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DECEMBER 1977

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ham radio magazine

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During the holiday season it's customary to take stock, to look back over the past year, and to make our resolutions for the next — resolutions, no doubt, which will be forgotten by the time the snow melts from the landscape and the trees begin to show their buds. The long winter nights are also a good time to plan that new antenna system or to dream about some new station equipment. With the snow swirling up to the window sills and the cold winds howling down from the north, perhaps it's a good idea to take some time to think about where amateur radio has been, and where it's going.

With the World Administrative Radio Conference (WARC) of 1979 now less than two years away, I can't help wondering what our amateur bands will look like in the 1980s. Will amateurs be given some of the additional high-frequency bands requested by the WARC planning committees, will the width of the amateur bands be pared down, or will we lose much or all of our high-frequency allocations? Nobody will know the answer to that until the final votes are tallied in 1979, but I suspect it will fall somewhere between the two extremes.

There are some who would have you believe there will be *no* high-frequency amateur bands after 1980, and very little vhf spectrum either, but I'm more optimistic than that. Optimism, unfortunately, leads to apathy and that, my friends, is our worst enemy. Perhaps it's best to prepare for the worst and approach WARC '79 with cautious optimism.

It must be remembered that the last international conference which had much effect on the highfrequency spectrum was held in 1947 when the United States and our Allies had considerable influence on the 50 member countries of the United Nations. Radio amateurs were highly regarded by our government for the part they played in war-time communications — not as amateurs, but because they provided a pool of trained technicians and communicators. To a lesser extent the same thing was true in Britain and the Soviet Union. Radio Amateurs were also the backbone of the communication networks set up by the resistance movement in Europe, and of the coast watchers in the South Pacific.

Governments which had severely curtailed amateur radio before the war now recognized its great potential as a national resource. Amateur radio was no longer considered a nuisance to be tolerated, but an activity which should be encouraged. Part of that encouragement was a new, exclusive 15-meter band. Old timers will hasten to point out that bits were shaved off the top ends of 10 and 20 meters, and 160 meters was dominated by Loran, but most amateurs agreed that 15 meters more than made up for the losses.

By the time the next ITU conference on high-frequency allocations was convened in 1959, the United States' sphere of influence had decreased and it looked like amateur radio was in serious trouble; the foreign broadcasters wanted big chunks of 40 and 80 as well as portions of 20 and it was uncertain if we could rally enough votes to save amateur radio. Fortunately some of the nations who weren't particularly friendly toward the United States but supported amateur radio came to the rescue, with the result that the amateur bands in the Western Hemisphere came through unscathed (amateurs in other parts of the world lost 50 kHz of shared space on 40 meters).

In general, the United States and other governments which were supportive of amateur radio in 1959 still are, but in the 20 years since that last conference the balance of power has changed; the emerging nations are now in the majority and they are not altogether in favor of amateur radio — a few ban it outright. Many of these nations have few amateurs, so to them the amateur bands represent wasted space — space they feel should be allocated to a radio service that better serves their national interest. These are the same countries which often oppose the policies of the rich western nations simply because it's in the vogue to do so.

Nevertheless, there's still hope, because many of the questions to be asked at WARC '79 will be answered on the basis of their scientific merits. There's bound to be a certain amount of political arm twisting, but if the delegates from the emerging nations can be made to see the value of an amateur radio service to the technological development of their nation, perhaps they can be persuaded to vote in favor of increased amateur spectrum.

Jim Fisk, W1HR editor-in-chief



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ARRL'S "CODE OF ETHICS" has been challenged in a formal written complaint filed with the Federal Trade Commission's Bureau of Competition.

<u>Specific Complaints</u> are that the Code will violate anti-trust laws by restraining trade, constitutes a "deceptive practice" as defined by the FTC, violates the First and Fourteenth Amendments to the Constitution, and that vendors who sign the ARRL pledge will become accessories after the fact in the above violations.

FCC'S DROPPING OF DOCKET 19759, the proposal that the 220-MHz Amateur band provide a home for a new CB service, doesn't mean that the band won't still become the new CB home. It does remove the immediate threat to the band, however, and many opinions have it that the longer the decision on where CB should go is delayed, the less likely 220 becomes as a choice.

In Announcing Its Termination of Docket 19759 the Commission pointed out that so many changes in related circumstances have occurred since several thousand comments were filed on it back in 1973, those comments were now obsolete. However, the question of a new CB band and where to put it is still very much alive, and 220 will undoubtedly be one of the options when the Commission considers the issue again in a future rulemaking.

A PETITION FOR RECONSIDERATION of the FCC's Report and Order on repeater deregulation (Docket 21033, Presstop, November), is being prepared by the ARRL. In it three issues will be emphasized: restoration of the WR-prefixed callsigns for repeaters, restoration of repeater licenses, and the need for formal consideration of the needs of the so-called "weak signal" vhf/uhf operations. <u>Plenty Of Support</u> for the League position appears likely, as many repeater groups al-ready oppose the dropping of repeater callsigns and licenses. In addition, reservations over the repeater sub-band expansion and even the proposed new bandplan for 144.5 MHz is starting to build among FM users as well as various SWOT and other VHF/UHF

144.5 MHz is starting to build among FM users as well as various SWOT and other VHF/UHF user groups.

AMATEUR LICENSING IRREGULARITIES will receive a full-fledged investigation run by an FCC Administrative Law Judge. The decision to go all-out on such a probe was reached at a closed meeting of the Commission, when information was presented that some Amateurs had apparently paid for the issuance or upgrading of their licenses or for special callsigns; that some of the same abuses may have occurred without payment; and that some Amateur callsigns have been issued inconsistent with normal FCC procedures.

ARTHUR C. CLARK, the noted science-fiction writer, was made an honorary AMSAT member in ceremonies attended by most of the AMSAT brass — Clark's honor came in recognition of his predictions of communications satellites and synchronous satellites in a 1945 Wireless World article.

1978 Orbital Prediction Booklets for OSCAR 7 only will be available shortly from Skip Reymann, W6PAJ, Box 374, San Dimas, California 91773. They're free to AMSAT Life members who request them, \$3 to AMSAT Annual members and \$5 to non-members — be sure to

include AMSAT membership number and an self-addressed label with orders. <u>OSCAR 7's Mode Schedule</u> will be changed effective January 1 to two days in Mode B for every day in Mode A, and the new schedule will be shown in W6PAJ's orbital calendar com-ing out in December. Ample Mode A operations will be provided by the Russian's "RS" spacecraft and AO-D, and OSCAR 7 is considerably more sensitive in Mode B than it is in Mode A.

A New Satellite Bandplan is also going into effect January 1 which will place CW only on the bottom third of the satellite <u>downlink</u>, mixed CW/SSB in the center third, and <u>SSB</u> only operation on the top third — the reverse of current practice. <u>OSCAR 6's Fifth Birthday</u> was October 15, but revival efforts from VE3SAT failed to bring any response from it. RIP.

The Amateur Space Program made the Congressional Record in October when K7UGA lauded it during a Senate discussion of Sputnik's 20th anniversary.

THOSE PACIFIC AND CARIBBEAN prefix changes may not be as drastic as originally announced. FCC's news release announcing the change has now been "cancelled," and while it appears that prefix changes will still be made they'll be done in such a way that the resulting callsigns should identify the individual islands or island groups (Presstop, November).

FCC'S "GAG" ON DISCUSSIONS of current matters will remain in place as a result of the Supreme Court's decision not to review the Court of Appeals decision in the "Home Box Office" case (Presstop, June).

WESTINGHOUSE SCIENCE TALENT SEARCH is open to any high school student in the United States and Puerto Rico who'll graduate before October 1, 1978. Teachers who have an out-standing student who'd qualify for one of the many scholarships and awards must request entry materials from Science Service, 1719 N. Street, Washington, D.C. 20036 — entries are due by December 17, 1977.

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Dave Olean, K1WHS, with his 160 Element DX-Array and Polar Mount EME System

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1-1 52-ohm Balun	DX-1BN	\$12.95	DX-2BN	\$12.95	DX-4BN	\$12.95
Vert. Pol. Bracket (20 El.)	DX-VPB	\$9.95	DX-VPB	\$9.95	DX-VPB	\$9.95



present-day receivers

some problems and cures

Some thoughts on and cures for problems encountered in modern amateur communications receivers

The modern-day communications receiver is going through a continuous evolution that has brought about significant improvement in certain operating features. Among these are greatly improved frequency stability and setability, better selectivity, a slow and consistent tuning rate from band to band, and a wide-range automatic gain control system that functions on CW and single sideband. At the same time, unfortunately, the design philosophies which have made the above advances possible have also reduced the typical receiver's ability to simultaneously handle weak desired and strong undesired signals. This absolute reduction in receiver dynamic range has occured at the same time the number of high-power signals on the amateur bands has been increasing.

Insufficient dynamic range in a receiver can result in one or more stages being over-driven into nonlinearity by undesired strong signals. The result is internally-generated intermodulation distortion (IMD) products. These undesired products can occur in any mode of operation, but are easiest to identify on CW. Two CW signals which are overdriving a receiver will generate IMD products, but only when both stations are transmitting simultaneously. In the extreme situation, not only may IMD occur, but one signal alone can block, deaden, or desensitize the receiver.

In a pileup or contest situation, many strong CW stations can cause serious receiver overload, intermodulating with each other, and resulting in multiple phantom signals; it will appear as if several operators are randomly tapping their keys, or that you are listening to the Novice band with a diode detector without a BFO.

Two or more ssb signals with the correct frequency relationship can also intermodulate with each other and result in IMD products on top of the station you are listening to. The interference, however, will be unintelligible. IMD can also occur from a single ssb station on an adjacent channel as the individual speech frequencies mix with their own harmonics. Generally speaking, transmitted IMD from an rf power amplifier will be worse than that internally generated in the receiver, with the result that the transmitted IMD may cover up a receiver's shortcomings. An operator may never be certain whether the unintelligible signals he hears are being generated within his receiver, or coming from the outside there is enough rf interference to contend with without the receiver creating its own!

The improvements mentioned in the first paragraph have been generally obtained by using a double- or triple-conversion scheme, plus a nonbandswitched master oscillator (PTO or VFO). Depending on the design technique, the first i-f may have a bandwidth of as much as 500 kHz, as in the Heath SB-104, or as narrow as 6 kHz in the Drake R-4B. Assuming that most of a receiver's selectivity occurs at the second intermediate frequency, you might think that the wider the bandwidth of the first i-f, the greater the chance of picking up more strong signals which could overload the second mixer. Of greater importance than this bandwidth, however, is the *net gain* between the antenna and the mixer that drives the narrow crystal or mechanical filter.

