resonance. Rugged phasing lines and preformed feed straps facilitate easy assembly and consistent results.

The TH7DX, complete with stainless steel hardware, BN-86 balun and heavy duty boom-to-mast clamp, is priced at $499.95. For more information, contact Telex Hy-Gain, 9600 Aldrich Ave., So. Minneapolis, Minnesota 55420; telephone 612-884-4051.

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The ESR24 earth station receiver has been introduced by the R.L. Drake Company, Miamisburg, Ohio. This 3.7-4.2 GHz receiver is designed for satellite television reception and features digital channel display, pre-set and variable audio subcarrier selector, AFC for stability, and full metering. For installation versatility, the down converter module (supplied) may be mounted internally or at the antenna. Accessories for the ESR24 include a remote control, a remote tuning meter, and splash-proof housing. Attractive styling makes the ESR24 suitable for commercial or private installations. Price is under $1,000.

R.L. Drake is recognized for high technology Amateur Radio, commercial, and maritime communications equipment. For more information, contact R.L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342; telephone 513-866-2421.

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April 1982
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(206) 775-7373

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Bring your family and enjoy a great weekend in Dayton.
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CODE PRACTICE TAPES FROM HRPG — Practice copying Morse Code anytime, anywhere. Whether you're upgrading your present license or just trying to up your code speed, a large assortment allows you to choose exactly the kind of practice you need.

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35 wpm code for 20 minutes
40 wpm code for 20 minutes

HR-SC3 - $4.95
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30 wpm code for 20 minutes
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Telephone: (305) 661-5534

Palomar Engineers
1924-F W. Mission Rd., Escondido, CA 92025
Phone: (714) 747-3343

NEW!
220-MHz model now available.

- Portable, Collapsible.
- Folds into its own base for portability.
- For backpackers, apartment dwellers — an antenna that stores compactly when not in use.

New portable quad extends the range of low power two meter transceivers by providing the gain and front-to-back discrimination of a two element quad. Gives the gain of a linear amplifier but does not require additional battery power. Users report full quieting on repeaters that are marginal with 5/8 wave whips.

The entire beam slips into an 18” carrying case to go in your suitcase. For use, it unfolds to form a two element full size quad complete with stabilized mounting bracket. Patented design lets you set it up or take it down in minutes. See the cover article QST September 1980 for full details.

ORDER DIRECT OR FROM YOUR FAVORITE DEALER
Model A-502 portable 2-meter quad or Model A-505 portable 220-MHz quad $87.50. Add $3.00 shipping/handling. California residents add sales tax.

GLB HIGH PERFORMANCE PRESELECTORS

MODEL P50 to P500

- 50 - 500 MHz
- Ultimate rejection over 80 dB
- Five large helical resonators
- Low noise
- High overload resistance
- Typical rejection figures:
  - 600 kHz at 144 MHz: -50 dB
  - 1.5 MHz at 220 MHz: -40 dB
  - 5 MHz at 450 MHz: -45 dB
- The solution to interference, intermod and desens problems on repeaters
- 12V DC operation
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We sell a complete line of transmitters and receivers strips and synthesizers for Amateur and commercial use. Write or call for our free catalog.
Food for thought.

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

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

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

**Group A**

<table>
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<td>118.8 ZB</td>
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<td>151.4 5Z</td>
<td>203.5 5M</td>
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**Group B**

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

**Model TE-64 $79.95**

**COMMUNICATIONS SPECIALISTS**

426 West Taft Avenue, Orange, California 92667
(800) 854-0547/California: (714) 998-3021
FT-230R: QUITE A SIGHT! (AND EASY TO SEE, TOO!!)

Sporting an all-new Liquid Crystal Display, the FT-230R is Yaesu's high-performance answer to your call for a very affordable 2 meter mobile rig with an easy-to-read frequency display! The FT-230R combines microprocessor convenience, a sensitive receiver, a powerful yet clean transmitter strip, and the new dimension of LCD frequency readout. See your Authorized Yaesu Dealer today — and go home with your new FT-230R!

- LCD five-digit frequency readout with night light for high visibility day or night.
- Two VFOs for quick QSY across the band.
- Ten memory slots for storage and recall of favorite channels.
- Selectable synthesizer steps (5 kHz or 10 kHz) in dial or scanning mode.
- Priority channel for checking a favorite frequency for activity while monitoring another.
- Unique VFO/Memory Split mode for covering unusual repeater splits.
- Up/Down band scan plus memory scan for busy or clear channel. Scanning microphone included in purchase price.

And don't forget! Yaesu has a complete line of VHF and UHF handheld and battery portable transceivers using LCD display!!!

Price and Specifications Subject To Change Without Notice or Obligation

FT-290R - 2 Meters
SSB/CW/FM Portable

FT-690R - 6 Meters
USB/CW/AM/FM Portable

FT-208R
FM Handheld
2 Meters

FT-708R
FM Handheld
70 cm
EIMAC's 4CW300,000G Power Tetrode. A new generation of high-performance power tubes.

EIMAC's 4CW300,000G combines all the desired features transmitter designers look for: high peak plate current, low grid emission, low internal capacitance and low internal inductance. This is the first of a new generation of high performance power tubes for LF, HF, VHF and pulse service.

**Laserlab pyrolytic graphite grids**
The control grid and screen structures of the 4CW300,000G are precision-cut by a laser beam. Each element is monolithic and combines extremely low coefficient of expansion with low structural inductance. These features permit the 4CW300,000G to have a very high transconductance—10⁶ micromhos—and allow efficient, high-frequency operation.

**Rugged mesh filament**
The EIMAC mesh filament provides exceptionally high peak plate current and permits low plate voltage operation. This leads to power supply economy, making the 4CW300,000G the economic choice for 300 KW AM broadcast service or long-pulse switch service, each of which demands a reserve of peak emission.

**Improved anode structure**
EIMAC's multi-phase cooling technique provides high plate dissipation to extract heat evenly and quickly from the anode, contributing to long tube life and operating economy.

**EIMAC expertise**
EIMAC's expertise in electron ballistics pyrolytic grid production, thermodynamics and circuit techniques combine to bring tomorrow's tubes for today's transmitter designs. More information is available from Varian EIMAC. Or the nearest Varian Electron Device Group sales office.

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