The Collins R-390A, for example, has three mixers and two separate gain stages ahead of its mechanical

By J. Robert Sherwood, WBØJGP, and George B. Heidelman, K8RRH, Sherwood Engineering, Incorporated, 1268 South Ogden Street, Denver, Colorado 80210



fig. 1. Block diagram of the Drake R4C receiver showing the gain redistribution. A shunt across the first i-f amplifier will reduce its gain the same amount as is added after the narrow i-f filters.

filters; it also has a set of elaborate, mechanicallytracked tuned circuits which have high Q and high insertion loss. Thus the net gain from the antenna to the major selectivity-determining elements is low enough to maintain good dynamic range.

Another receiver, the Heath SB-303, has a 500-

One topic that has received considerable attention by amateurs in recent years has been that of receiver performance and design. Many approaches have been covered, from the initial design of the "super receiver" to modification of existing equipment; but to the person with just a casual interest, the reasons behind some designs may not be readily apparent. In fact, the problems themselves may not be noticeable to the ordinary amateur. This article is another in a continuing series that shows you how to recognize the problems in typical modern receivers; in addition, it discusses modifications applied to one receiver and the motives behind these changes.

Of major importance is the reason for the modification. The intent of this article *is not* to prove that one particular receiver is superior to another for whimsical reasons, but to realistically and fairly compare different receivers by presenting test results on comparable circuits. On the basis of the test results, design changes were made in one receiver in an attempt to improve overall performance. You will notice while reading the article that the results are given in very specific terms; this will help you to better understand the basics of receiver performance standards. With this knowledge, *you* will be able to judge the merits of the different receivers on the market and choose one according to your own needs. **Editor** kHz wide first i-f *window*, but unlike the R-390A, it has little selectivity ahead of its narrow filters and too much gain. This results in higher susceptibility to overload from strong signals anywhere in the band, which then cause undesired IMD products to be generated within the receiver.

At the opposite end of the bandwidth scale is the Drake R-4C with its 8-kHz wide first i-f filter at 5645 kHz. This four-pole crystal filter does an excellent job of keeping most of the undesired signals in the band from passing on to a second high-gain mixer. However, any undesired strong signals that do pass through this 8-kHz window can proceed to the second mixer with disasterous results. The net gain from the antenna to the narrow second i-f crystal filter can be as high as 50 dB when a desired weak signal (S1) is being received; this puts an impossible demand on the i-f stages, since the 1-dB compression point of the second mixer output has occured with any signal 30 dB over S9. An undesired signal, outside the narrow selectivity but inside the first i-f window, that is S9+40 dB (-33 dBm or 5 mV)across the 50-ohm antenna input) for example, would have to be linearly amplified to a level of +17dBm (1.58 volts across the 50-ohm narrow-filter input) and then be rejected by the filter. To supply this power level to the filter, the high-impedance plate of the second mixer would have to linearly swing more than 40 volts to yield a signal that is as great as 15 volts rms; even if this level could be produced in a low noise mixer, which is highly unlikely, the filter could be damaged.

What actually results when there are two undesired signals at S9 + 40 dB with the correct frequency relationship, over loading the second mixer, is a spurious third-order IMD signal that is greater than S9 in strength. This would certainly be strong enough to obliterate the desired weak signal!

One possible reason why such net-gain design errors are overlooked is our present method of testing receiver dynamic range. This subject has received considerable attention lately in ham radio^{1,2,3} and QST.⁴ An increasingly popular method of testing for dynamic range has been developed by Wes Hayward, W7ZOI, and is used by the ARRL.5 Basically, it consists of applying two well-isolated, equal-strength signals, 20-kHz apart, to a receiver's input and then adjusting their level so that the undesired third-order IMD products generated within the receiver are just equal to the noise floor of the receiver. The difference in level between the noise floor and the test signals gives the receiver's dynamic range. The higher the receiver's dynamic range, the better it can handle both desired weak and undesired strong signals at the same time.

The choice of 20-kHz spacing for the two test signals is arbitrary and in many cases satisfactory. In a receiver which has all its significant selectivity far



Installation of the 600-Hertz first i-f filter. The filter is installed on a vertical shield near the original 8-kHz filter. The devices with 8 leads are TO-5 size relays that are used to select the appropriate filter.

down the i-f chain, this signal spacing is relatively unimportant. If the early-stage bandwidth is narrower than the test signal spacing, however, its selectivity will partially or completely reject one or both of the test signals, resulting in a highly inflated dynamic range reading. We feel these measurements should cover worst-case conditions since real-life interference on the amateur bands may be spaced less than 20 kHz.

Third-order IMD products, with 20-kHz spacing, will occur 20 kHz below the low frequency test signal and 20 kHz above the high frequency test signal. When the receiver is tuned to a third-order internallygenerated spurious IMD signal, the test signals are 20 and 40 kHz up or down the band. The 25-kHz-wide crystal filter in the first i-f of the Signal-One transceiver, to name just one example, will greatly attenuate the test signals before they can reach the following stages. Thus, 20-kHz spacing will test only the front end and first mixer. What is needed is spacing narrow enough so that both test signals can pass through any selectivity prior to the narrow filter. We feel a spacing of 2 kHz will satisfy this requirement, and at the same time be wide enough so the narrow filter will adequately reject the test signals when the receiver is tuned to an IMD product.*

The Drake R-4C, with its 8-kHz-wide first i-f filter, shows an inflated 20-kHz dynamic range of 83 dB. This reading has remained quite consistent over several receivers, including one we tested at the ARRL laboratory.† When the test signals are placed 2 kHz apart, however, so they *both* pass through the 8-kHz filter, the dynamic range drops to around 58 dB.

improving receiver performance

There are three ways to improve a receiver's dynamic range. If the second mixer cannot handle the required level, one option is to replace it with a mixer that will do the job. Unfortunately, as WB4ZNV discovered,⁶ the process of replacing an active mixer with the superior passive double-balanced mixer is a laborious task, even if it does improve the receiver's overload characteristics. Oscillator injection levels and impedances are usually not compatible with existing circuitry.

Another remedy is to redistribute the gain in the receiver, reducing it ahead of the overloaded stage and building it up again after the narrow filter. A third method is to insert more early-stage selectivity into the receiver so strong interfering signals are not as likely to get past the first mixer. We chose to inves-

^{*}When performing a 2-kHz IMD test, one very important factor must be taken into consideration: the noise sidebands of the signal generators. General test equipment, oscillators, or VFOs are more than adequate for testing, until a receiver's dynamic range nears 100 dB. At this point it will be impossible to accurately measure true receiver IMD products if the signal generators are producing excessive low-level spurs and noise. At this time there are only two or three generators that have the necessary sideband suppression; one manufactured by Hewlett-Packard and another by Rohde and Schwartz.

tThe ARRL laboratory uses a pair of AN/URM-25 signal generators to perform IMD tests. A 2-kHz IMD test produced results within 2 dB of those obtained by the authors while using the high quality, low-noise sideband Rohde and Schwarz XUA signal generator.

tigate the latter two options, using our own R-4Cs.

The initial gain redistribution began with a 20-dB reduction of the signal level as seen by the second mixer. This gain loss was then restored after the narrow filters at the high-impedance grid of the third mixer. The original amplifier used a single jfet plus a step-up transformer to provide the necessary gain, but the circuit suffered from instability problems and noise. It was then decided to relocate the added gain outboard from the receiver and insert it at a convenient 50-ohm point, the output of the switchable second i-f crystal filters (see fig. 1).

A cascode jfet amplifier, with 50-ohm input and output impedances (**fig. 2**), was built and inserted into the i-f chain just prior to T-6. The coax cable that connects T-6 and the mode switch was lifted at the switch end; two lengths of miniature coax (RG-174/U) were then run out through a slot in the rear of the receiver. The first length is connected to the lugs on the mode-switch wafer, while the second is spliced into the cable that feeds the transformer.

This amplifier can possibly be located inside the receiver. Regardless of its location, it should be mounted in a metal box or other well-shielded enclosure. Two toroidal transformers provide the necessary impedance changes, their associated trimmer capacitors forming resonant circuits. While both trimmers can simply be peaked for maximum signal, the input may be fine-tuned for the best compromise signal-to-noise ratio among the switchable narrow filters. (The 2N5950 and 2N5953 jfets may be purchased from G. R. Whitehouse Company, Amherst, New Hampshire 03031).

We found the best way to attenuate the signal level into the second mixer was to swamp the output of the first i-f amplifier Q1 (V3/6BZ6 in early receivers). A miniature 5000-ohm multi-turn trimmer, from noise blanker socket pin 4 to ground, made a convenient way to adjust this level. Simply adjust the trimmer to drop the calibrator signal 20 dB on the S-meter; then adjust the gain pot on the cascode amplifier to restore the S-meter to its previous level. On certain receivers it may be necessary to peak T-6 to obtain 20 dB of gain from the cascode amplifier; always readjust both cascode trimmers after making a gain change.

If the noise blanker is installed in the receiver, significant IMD products can occur in its stages, too. Due to noise limitations, however, the blanker cannot be starved a full 20 dB. Instead, after replacing blanker resistor R1 with a 0.001 μ F disc capacitor, reduce the gain to the blanker about 12 dB, and then turn down the blanker output pot 8 dB to achieve the 20 dB reduction at the second mixer. Alternately, the gain of blanker transistor Q2 can be decreased by reducing its emitter resistor bypass capacitor, rather than readjusting the blanker output pot.

Take care not to use too much cascode amplifier or blanker gain; otherwise amplified 5645-kHz oscillator leakage can degrade system performance. With the antenna disconnected and the top and bottom covers of the receiver in place, make sure the S-meter does not kick upward more than one-quarter S-unit when the passband tuning is slowly turned through its range. In some receivers it may be necessary to jumper the cable-braid ground point of the Q4 oscillator board with a short clip lead to the shield tray on which the blanker board rests to reduce this oscillator leakage to an acceptable level. It might also be necessary to insulate the frame of the rear carrier-oscillator jack from the chassis ground.



The new product detector is installed next to the audio transformer and behind the variable capacitor used for passband tuning. The entire assembly is mounted on a $1-3/4 \times 1-5/8$ inch (4.5x4.1cm) board.

Also, if the cascode amplifier breaks into oscillation when the mode switch is between detent positions, reverse the leads of a high impedance winding of one of the toroids.

Proper operation of the gain redistribution circuits provided greatly reduced susceptibility to IMD overload problems on both CW and ssb, as was visibly demonstrated with strong nearby DX contest signals; yet the receiver was still able to meet its sensitivity specification. Agc attack distortion was also reduced somewhat. Dynamic range improved from 58 dB to around 70 dB, while using our 2-kHz spacing test method.

i-f filters

As an additional CW remedy we chose to increase the selectivity (possibly on a switchable basis) following the output of the first mixer; the bandwidth is presently determined by an 8-kHz wide four-pole crystal filter. This bandwidth is needed on phone to pass an upper and/or lower sideband signal. A bandwidth of at least this magnitude is also required to pass undistorted noise pulses to the blanker. A noise blanker's usefulness, however, is marginal at best with one or more strong nearby signals, due to its agc greatly increasing the blanking threshold, or possible false triggering. Thus, the need for narrowing first i-f selectivity ahead of the noise blanker, which reduces blanker effectiveness, occurs under conditions which are usually unfavorable to blanking in the first place.

Circumstances could occur where blanking would be necessary at all times, such as when you suffer from a continuous very high level of blankable noise. In these cases, the 8-kHz first i-f filter must remain ahead of the blanker. Then a properly-terminated narrow filter could be inserted just after the blanker, but before the second mixer. The signal path can be switched between the narrow filter and an attenuator equal to its loss. While the chance of second mixer overload is greatly reduced with this arrangement. there is no such narrow bandwidth IMD protection for the blanker; this limits the receiver's potential dynamic range considerably below what is otherwise obtainable. It is therefore mandatory to use the cascode gain redistribution system with this special, optional filter arrangement. With this arrangement close-in dynamic range will be in the high 70s.

We decided that the first i-f CW selectivity should be equal to the widest desirable under contest conditions. We then designed a new 600-Hz six-pole filter, keeping in mind package size limitations and insertion loss requirements. We've also developed a miniature relay system which allows instant interchange of our internally-mounted, CW-bandwidth, first i-f filter with the existing 8-kHz phone unit.

The project of minimizing overload in the R-4C was now complete and totally successful. When measured using our worst-case 2-kHz test method, the receiver's dynamic range jumped from an original unacceptable 58 dB to a final excellent 85 dB. This value ranks with the best of the commerciallyavailable amateur gear on the market today, and



fig. 2. Schematic diagram of the cascode amplifier used for the gain redistribution. There is only one ground return on the circuit, through the input coax cable. The braid on the output coax cable goes to the primary of T6 which is not grounded at that point. T1 and T2 are wound on Micrometals T-50-2 toroidal cores. The high-impedance windings are 80 turns of no. 30 AWG (0.25mm) while the low-impedance windings are 5 turns of no. 24 AWG (0.5mm).

should be more than adequate for most practical situations. As a side note, a similar arrangement of first i-f filter switching can be used on ssb by inserting a set of 2.6 or 2.3-kHz phone filters in the first i-f for improved phone selectivity.

simple receiver testing

While we made use of a considerable amount of test equipment during this project to measure dynamic range, you can make comparative tests using only a crystal calibrator and transmitter vfo, *loosely* coupled into the receiver. Comparative noise floor measurements, with no antenna connected, can be made by measuring the preselector noise peak (above later stage noise) with an ac voltmeter connected to the audio output line.

When making gain redistribution or selectivity changes, adjust the receiver to maintain its original net gain by measuring the calibrator level on some specific frequency. We use 7.2 MHz as our reference frequency. Here the calibrator level should read about 15 to 20 dB over S9 with nothing connected to the antenna input. (Don't readjust the S-meter sensitivity pot.) Two strong test signals, accurately set to a specific S-meter level, will produce a repeatable reference IMD that can also be measured on the S-meter. As improvements are made the IMD, read on the S-meter, will drop. We made our 2-kHz tests at S9 + 40 dB, and ended up reducing the IMD from greater than S9 to less than S3.

filter rejection

The 600-Hz first i-f filter, in addition to greatly reducing the chance of overload, had the extra benefit of eliminating the annoying signal leakage around the narrow second i-f filters. This problem of not being able to realize the ultimate rejection capabilities of a well-designed filter is one that plagues all equipment that, to our knowledge, is presently on the market. It is really quite difficult to even design a test fixture to correctly measure the ultimate rejection of a filter. Obtaining adequate ultimate attenuation, which should be in excess of 100 dB for an eight-pole filter in a receiver or transceiver, requires tedious attention to detail. Current ground loops and stray capacitive coupling are the main problems that must be eliminated. We have had many frustrated amateurs ask us to provide a filter for their receiver or transceiver which would not leak like the factory installed units. Unfortunately, some of the limitations were in the receiver and not the filter. Although replacing or adding to an existing late narrow filter can often considerably improve skirt selectivity, the only way to eliminate the last traces of these leakage problems, in existing popular receivers, is to add a filter earlier in the set with a

bandwidth closer to that of the main filter. The early filter should preferably be on a different frequency from the later one, such as in the R-4C or 2B.

We tested one all-solid-state American transceiver that had so much leakage around the CW filter that a 2-kHz dynamic range test could barely be made. The IMD was masked by the test signal leakage until special audio filtering was employed.

While discussing filters, we would like to emphasize the importance of a great variety of bandwidths being available to the operator. Most of the equipment on the market has just one standard phone bandwidth, with one CW filter available as an option, and when installed it must be used at all

NEW PRODUCT DETECTOR

with this trade-off, there is an additional insertion loss of 5 to 7 dB compared to the phone filter, and relatively poor skirt selectivity.

As a minimum, the receiver net gain should be designed around the lossiest filter, with the losses of the other filters increased to that constant level. Another school of thought suggests that the noise integrated by each of the filters should be the same, requiring increasing gain (or decreasing insertion loss) as narrower filters are selected. To our knowledge, no amateur equipment manufacturer is currently keeping the integrated noise constant, and only the R-4C provides for constant insertion loss with narrow bandwidth filters.

EXISTING PRODUCT DETECTOR



fig. 3. The MC1496L can be used as a product detector as shown in A. The IC plus associated components are mounted on a small circuit board which is installed next to the audio transformer in the receiver. C1 and C2 are critical values and should not be substituted. For smaller size, the 1- μ F capacitors may be tantalum. B shows the interconnections between the detector and the receiver.

times for that mode. Many of the imported rigs are examples of these limitations. The Yaesu FT-101B has only a six-pole 600-Hz filter, and the Kenwood TS-820 is limited to only a six-pole 500-Hz unit.

By today's standards a six-pole 500-Hz filter is quite broad and has a poor shape factor. One possible reason for offering only these filters is that the design of the equipment was based on the use of an ssb filter having an insertion loss of only 2 to 4 dB. Unless a manufacturer employs special technology in building, say, an eight-pole 350-Hz filter that is more advanced than required for a phone filter, the insertion loss will rise to an unacceptable 14 to 16 dB. It is quite undesirable to have the signal drop 12 dB when the CW filter is used; a compromise is made, and the six-pole filters mentioned above are offered. Even We have noted with interest the comments from some of our Japanese and German filter customers about American rigs such as the R-4C and T-4XC. The cost of these units in their home countries, due to import duties, is 30 to 50 per cent higher than here in the United States, but the discriminating foreign amateur is willing to pay that premium partly because of the excellent filters which are available. Compared with the typical filter in the average set, the Drake eight-pole 250-Hz and the Sherwood eight-pole 125-Hz CW filters are valuable assets. Similarly, an optional 1500 to 1800-Hz ssb filter* can make the dif-

*Drake also offers the FL1500, a 1500-Hz filter. Though publicized as an RTTY filter, it provides exceptional performance, especially under difficult phone contest conditions. **Editor**.



Cascode amplifier used for gain redistribution is installed in a small enclosure. The shield must be in place between the stages of the amplifier.

ference in being able to hold a contact under heavy interference and contest conditions.

It takes some practice to become proficient at using a narrow i-f filter, just as in learning to tune with the wide-skirted audio filters. But during crowded band conditions a 250-Hz filter can often be too broad! One CW operator used the 125-Hz filter in his R-4C almost exclusively during the hectic 160-meter contests.

The entire line of filters for the R-4C is excellent and can be adapted to any receiver or transceiver. A construction article in the 1977 ARRL *Handbook*⁷ describes a method of adding bandpass tuning to a receiver lacking this feature. This circuit uses 455-kHz filters and is inserted in the receiver i-f chain by converting down to 455 kHz and back up again. This basic idea can be used with any pair of filter and receiver intermediate frequencies.

You could convert from 3395 kHz up to 5695 kHz and back down again, for example, or down from 9 and up again. As the difference between the two i-f frequencies becomes smaller, the difficulty of the conversion process increases. A Drake R-4B owner who wishes to add R-4C filters to his receiver has to cope with a conversion frequency difference of only 50 kHz. Howard Sartori, W5DA, has developed a circuit for use in his R-4B which can be adapted to any i-f by simply changing one crystal oscillator. It has been used on intermediate frequencies as low as 50 kHz and as high as 30 MHz with excellent results. His circuit is described on page 20 of this issue of ham radio. One precaution, when adapting the Handbook circuit or W5DA's i-f converter to a transceiver: make sure the transmitted signal does not have to pass through the added filters. Otherwise, with use of the two narrowest filters (the FL-250 and CF-125/8), the

transmitter carrier offset frequency adjustment would become quite critical, and keying on the transmitted signal could be too soft.

The Kenwood TS-820, which we have in the lab. has a noise floor and dynamic range in the ssb mode that is virtually identical to that of the Drake R-4C. Both units perform very well on phone; when you want to dig out a weak CW signal on a quiet band. however, the R-4C is significantly better. The R-4C's gain remains constant when a CW filter is switched in, but the TS-820's drops off 5 to 6 dB. Even if a weak received signal is above the noise floor, this gain reduction increases the agc threshold to the point where it may become necessary to manually ride the gain control. The Yaesu FT-101B we tested had a dynamic range, at any test signal spacing, as bad as the unmodified R-4C when measured with the worst-case 2-kHz test method. The bulk of the problems in the FT-101B were caused by a bipolar transistor in the noise blanker which was being overdriven.

A receiver's maximum net gain from the antenna to the detector can change significantly from band to band without having much effect on the measured sensitivity. Two sets with similar signal requirements for a given signal-to-noise ratio can have vastly different capabilities in handling weak, fluctuating signals, especially on the 10- and 15-meter bands. As the net gain falls off, more and more signals will fall below the agc threshold. The R-4C, for instance, holds a much more consistent net gain from 80 to 10 meters than the TR-4C. The TS-820 increases the net gain on 10 meters compared to 20 and 15 by changing a capacitive tap on the rf amplifier drain. Its gain, however, is too high on 160 meters, resulting in a higher susceptibility to overload by broadcast stations. When connected to a nearly self-resonant 160-meter vertical antenna at our lab in Denver, the TS-820 grossly overloads with the eighteen local broadcast stations, developing more than 1 volt across its antenna input. Without the 20-dB rf attenuator switched in, the 160-meter band is nothing but a solid mass of S9 + 30 dB IMD products.

The TS-820's front end is not selective enough to cope with this admittedly unusual receiving situation. On 1.8 MHz, the preselector attenuates signals that are 100 kHz off frequency by 18 dB. In comparison, the R-4C attenuates these same signals by 38 dB. On 3.6 MHz, the TS-820's front end is down 8 dB at 100 kHz off frequency, the TR-4C by 12 dB, and the R-4C by 24 dB. When tested on 10 meters, the *500*-kHz attenuation is 8 dB on the TS-820, 8 dB on the TR-4C, and 15 dB on the R-4C.

One way to eliminate the need for a sharp preselector is to use an up-conversion scheme, with the first i-f above 40 MHz. The input may only need a bandpass filter that rejects signals below 1.8 and above 30 MHz. Then image signals would fall above 80 MHz and be virtually eliminated by the bandpass filter. The first mixer must have a much greater signal-handling capability than in present receivers, however, because it would see all stations between 1.8 and 30 MHz. Two strong local signals, one on 14 and the other on 21 MHz, could produce a 7-MHz IMD product.

The R-4C and the TS-820 show a 20-kHz testsignal-spacing dynamic range in the ssb mode of about 80 dB when tested on 20 meters. At this frequency, the preselectors do not significantly enter into the dynamic range test, since they will not attenuate the test signals more than 1 dB. This is not the case on 160 meters, especially with the R-4C. Here, its high-Q front end attenuates the 20-kHz signals enough to raise the dynamic range by 12 dB. On the other hand, some receivers have too much gain on 80 and 160 meters which, even with sharp preselectors, could yield a dynamic range no better (or even worse) than on 20 meters.

While the 20-kHz dynamic range of the R-4C improves on the lower frequencies because of its preselector, the 2-kHz dynamic range measurement remains quite constant at just under 60 dB. Similarly, it is consistently above 83 dB with the 600-Hz first i-f filter that cures its *window* overload problem. The TS-820 does not have this *window* problem since it is a single-conversion design and has no overloadable stages between the wide noise blanker filter and its narrow filter. Any improvement in dynamic range with increasing frequency separation of the test signals can only be attributed to its preselector.

A detailed review of the TS-820 in *CQ-DL*,⁸ far more comprehensive than anything published in this country, showed a 6-dB improvement in dynamic range as the test signal spacing was increased from 2 to 50 kHz. It is interesting to note that *CQ-DL* also feels that a close-in 2-kHz spacing is necessary for proper evaluation.*

The Atlas 210X, without its noise blanker operational, has a better than average dynamic range of about 90 dB, which would be even better if its double-balanced mixer were properly terminated above the i-f frequency.² This could be accomplished with the use of a diplexer, as described by Wes Hayward,⁴ or with a power jfet, as related by Ulrich Rohde.^{2,3} There is one limitation in the 210X that cannot be easily remedied, however; its potential strong-signal handling capabilities cannot be fully realized due to its noisy conversion oscillator. Since this oscillator has noise sidebands that are only 65 dB down 10 kHz on each side of its center frequency, all

*A recent independent measurement by DJ2LR showed the intercept point of the TS820 to be -12 dBm.

the signals passing through the mixer will take on similar noise sidebands. Consider a strong station near a desired signal that is weaker in amplitude. Reciprocal mixing of oscillator noise can cause noise sidebands to be transferred to the strong nearby station and cause interference to the desired signal. Thus, even if the i-f filter's ultimate rejection is actually realized in the receiver circuitry, which is doubtful in practice, this high level of rejection can be negated by wide-band mixer noise. So while it takes two strong signals to cause IMD which can interfere with weak signal reception, a noisy oscillator and one strong signal can cause the same unfortunate results.⁹

The noise blanker in the Atlas 210X also degrades its dynamic range, diminishing the advantage of the double-balanced passive mixer. The 210X transceivers we tested had a dynamic range of between 73 and 81 dB, depending on the band selected. When the blanker was turned on, these numbers dropped by 3 dB.

There is little reason for a noise blanker to include additional gain stages which can degrade receiver performance. The TS-820 has only a 4-diode balanced blanker gate in its i-f chain; therefore, it does not reduce the overload capability or significantly increase the noise floor. Alternately, a balanced mixer or push-pull i-f stage can be gated for noise blanking; this requires no additional gain stages in the signal path.

product detectors

Another area that could use additional work is that of the product detector. As the name implies, its output should be the product of the two input signals. If



IMD generated at the output of the R-4C second mixer by two 5 mV signals at the antenna input. The signal spacing was 2 kHz. The receiver was tuned so that the narrow second i-f filter was positioned away from any test signals or IMD products. Therefore, with no signal reaching the AGC, the receiver gain is at maximum and the S meter reads S1.

BFO injection is removed, output should go to zero. If this is not the case, as in the Heath HW series, envelope detection is also occurring, which causes audio distortion. On the other hand, the 6GX6 product detector in the Drake R-4, TR-4, and TR-4C, and the 6BE6 in the Drake 2A and 2B, works very well.

Other extraneous outputs can occur even if the detector is acting solely as a product mixer. A detector should be a double-balanced, or other arrangement, which provides good isolation between input and output. The two-diode detector in the R-4B and R-4C is not a double-balanced design and allows the detected audio to leak back and envelope modulate the last i-f stage. This resultant signal is detected in the agc, which then tries to follow it at an audio rate, especially (but not only) when the faster time constants are in use. This audio output sounds slightly distorted, and is noticeable on ssb as well as CW. In addition, BFO injection is marginal, causing additional distortion on AGC attack.

We decided to replace the product detectors in our R-4C receivers, but wanted to use a device that was compatible with the existing drive and impedance levels. The MC1496L active double-balanced mixer looked like a good choice, and with minor circuit changes from the data sheet, was installed in the receiver. The modulation of the i-f by the detected audio was eliminated, resulting in cleaner sounding audio. AGC attack distortion was further reduced.

The MC1496's main drawback is its high number of associated components. Eleven 1/4-watt resistors, nine capacitors, and the IC had to be squeezed on a 1-3/4 by 1-5/8 inch (4.5x4.1cm) board which was nestled between the audio output transformer and the adjacent PC board (see fig. 3). All R-4C owners, whether they change product detectors or not, should add a 0.0015 µF capacitor across R83 in the audio amplifier. This corrects a phase error in the feedback circuit, and eliminates an undesirable peak in the audio frequency response which accentuates harmonic distortion. The Kenwood TS-820 and the Atlas 210X both use a doublebalanced diode product detector that works quite well, and needs considerably fewer parts, but they are low-impedance devices not easily adapted to some circuitry.

conclusions

We have discussed several popular receivers and noted some of their strengths and weaknesses. Some problems can be corrected in the field, while others go beyond the scope of a weekend project. We've also investigated two ways to improve a receiver's susceptibility to overload, so that it can better handle today's high-level rf environment: redistributing the gain and increasing the early-stage selectivity with an additional filter. The importance of having a wide choice of adequate narrow filter selectivity, without leakage, was also mentioned. While most of our circuit changes have been applied to one specific popular receiver, the Drake R-4C, the ideas can be extended to other sets. A method of checking a receiver's overload capabilities which requires no test equipment was also described. Thus receiver changes can be evaluated as to their effect on dynamic range.

The real key to how a receiver performs is its net gain distribution, particularly in relation to the location of selectivity determining elements. A receiver must have a great deal of gain from its antenna to the speaker to be able to receive weak signals. But if too much gain is placed ahead of a narrow filter, the receiver is bound to overload and generate interference of its own.

How a receiver will perform in real-life situations can be determined in the lab, but only if it is tested in a manner that approximates the real world. We feel that the present 20-kHz signal-spacing method can be quite misleading, and should be augmented with our 2-kHz test procedure. If the two readings are significantly different, then further investigation is warranted.

As we stated at the beginning of this article, receivers have improved in many ways, especially over the past 15 years; at the same time, dynamic range has diminished. Amateur radio operators should be demanding excellence in this critical parameter. Improvements in receiver versatility need not reduce system performance, as we have so often observed. Potential problems can be eliminated in new equipment by state-of-the-art design or by retrofitting existing receivers. All that will be lost is some internally-generated rf interference!

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Compare the Atlas 350-XL with other transceivers . . .

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MODEL	ATLAS 350-XL	TEN TEC	YAESU FT-301	DRAKE TR4-CW	HY-GAIN 3750	KENWOOD TS-820	TEMP0 2020
INPUT POWER	350 WATTS	200	200	300	200	200	180
BANDS	10-160 M	10-80M 160M OPT	10-160M	10-80M	10-160M	10-160M	10-80M

. . . and see why it's your best buy!

Above is a chart comparing leading HF Transceivers that fall in approximately the same price range as the Atlas 350-XL. The Drake TR4-CW is least expensive, while the HY-Gain 3750 is the highest. Rated power input (SSB) and bands covered are listed in the chart, but below is a discussion on a number of other interesting comparisons which will help you choose the right transceiver for your station.

1. STATE-OF-THE-ART, ALL SOLID STATE

The first 3 transceivers listed above are all solid state. The real designs of the future! Having manufactured and sold over 12,000 of our little 210x/215x's, we can attest to the high performance and reliability of all solid state design. Tubes for the driver and P.A., with their tuning circuits and high voltage power supplies are rapidly becoming obsolete. As a result their resale value will be declining.

2.POWER RATING.

The higher power rating on the 350-XL provides you with a comfortable edge over the others. Running barefoot you can easily ride over the competition. If you're driving a linear you don't have to strain for every bit of drive from the transceiver. It can loaf along with ease. The 350 watt input rating is really very conservative. Typical input power runs upwards of 400 to 450 watts without flat-topping. Considerably more than the others.

3. BAND COVERAGE

Not only does the 350-XL cover the 10 through 160 meter bands (including all of 10 meters in four 500 kHz segments), but one of its exclusive features is that you can install up to 10 auxiliary 500 kHz ranges anywhere from 2 to 5 MHz, and from 6 to 23 MHz. This gives you great flexibility for MARS operation and possible future amateur bands. Crystals for Auxiliary Ranges are installed internally. In addition, the 350-XL provides reception of WWV at 5, 10, and 15 MHz, without having to add any auxiliary range crystals.



4. DIGITAL FREQUENCY READOUT

On the 350-XL, the optional Digital Dial can be installed, and you still retain the conventional analog dial, with the option of switching the digital dial off if you wish. With the Ten-Tec or Yaesu 301, you lose the analog dial if you purchase the digital dial model, making you totally dependent on the digital dial.

5. FULL BREAK-IN CW

Only two rigs offer this feature; the Atlas 350-XL and the Ten-Tec ! The others are all "semi-breakin". And the Atlas includes CW sidetone with pitch and volume adjustments.

6. NARROW BAND CW FILTER

This is another standard feature in the Atlas, optional on the Ten-Tec, Yaesu, and Kenwood. Ours is an I.F. filter with 500 Hz bandwidth, and shape factor of better than 3 to 1.

7. A.F. NOTCH FILTER

This 350-XL standard feature permits nulling out heterodynes and other interference. The Yaesu, Hy-Gain and Kenwood include a similar feature.

8.SPEECH COMPRESSION

The standard Atlas ALC system provides up to 20 dB of R.F. compression which increases your talk power and at the same time reduces "flat-topping" and splatter. An optional speech processor to provide up to 20 dB additional A.F. compression will be

* We're very proud that every Atlas transceiver is made right here in America, (as are the Ten- *

🗶 Tec and Drake). We think the American worker, and our employees in particular, are the most 🖈

x talented, industrious people in the world. The quality and versatility of our transceivers are proof of this.

And by using this American quality workmanship, advanced value engineering in design and manufacture, and rigid quality control, the Atlas transceiver is not only competitively priced with the imports, but is actually a better value!

Merry Christmas and Holiday Greetings from all the gang at Atlas!

available soon for installation in the AC supply. The Hy-Gain, Kenwood, and Yaesu also provide some form of speech processing.

9. AUXILIARY VFO

All of the rigs listed offer an optional second VFO for split frequency operation. But Atlas is the only one with an Auxiliary VFO that is not an add-on box. The Atlas Auxiliary VFO plugs right into a space provided in the upper right hand corner of the front panel. Although miniature in size it tunes the same 500 kHz as the primary VFO, and does it smoothly with coarse and fine controls that have 10:1 planetary drives. Green, yellow, and red LED's let you know which VFO you have set up for receiving and transmitting. Very neat, and all self-contained.

An option to the Model 305 Auxiliary VFO is the Model 311 crystal oscillator that provides up to 12 crystal controlled channels. It also plugs into the front panel just like the 305. Vernier controls provide fine tuning of the crystal frequency.

10. MOBILE/PORTABLE OPERATION

The Atlas, Ten-Tec , and Yaesu, being solid state, are unique in that they will operate mobile or portable directly from a 12-14 volt DC battery. Also, the solid state rigs are considerably smaller and lighter weight than the hybrid rigs. The Atlas is unique in having a very handy plug-in mobile bracket for the 350-XL that makes it a simple matter to plug-in and go mobile.

11. OTHER 350-XL STANDARD

FEATURES include R.I.T., VOX, Crystal Calibration, ANL, and Noise Blanker.

Compare the Atlas 350-XL SSB-CW Transceiver with the others, and we think you'll agrea the Atlas has everything you'll ever need in a transceiver. And it's made in America.

And let us not forget to mention Our Customer Service which is second to none. Just ask the ham who owns one.

Model 350-XL (less options) \$995. Model DD6-XL Digital Dial \$229. Model 305 Auxiliary VF0 \$155.



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More Details? CHECK - OFF Page 150

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an up/down filter converter -

matches any bandpass filter to any receiver i-f

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Design and construction of an up/down converter that will interface any crystal filter with any receiver i-f. A design example shows how to add a 125-Hz filter to the Drake R-4B

Ultimate receiver performance is viewed by most amateurs as a moving target, with increasing cost just one factor that keeps the target out of reach. New inventions and techniques are constantly being announced by the fast-moving electronics industry; yesterday's dream of an ideal receiver becomes history long before the final receiver payment is due. Giant strides in IC technology have made receivers comfortable and easy to use through the addition of synthesizers, diode switches, and frequency counters. However, most *real* receiver performance improvements, in terms of signal-handling capability and selectivity, are still to be made. The name of the game is picking the weak signal out of the interference caused by many nearby strong signals. Then receiver performance specifications such as third-order intercept point, dynamic range, receiver desensitization, and mixer overload suddenly come to mind.

One goal of modern high-frequency receiver design is to process the desired signal through the narrowest available filter, with the smallest number of active components. Maintaining a credible noise figure, however, tends to legislate against throwing out all of the active front-end components except the mixer.* While giant strides have been made in semiconductor development, filter technology has been advancing rapidly, too.

This article will discuss the use of available crystal filters and will show you how to easily add highperformance filters to receivers without facing the frustrations of mixer design — frequency conversion, loss of sensitivity, and degradation of dynamic range.

filter characteristics

A complete line of filters optimized at the same center frequency for CW, RTTY, ssb, and a-m, particularly the i-f in your receiver, is hard to find at a price you can afford, especially from a single manufacturer. Many receivers place the ultimate selectivity (that filter which passes the information bandwidth, such as a 2.4 kHz ssb filter) in the second i-f stage, or further down the active component chain than is desirable.

Frequently a receiver manufacturer does not offer filters which are optimized for RTTY or CW. If you find a filter with the desired response characteristics, chances are that it won't match the receiver i-f.

If you look to filter manufacturers who specialize in only crystal filters, you'll find that, within the past year or two, excellent crystal filters have become available that will optimize filtering for any mode of radio communications. The first stumbling block is the wide variety of filter center frequencies, typically

*Recent developments in solid-state mixer design actually permit the omission of all active stages prior to the mixer. It is now possible to obtain a mixer noise figure of 10 dB or less on the highfrequency bands. This will be sufficient for all but the most demanding reception requirements, such as OSCAR 7, Mode A on 10 meters. Editor

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fig. 1. Block diagram of the up/down frequency converter for adding an additional filter to a receiver. Z1 and Z2 represent the impedance matching networks necessary to interface the mixers and filter.

in the range from 5 to 11 MHz. Rarely does one manufacturer produce a complete line of filters, optimized for the information bandwidth of each of the operating modes used by amateurs.

Cost has been a major factor in the past, but crystal filter production techniques have vastly improved and costs have turned downward. To put cost into perspective, and to consider the effects of inflation, excellent 8-pole crystal filters, with signal rejection *floors* below 100 dB, are now available for about the same price level as 4- and 6-pole crystal filters were about 5-10 years ago.

Describing a crystal filter by its shape factor, normally defined as the ratio of the 6-dB to 60-dB bandwidths, doesn't tell the whole story. This measure of squareness is typically 1.7 to 2.2 for a good quality ssb filter, the slope of the response curve in a simple filter is determined by the characteristics of the crystals. For a 2.4 kHz ssb filter with a shape factor of 1.75, for example, the attenuation/ Δ frequency of the filter slope would be 54 dB/900 Hz. When narrower filters were designed, the bandwidth between corners was reduced but the slope remained essentially the same; a 500 Hz filter had a shape factor greater than 4.0. In fact, the bandwidth of the slope itself on one side of the ssb filter was wider than the 60 dB bandwidth of an optimized CW filter!

Eventually new fabrication techniques, such as mounting all crystals on the same header, permitted development of high-performance crystal filters. In some filters the entire passband may move as much as 2 Hz/°F, but this is not objectionable when the entire filter shape moves. To further illustrate the tremendous achievements in crystal filter technology that have occurred during the past several years, consider the 125-Hz CW filter now on the market. At one time, not too many years ago, 125-Hz crystal filters were a novelty of the laboratory. Today CW filters with 125-Hz bandwidths and shape factors of 2.5 are available for approximately \$125.



fig. 2. Block diagram of the Drake R-4B showing the receiver frequency conversion scheme. The conversion is necessary for the additional filter inserted between the first and second mixer. This diagram can be used to check for any sideband inversion.

By using the newer 8-pole crystal filters, receiver performance can be improved in the following ways:

1. Filters are optimized for the information band-width.

2. Signal-to-noise ratios are improved.

3. Filters placed as close to the front end as possible improve dynamic range.



fig. 3. A representative schematic diagram of the Texas Instruments TL442 doubly balanced mixer IC. Input 1 has an input impedance of 600 ohms, while input 2 is 50 ohms. The output can be connected for 600 or 1200 ohms output impedance, depending whether the supply voltage is connected to $+V_{\rm cc}$ or OUTPUT E.

4. Cascading with an existing filter to achieve better filter skirt and out-of-band performance.

When you optimize filters, you not only reduce susceptibility to interference — you also reduce listener fatigue. Improved signal-to-noise ratios are particularly noticeable on the low bands for several reasons. First, if only an ssb filter is available for CW, when a 125-Hz bandwidth filter is switched into the circuit, the bandwidth improvement ratio will be 10 log 2400/125 or 12.8 dB. Even if a 500-Hz CW filter were available, a 125-Hz filter would improve the signal-to-noise ratio by 6 dB. In addition, by the very nature of impulse and static noise, the filter will prevent overloading the following stages; this means that the signal-handling ability of the receiver has been improved.

Some receivers have a nominal 2- or 4-pole filter in

the first i-f, and the final selectivity in the second or last i-f stage. Obtaining all the needed selectivity at one i-f is difficult, however, because of signal radiation and leakage (even with the best shielding). By placing a high-performance crystal filter in the first i-f, overload of the second mixer can be greatly reduced. Further, spreading the selectivity over several stages is an excellent way to improve ultimate signal rejection.

Since the crystal filter you want to use will probably not agree with your receiver i-f, to say nothing of the input and output impedances, a convenient method is required to interface additional filters. Many receivers have a simple general purpose filter in the first i-f with an output impedance of 500 to 1000 ohms. This is an ideal place to add an outboard crystal filter.

Designing a conversion scheme can be a complicated process; consider all the variables. Fig. 1 shows a block diagram of the general approach for heterodyning the first i-f signal up or down to a highperformance crystal filter, and then heterodyning back to the receiver. The first mixer (up converter) can easily overload; the down-mixer is not nearly as susceptible to overload. In addition, the local oscillator can act as a source of spurious radiation for birdies in the amateur bands. Since only one local oscillator is generally used for the two mixers, it may serve as the leakage path for the signal around the filter. Or, the local oscillator signal could feed through the down-mixer into the next i-f stage. And consider the fact that since almost all commercial crystal filters are in the 5 to 11 MHz frequency range, the receiver i-f and the crystal filter frequency should be very close.

Having both the receiver i-f, fif, and the filter, fEL, at nearly the same frequency is probably the leading cause for abandoning the project. Mixers not only mix; they can amplify. Therefore, if f_{if} cannot be filtered out by the LC bandpass filter (BPF) at the output of the up-mixer, overload may eventually become a problem because both signals could be substantially amplified. To make matters worse, the conversion process usually results in some loss of desired signal; as much as 30 or 40 dB difference between the two mixer output signals, f_{if} and f_{FL}, is not uncommon. As a result, the filter signal rejection floor is greatly diminished. If the two frequencies are within several hundred kilohertz, the isolation of f_{FL} by the bandpass filter may be as big a task as manufacturing the high-performance crystal filter in the first place.

Finally, the filter must be very carefully matched to the up-mixer output and the down-mixer input. These matching networks are shown in **fig. 1** as Z1 and Z2.

Selecting a frequency conversion scheme should be done with care. **Fig. 2** shows the first i-f signal, filter i-f, and local oscillator frequencies. It is well to consider the particular sideband, too, since sideband reversal may not be desirable. **Fig. 2A** shows the complete receiver conversion scheme, with the Drake R-4B used as an example. The sideband slope diagrams show the relative sideband with respect to the incoming rf signal.

Fig. 2B shows how the up/down filter converter is integrated into the receiver's first i-f. Using the filter's center frequency, the required local oscillator frequency can be determined from $f_{LO} = f_{FL} \pm f_{if}$. The Sherwood Engineering CF-125/8* CW filter center frequency is 5695.0 kHz. Assuming an 800-Hz tone, the incoming frequency, f_{if} , would be 5645.0+0.8=5645.8 kHz; the local oscillator frequency would be 5695.0+5645.8=11340.8 kHz (or 5695.0-5645.8=49.2 kHz). The 11340.8 kHz frequency is, fortunately, not in any amateur band; but the 49.2-kHz local oscillator signal would fall within the passband of the receiver's second i-f! Therefore, 11340.8 kHz will be used as the local-oscillator frequency.

solving the problems

The close proximity of the receiver's first i-f and the crystal filter frequencies was the toughest prob-

*Sherwood Engineering, 1268 South Ogden Street, Denver, Colorado 80210.



fig. 4. The capacitive tap-down network is used to match the up-mixer to the crystal filter. It can also be used to match a high-impedance first i-f stage to the 600-ohm input of the TL442 mixer IC.

lem to solve. Convenience and easy-to-do were words which guided this design project for more than six months. Building the LC filter at the output of the up-mixer, shown in **fig. 1**, however, was anything but easy. Combining it with the impedance matching network, Z1, was complicated and certainly not repeatable without diligent tuning. Doubly balanced mixers were considered, but many of them required large numbers of external components and even null adjustments.

Finally, a doubly balanced mixer IC was found that provides internal preset nulls in excess of 30 dB, for



fig. 5. Schematic diagram of the up/down converter. All resistors are ¼ watt, 10 per cent. The TL442 pin-out is shown for the dual in-line package. The active devices can be obtained from Texas Instruments Supply, 6000 Dentron Drive, Dallas, Texas 75235.

both the input and the local oscillator signal. **Fig. 3** shows a diagram of the Texas Instruments TL442 (old designation SN76514). This circuit was designed specifically for radio receiver applications. Its features include

1. Flat frequency response to 100 MHz; with tuning usable to 300 MHz, C_i = 3-5 pF; C_o = 10 pF

2. 50 and 600 ohms input impedances and 600/1200-output impedance

3. Factory-tuned null adjustments for both signal and local oscillator

4. Single- or double-ended voltage source

5. Differential amplifier with large signal-handling capability

6. Low-level local oscillator requirement

7. Noise figure of approximately 6 dB

8. Typical conversion gain of 14 dB

In the TL442 IC, uhf transistor chips are matched and the resistors are etch-trimmed in the manufacturing process to achieve balance. The IC actually consists of two cross-coupled differential amplifiers whose emitters are driven by a third differential amplifier. A constant-current source is connected to the third differential amplifier emitter. This device works best with 250 mV local-oscillator injection, and performs without significant overloading, up to about 300 mV of rf signal. Hence, the signal-handling characteristics of the TL442 are as good as or better than most vacuum-tube converters in current receiver designs.

An excellent description of the TL442 is also available from Texas Instruments.¹ Cost of the doubly balanced mixer is \$2.40, an excellent trade-off when you consider that no external components are required. With more than 30 dB separation between the desired f_{FL} signal and the nearby f_{if} signal, the re-

mainder of the high-performance crystal filter converter design is downhill.

Impedance matching, or the lack of it, is a big benefit of using the TL442. The fixed 600/1200-ohm output required no LC network, and only the most simple matching circuit to match the 50-ohm crystal filter. Going from the filter to the down-mixer does not require matching when using the 50-ohm input of the mixer! With isolation between the local oscillator and the output port of more than 30 dB, the local oscillator signal will have only minimal impact upon the receiver, and will provide more than 60-dB protection against signal leakage across the filter.

The gain/loss in the conversion process is also worth planning. The Sherwood CF-125/8 filter has a typical loss of 9 dB (maximum 11 dB). Another factor is the bandwidth reduction loss from 2.4 kHz to 125 Hz, which was shown to be about 13 dB. I like background noise to remain constant rather than to keep the signal strength constant when switching between the two filters; the noise floor is always a ready reference and a 13 dB drop in the noise floor is a noticeable deadening of the receiver! If the TL442 is connected for a 1200-ohm output impedance, about three S-units of excess gain can be provided to slightly more than account for loss of background noise due to bandwidth reduction.

Designing the matching network, from the TL442's 1200-ohm output impedance to the CF-125/8 crystal filter's 50-ohm input impedance, is based upon the capacitive tap-down network shown in **fig. 4**. The IC output impedance is 1200 ohms in parallel with 10 pF of source capacitance. L1 is used to resonate this 10 pF and the series connected tap-down capacitors, C1 and C2. **Fig. 4** shows the relationships between the network components and the termination parameters. As discussed before, one of the advantages of the TL442 is that the crystal filter will directly match the 50-ohm input of the second TL442 mixer.



fig. 6. The 1:1 line isolation amplifier is shown in A, while the amplifier with the variable ratio is shown in B. This design is capable of handling large signals with low cross-modulation.

All that's needed to complete the converter are three semiconductor devices and one tuned circuit. **Fig. 5** shows the schematic. Both mixer ICs are configured in the same manner. Rf ground potentials are carefully bypassed with monolithic capacitors using short leads. The TL442 outputs are single ended, and the impedance is raised to about 1200 ohms by applying the supply voltage to pin 13. The constantcurrent source resistor network derives its voltage from the 3k resistor connected from pin 12 to pin 4; an additional 4.7k resistor is connected to pin 4 to increase gain. At the signal input, the 600-ohm input



fig. 7. Frequency response of the Sherwood Engineering CF-125/8 crystal filter.

(pin 11) is used for the receiver first i-f signal because matching to 600 ohms is convenient.

The local oscillator was carefully designed to provide as much decoupling from the supply voltage as possible and also to provide a very low output impedance. A T-pad attenuator between the two mixers further decreases the possibility of signal leakage from the signal input through the input mixer, oscillator, and through the output mixer.

The capacitive tap-down network from the upmixer, U1, to the crystal filter, FL1, is composed of L1, C6, and C5. Depending upon the Q of L1, and any other filter impedance, C5 and C6 can be adjusted to give the proper ratio for a good filter match. This occurs at or near maximum signal strength without objectionable ripple in the passband or outof-band ripples.

Retune L1 each time C5 or C6 are changed; the adjustment is straightforward and noncritical. C15 provides a fine-tuning adjustment for the crystal.

line amplifiers

When the output impedance of the receiver's first i-f is greater than one or two thousand ohms, a matching circuit will be required. Again, the capacitance tap-down network will work well for ratios of 24:1 or more. Further, there is sufficient gain in the TL442 IC mixers to recover a few dB of circuit loss. When excessive loss is encountered, a line amplifier (**fig. 6**) will help recover gain, or match extremely high-impedance circuits to low-impedance circuits. This circuit features several S-units of gain while exhibiting very large signal-handling capabilities with low cross-modulation distortion. The output impedance is 1200 ohms, untuned, and should be easy to match to the second-mixer circuit in any receiver.

construction

The whole system was built on a double-sided printed circuit board that fits over the pins on the CF-125/8 crystal filter. The filter is securely grounded to the back plane of the printed-circuit board to reduce signal leakage around the filter. A piece of double-stick *Scotch* mounting tape was used to attach the entire up/down crystal filter converter assembly to an unused panel inside the receiver. One word of caution: always place a metal shield between the input wafer and the output wafer of the crystal filter switch to minimize signal leakage.

results

True single-signal reception with the CF-125/8 crystal filter in tandem with an ssb filter is most gratifying. Fig. 7 shows the frequency response of the CF-125/8 crystal filter by itself. Other crystal filters give equally impressive results. My R4-B receiver is equipped with a 1:1 line amplifier which drives the Drake ssb filters from the 2-crystal filter in the first i-f stage. The output from the ssb filter drives the line driver (adjusted to make up the 6 dB filter loss) and then the second mixer. The up/down crystal filter converter is switched in between the ssb filter and the second line amplifier. The tandem combination of filters does not ring, and 40 word-per-minute CW copy is possible. The noise and static effects that were so bothersome when a 125-Hz audio filter (shape factor 3) was used are now annoyances of the past.

If you are lucky enough to find crystal filters that are on the same frequency as your receiver's first i-f, all that is needed is a line isolation amplifier to buffer the filter, a matching network, and a line amplifier to make up the gain of the return signal.

reference

1. Balanced Mixer Application Note, Section 6.6 SN76514/TL442, Linear Circuits Application Department, Mail Station 964, Dallas, Texas 75222.

ham radio

how to select TTL sub-series ICs

for different digital designs

The popular TTL family of digital ICs is widely used in amateur applications, but low-power, high-speed, and Schottky TTL have been largely neglected here's how to select the best TTL sub-series for your own designs

Through the years, as the 7400 series of ICs has become the mainstay of TTL logic designers, more and more devices have been added to the family. In the last few years, both Fairchild Semiconductor and Texas Instruments have increased their commitment to the market by introducing expanded lines of high speed, low power, and Schottkyclamped devices. At the same time, extensive use of foreign production facilities has allowed a 50 per cent drop in prices, which distributors are now beginning to pass along to the consumer. Where does that leave you when you decide to build that new keyer or frequency counter? Consider the popular 7400 quadruple 2-input NAND gate, for example. There are five versions: the 7400, 74H00, 74L00, 74LS00, and 74S00. Which version is best suited for your purposes? What advantages does one version have over another?

Actually, each 7400 sub-series (H, L, LS, and S) has clear cut strengths and weaknesses which make the choice a lot easier than it may appear. The two major differences between each sub-series are speed (maximum operating frequency) and power consumption. In general, to gain speed, power consumption must be increased. This speed-power trade-off would probably settle the matter because you would pick the lowest power version that meets the required speed and stop right there, but a third factor comes into play: cost. To either increase speed or decrease power, the cost at least doubles over that of the standard (and the least expensive) version. A good rule of thumb is to use these special devices only when the standard 7400-series chips can't do the job.

performance comparison

Let's go back to the 7400 quad 2-input NAND gate and look at the differences between each version and set down some general characteristics for each sub-series.

The 7400 typically operates from dc to 35 MHz, as will the remainder of the 7400 series. This is a typical specification and does not hold true in devices of higher complexity such as the Texas Instruments SN74144, which contains a BCD counter, a four-bit latch, and a BCD to seven-segment decoder-driver. The SN74144 is intended to be a one-chip replacement for the popular SN7490A counter, SN7475 latch, and SN7447 decoder-driver combination. Because of the high component density in the SN74144 (an equivalent of 86 gates on one chip), however, the typical counting frequency only extends to 18 MHz. Some of the earlier devices were equally slow. There was a time, not too long ago, when it was sometimes necessary to go through a handful of 7490 counters before you could find one that would work to 30 MHz. Therefore, it's wise to buy only devices with current date codes, unless the application isn't critical. Problems shouldn't pop up with any major suppliers, like those who advertise in ham radio, because the turnover is too high for 1973 chips to be still floating around in the open market.

Digressing a moment, the date code is a three- or four-digit number standardized by the EIA (Electronic

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Industries Association). It is stamped on every integrated circuit, usually, but not always, after inspection. Contrary to what you might think, you can get an untested IC with all the same markings as a first-rate unit. If the number has four digits, the first two represent the year of manufacture, such as 75, and the next two stand for the calendar week, such as 38, which would mean the thirty-eighth week of 1975 (the third week in September). If the number has three digits, the first is the year of manufacture (in the example above, the year would be cropped to 5), and the last two digits refer to the calendar week.

The typical low power 74L00 version will operate to 3 MHz, making it somewhat slower than the CMOS family. Every other sub-series is faster than the original type. Below is a list in terms of *typical* speed:

74L	3 MHz
74LS	45 MHz
74H	50 MHz
74S	125 MHz

These figures represent the highest typical clock rate for flip-flops. Once again, remember that each device must be considered on a one-by-one basis where speed is concerned, with the higher-density units having lower maximum frequencies than their less complicated brothers.

Power consumption is often compared using the power dissipation per gate for each series. This information is given in **table 1**, along with all other comparative figures, but in this case it is based on the average supply current, per gate, assuming a 50

the SN7490A, SN74L90 combination, and 13.15 for the SN7473, SN74L73 flip-flop pair. The average supply current values for the remainder of the devices are: 4.5 mA for the 74H00, 0.4 mA for the 74LS00, and 3.75 mA for the 74S00. Thus, if the 7400 is used to establish the standard unit of power consumption (2 mA = 1 unit), then the relative standings are 0.1, 0.2, 1.9, and 2.3 for the L, LS, S, and H sub-series, respectively.

selecting a sub-series

There are three variables that must be considered when choosing the proper series: price, speed, and power. In general, the first and deciding requirement is that the chip will work up to the desired frequency. It is possible to approach the choice from a power consumption standpoint, but if power conservation *is critical*, it would be a good idea to see what can be done with a very low power series like the RCA CD4000 COS/MOS family. On a cost-effective basis the low-power 74L00 series is not as good as the COS/MOS family, which has a much lower powercost product; COS/MOS will also work at higher frequencies (5 MHz for counters, 10 MHz for gates and flip-flops).

This brings us to the method for selecting the best sub-series once the speed requirements have been fulfilled: the *power-cost product*. Since both cost and power are to be minimized, it is easier to multiply the two figures together and deal with one variable instead of two. The lower the power-cost product, the more performance you get for your

table 1. Comparison of the various TTL sub-series showing clock, rate, power dissipation, propagation delay, relative cost, and power-cost product

series	maximum flip-flop clock rate	power dissipation per gate	gate propagatic delay time	cost increase on over standard e series	power-cost product
74L00	3 MHz	1 mW	33 ns	3.1	0.62
7400	35 MHz	10 mW	10 ns	1.0	2.00
74LS00	45 MHz	2 mW	9.5 ns	1.4	0.57
74H00	50 MHz	22 mW	6 ns	1.9	8.49
74S00	125 MHz	19 mW	3 ns	3.4	13.07

per cent duty cycle. The average supply current data is readily available for individual devices, whereas power dissipation is generalized for all gates in the series. The difference in consumption will hold true, when comparing more complicated devices, so long as it is treated as an approximate ratio. In other words, if the average supply current for one gate of a 7400 is 2 mA, and the average supply current, per gate, of a 74L00 is 0.2 mA, then it is fair to say that *any* standard 7400 series device will require approximately ten times the amount of power than an Lseries unit does. In practice, the actual ratio may be more or less.

To take several cases, the power ratio is 7.25 for

money. In practice, the product is calculated by multiplying the average current per gate (in mA) for the series, by the average increase in price of the series over that of the standard 7400 series (given as a multiple, such as 3.1 times cost). Data is provided in **table 1**. This is a method for standardizing the selection process, or a mathematical replacement for common sense.

As an example of how to use the chart, suppose you are planning to built a 10-MHz frequency standard. The 10-MHz specification puts everything in the running except the L series. If cost effectiveness is the object, a look at the lowest power-cost product reveals that the LS series is your best bet. Sheer low cost, providing that a husky power supply is available, would be provided by a switch to the standard series. Not all device selections are that simple.

Let's assume you are designing a frequency counter. The goal is to build a model capable of counting to the highest frequency and requiring the least possible power, and using only the TTL series (no CMOS or ECL integrated circuits); price is no object. In case the design goal will not be met by using only one 7400 sub-series, the lowest power version having the necessary speed will be selected. Excess speed margins, when not needed, will be sacrificed for power conservation. Starting with the 10 MHz oscillator, choose a 7400. The 74LS00 and other subseries chips have a reputation for not performing well in oscillator service. A key to this problem is the different biasing requirements for each sub-series. It is impossible to just simply remove a 7400 from an oscillator circuit and plug in a 74LS00 without changing external resistor values. There are many proven oscillator circuits based on the 7400, but little published information about biasing for oscillator service, so sticking to the well trodden path will assure success.

build a higher current power supply than to purchase twenty-five special ICs at three times the cost of their standard TTL equivalents.

TTL sub-series compatibility

One of the original design objectives for the different TTL sub-series was compatibility. All have the same maximum supply voltage rating of 7 volts, except for the L series, which is 8 volts. This gives plenty of leeway above the typical 5.0 V supply voltage, which is common to all sub-series. Operating temperature range extends from 0 to 70°C (32 to 158°F). The maximum input voltage for the L series is 7 volts, with 5.5 volts as the limit for all others. Because of these similarities, mixing devices from different sub-series will produce no problems so long as fan-out limits are observed.

Fan-out (the number of inputs a single output can drive) is figured only on the basis of outputs driving inputs from the same sub-series. The standard fan-out is 10 loads, except for the L and LS series, where it is 20. Mixing of devices is permitted as long as the output can source (provide) or sink (absorb) the *total* current to or from all inputs.

The high- and low-state input requirements are shown in **table 2** along with output sink capabilities.

The first divider must be able to toggle up to 10

table 2. Input and output data for the various TTL sub-series. Note that the L series had two different standard inputs; assume highest input current when calculating output requirements. Negative signs represent current flow out of terminal

series	input current (high state)	input current (low state)	maximum output sink current	maximum output source current
74L00	10/20 μA	-0.18/0.8 mA	3.6 mA	200 μA
7400	40 µA	-1.6 mA	16 mA	-400 μA
74LS00	20 µA	-0.4 mA	8 mA	40 0 μA
74H00	50 µA	-2.0 mA	20 mA	500 μA
74S00	50 µA	–2.0 mA	20 mA	−1000 μA

MHz; a 74LS90 will require the lowest power. All remaining dividers operate at 1 MHz or below, making the 74L90 the best bet. The counter control circuitry, which generates the count enable, strobe, and reset pulses functions at a very low rate, since most counters can make no more than 10,000 counts per second, so L-series devices can be used. The gate must pass the highest counted frequency, as must the first decade counter, and this application calls for Schottky ICs such as a 74S00 for the gate, and a 74S196 for the first counter. The second decade counter must be LS to count up to 12.5 MHz, but the remaining counters may be L versions. Latches and decoder drivers can also be chosen from the L series.

It is important to note that no standard series TTL logic was used in this circuit. Only when performance can be sacrificed in favor of price is standard TTL a wise choice. Price is almost always important, which explains my rule of thumb which suggests that the standard series be used exclusively, except when it just won't do the job. It's less expensive to From these figures it's easy to check to see whether a particular output can handle its loads. Just add up the low-state currents for all loads (inputs), and then check that the totals fall within the maximum sink limit for the output. The negative values of input current represent a flow out of the terminal, back into the output. If the output can sink the required current, it will always be able to source enough current for the loads.

conclusion

While the use of standard and LS series TTL ICs has certainly caught on for amateur projects, the H, L, and S series have been largely neglected. As so often happens with new products, this is due more to insufficient information than it is to a lack of applications. It is hoped that this article has provided enough information to generate more interest in using the various TTL sub-series ICs in future designs.

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- 5. Numerous small lizards and spiders were crawling in

Despite all these adversities, the two Drake C-Lines were

Despite all these adversities, the two Drake U-Lines were in operation for over 145 hours; the only "off time" was in operation for over 140 nours; the only "Olf time" was during the six hour trip from Palmyra Island to Kingman Reef. during the six hour trip from Paimyra island to Kingman Meer A total of over 16,000 contacts were made from KP6AL Palmyra A total of over 10,000 contacts were made from KPOAL Pa and KP6BD Kingman Reef without a single transmitter or I know that customer satisfaction has been the cornerstone

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the dragon one supply

This article describes a power supply system that delivers a whopping 500 watts of clean, regulated power. The regulation is better than 1 per cent from no load to full load, with a ripple voltage of less than 10 mV peak-to-peak. Safety features such as current limiting, overvoltage shutdown, and short-circuit protection are all built in. Optimum regulator efficiency is approximately 65 per cent at an output of 450 watts.

Fig. 1 shows the regulated-power supply schematic. Transformer T1 steps down the line voltage to 22 volts, which is rectified by full-wave bridge rectifier CR1. Filtering is by C1, which is a computer-grade electrolytic capacitor having a capacitance of 18,000 μ F. R1 discharges C1 after the power supply is turned off.

Regulation is provided by U1, the popular 723 regulator IC. The 12-15-volt voltage adjustment is by R4. C5 ensures oscillation-free operation of regulator

By C. C. Lo, WA6PEC, 5414 Barrett Avenue, El Cerrito, California 94530





table 1. Parts list for the 500-watt regulated supply.

6 resistor 15 milliohm Lotronics R15M Dragon One assembled and tested - \$209.50	iat sinks 10° x 6° (255x152mm) Lotronics H10-6, one piece Items 1-9 Dragon One major component kit – \$129.50 4° x 6° (102x152mm) Lotronics H4-6, two pieces Dragon One complete component kit – \$169.50	relay 24V coii dpdt, contacts rated at 10A 125V each 723 regulator DIP package R1, SCR2 MCR 103 or equivalent (50V 200μA gate current) fuse holder and 8A fuse V1 toggle switch dpst on-off scellaneous wire, screws, washers, terminal block, line cord. e following parts are available from Lotronics, Box 975, El Cerrito, California 94530: ms 1-9 Dragon One major component kit – \$129.50 Dragon One assembled and tested – \$205.50 Dragon One assembled and tested – \$205.50	16V zener 1W, 1N4745 or equivalent diode 50V 1A red light-emitting diode transformer Lotronics T2230 bridge rectifier Varo VK 148 or equivalent electrolytic capacitor - 18,000 µF, 35V uit board Lotronics PC723SO 10 [~] x 6 ⁻ (254x152mm) Lotronics H10-6, one piece 4 [~] x 6 ⁻ (102x152mm) Lotronics H10-6, two pieces resistor 15 milliohm Lotronics R15M	42 13 13 11 11 11 11 11 11 11 11 11 11 11
	16 resistor 15 milliohm Lotronics R15M Dragon One assembled and tested — \$209.50	r all above add \$10 for shipping, insurance, and handling. California residents add 6 per cent tax Instructions includ	7	hassis
Heat sinks 10° × 5° (254×152mm) Lotronics H10-6, one piece Items 1-9 Dragon One major component kit – \$129.50 4″ × 6″ (102x152mm) Lotronics H4-6, two pieces Dragon One complete component kit – \$129.50		e following parts are available from Lotronics, Box 975, El Cerrito, California 94530:	uit board Lotronics PC723SO	Printed circ
Printed circuit board Lotronics PC723SO Heat sinks 10" x 6" (254x152mm) Lotronics H10-6, one piece Items 1-9 Dragon One major component kit – \$129.50 4" x 6" (102x152mm) Lotronics H4-6, two pieces Dracon One complexes component kit – \$129.50	The following parts are available from Lotronics, Box 975, El Cerrito, California 94530:	scellaneous wire, screws, washers, terminal block, line cord.	electrolytic capacitor - 18,000 μ F, 35V	5
C1 electrolytic capacitor - 18,000 μF, 35V Miscellaneous wire, screws, washers, terminal block, line cord. Printed circuit board Lotronics PC723SO Heat sinks 10 x 6 (102x152mm) Lotronics H10-6, one piece Items 1-9 Dragon One major component kit – \$129.50 4 x 5 (102x152mm) Lotronics H4-6, two pieces Items 1-9 Dragon One major component kit – \$129.50	C1 electrolytic capacitor - 18,000 μF, 35V Miscellaneous wire, screws, washers, terminal block, line cord. Printed circuit board Lotronics PC723SO The following parts are available from Lotronics, Box 975, El Cerrito, California 94530:	V1 toggle switch dpst on-off	bridge rectifier Varo VK148 or equivalent	CR1
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723. U1 output drives Q2, an MJE3055, which in turn, drives Q3-Q6, 2N3055s. Current sensing is by R16, a special 15-milliohm resistor. The two output terminals are isolated from chassis ground. Grounding is achieved by connecting the positive or negative output terminal to the ground terminal with a jumper. A light-emitting diode indicates the presence of dc output voltage. R3, R4, and R5 make up the output voltage sensing divider; the voltage control signal is connected to U1 inverting input.

To protect the power supply from burning itself up in case of excessive load current, the short-circuit shutoff is done in conjunction with the current limiting provided by the regulator through R16. As load current exceeds 35 amps, the output voltage starts to drop. When the voltage drops below 8 volts, Q1 turns off and SCR1 turns on, pulling the regulator noninverting input close to ground potential, thus turning off the output power. This condition remains until the power supply is turned off and SCR1 unlatches.

Overvoltage shutdown is designed into the system to protect your expensive transceivers and linear amplifiers. If anything should happen to the regulator or any of the series transistors, chances are one of these devices will short out, putting the full voltage across C1 at the output. This could be disastrous to transceivers and amplifiers. Relay K1, together with CR2 and SCR2 ensure that this will not happen, even if all the pass transistors and regulator are shorted. As the voltage exceeds 16 volts, CR2 starts to conduct, supplying gate current to SCR2, which turns on and activates K1. In doing so, the main dc supply is cut off and will remain off for as long as the power is on and the defect has not been corrected. This special feature is valuable and its additional cost is well justified, although the overvoltage shutdown feature may never be needed in the lifetime of the power supply.

construction

All components are packaged in a 7 x 8 x 10-inch (178x203x254mm) steel chassis box. Three heatsinks are used (photo).

All components shown inside the dotted line in the schematic diagram are mounted on the printed circuit board. Since this circuit is a high-current power source, no. 12 (2.1mm) wire should be used for all high-current paths. However, no. 16 (1.3mm) wire can be used for interconnections from the two relay contacts, which are wired in parallel to the individual transistor collector and from the individual emitter to point L or R16. R19 and C6 are mounted behind the output terminal block. Holes are punched on the top and bottom panels for ventilation purpose. Output voltage can be adjusted between 12-15 volts dc. Load current is rated at 35 amps intermittent, and 22 amps continuous duty. For prolonged operation at high current and low output voltage (below 13 volts), a small external fan is recommended for cooling the heat sinks. However, the power supply can deliver 22 amps continuusly without forced-air cooling if ambient temperature is below 77°F (25°C). The temperature of the pass bank tran-



Underchassis view of the power supply. Three heatsinks are used. The heatsink mounted on the rear of the chassis box is isolated from chassis ground. The four 2N3055s ($\Omega_3 - \Omega_6$) are mounted directly on the heatsink. The heatsink on the right-hand side is for bridge rectifier CR1; the other heatsink is for Ω_2 . Heatsink compound was used for mounting $\Omega_2 - \Omega_6$.

sistor under this condition stabilizes at around 221°F (105°C). With a 25-30 cfm (7 x 10^5-8 x 10^5 cm³/minute) fan blowing at the rectifier and the transistor heatsinks, the transistor heatsink temperature stabilizes at 122°F (50°C) with 30 amps continuous load current operation for one hour. Regulation is below 1 per cent from no load to full load (35 amps). The taps on transformer T1 are for optimum efficiency operation. It's obvious that if the line voltage is high, the unregulated dc voltage will also be high, making the voltage drop across the pass bank transistor high. That means higher power dissipation and lower system efficiency. Hence, if the input line voltage is connected to the proper tap, an optimum system efficiency is achieved.

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voice-operated gate

to replace voice-operated relays for carbon microphones

Presenting a circuit using four ICs plus a couple of transistors and diodes to replace the old voice-operated relays in ssb transceivers **VOR is an acronym** for what is often called the "voice-operated-relay" or "squawk-to-talk" circuit, as used in many modern ssb and fm transceivers. "Voice-operated-relay" was an adequate description when tubes and relay circuitry were used, but it's rather unusual to find such relays in today's all-solid-state designs. And so now we have the Voice Operated Gate, or VOG.

The VOG described here used four ICs plus a couple of transistors and diodes to accomplish preamplification, bandpass filtering, and audio gating. A logic output also comes out of the VOG (choice of 1 or 0 for *gate-on*), to serve as a turn-on signal for other sections of the system being voice controlled.

Incorporated in the VOG is a lowpass and highpass filter pair providing the equivalent of a 300-3000 Hz bandpass filter with 40 dB per decade rolloff at each edge. These filters are of the active type, built around operational amplifiers. Only the audio passing through the filters can actuate the gate (and thereby pass through the VOG); this helps to discriminate against ambient noise.

VOG circuit

A diagram of the VOG is shown in **fig. 1**. The first section is a microphone preamp with an fet constantcurrent source for a carbon microphone. The carbon microphone is a variable resistance, so the injection of a constant-current into it causes the voltage across it to be representative of the variations in resistance of the microphone. The op amp that forms

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fig. 1. Voice-operated-gate (VOG) circuit block diagram. The circuit is a replacement for the old voice-operated relay systems prevalent in many modern ssb and fm transceivers.

the microphone preamp has a voltage gain of 100, which provides a voltage output of about 3 volts rms for usual carbon microphones during average closetalking use.

Following the microphone preamp is the highpass active filter followed by the lowpass active filter, each consisting of one section of the same quad op amp (U1) that's used as the preamp (see **fig. 2**). The last section of U1 is used as an *active* diode detector CR1, CR2, in which the op amp *linearizes* the detector. The diode detector is arranged to furnish the negative polarity of rectified audio.

The rectified audio from CR1, CR2 is then averaged by U2. Since the averager (U2) is also an inverter, the negative rectified audio is inverted and averaged to become a smoothed, long, positive pulse of the duration of the audio burst originally delivered by the microphone. This positive pulse is processed by U3, a Schmitt trigger, which sharpens the pulse leading and trailing edges and makes it



fig. 2. Schematic showing the carbon microphone preamp, bandpass filter, and voice-operated gate (VOG). Note that U1 is the equivalent of four µA741 op amps and could be replaced by four such ICs.

CMOS-logic compatible. The Schmitt trigger also inverts the pulse and adds an effect called hysteresis. That is, U3 output (pin 7) will go from 1 to 0 at an input level (set by the threshold-adjust pot) of say, 2 volts. U3 output will not return from 0 to 1 until the input voltage has dropped substantially *below* 2 volts. This hysteresis action prevents noise on the audio and minor voice level wavering from causing a chopping effect.

After the Schmitt trigger comes Q2, a simple transistor inverter, which inverts the audio-derived pulse to provide the proper polarity to turn on analog gate U4 when an audio signal is present. The inverter output and the Schmitt trigger output provide both **0** to **1** and **1** to **0** logic-level outputs, which can be used to actuate the turn-on function of the transmitter. Both polarities are handy, because this unit may be used with a number of transmitter designs.

The analog gate, U4, is a member of the RCA CD4000 CMOS logic family, which makes it much less expensive than some of the hybrid analog gates on the market. U4 consists of four analog gates. Since we need only one, all four sections have been wired in parallel. The CD4016 doesn't tolerate very large ac voltages without distortion, so the (filtered) audio input is attenuated at a ratio of 3:1 by a voltage divider at the analog input.

adjustment and testing

Setup of the VOG is simple. Connect it to a carbon microphone and a \pm 15-volt supply. Connect a scope or ac VTVM to U1 pin 3 of U1. Talk into the microphone and adjust the LEVEL pot (**fig. 2**) until about 3 volts rms is seen, then adjust the THRESHOLD pot until about +2 volts is seen at its wiper arm. Connecting a scope or ac VTVM to the output should now show a pulse of audio when speaking into the microphone. A dc voltmeter at U3 pin 7 should jump from +15V for "no talking" to near zero for "talking". The same dc voltmeter at the O2 collector should react in the opposite way: near zero for "no talking" and +15 volts for "talking."

closing remarks

This VOG circuit was originally designed to replace one of the special-purpose ICs made by a large linear IC manufacturer. It surpasses the device it replaces in every way.

Note that U1 is the equivalent of four μ A741 op amps and could be replaced by four such ICs. Also, two μ A747s (dual μ A741 op amps) or two MC1458s could also be used. U2 is best left as an LM301A, since the requirement here is for low input bias currents. When using other-than-called-for ICs, however, pin changes will have to be made.

ham radio



accurate low power rf wattmeter

for high frequency and vhf measurements

How to build an accurate low-power wattmeter that measures up to 10 mW from 1 to 500 MHz it uses small lamps as barretters

A pair of subminiature lamps used as an rf power detector make up the heart of a simple but accurate rf power meter, which can be calibrated directly from dc measurements. The instrument described in this article can be used to accurately measure rf power from 10 mW down to about 0.2 μ W, over a frequency range from 1 MHz to 500 MHz. Its high sensitivity makes it useful for a host of purposes including antenna gain measurements, local oscillator measurements in conjunction with low-power signal generators. Its maximum power capability can be extended to any level through the use of external calibrated attenuators or directional couplers. In

addition, homebrew attenuators and directional couplers can themselves be calibrated using the power meter.

The rf power detecting element in the wattmeter consists of a pair of incandescent lamps used as barretters. Barretters have been used for many years in commercial wattmeters and have been discussed in several previous articles.^{1,2}

A barretter is a wire element whose resistance increases with temperature. Suppose a barretter is heated to a specific resistance (say 50 ohms) by a variable power source whose level is known. As long as the total power dissipated and the ambient temperature remain constant, the barretter resistance will remain at 50 ohms. Now suppose the barretter is also heated with power from a separate source (an rf generator in this case) whose level is unknown. The resistance of the barretter will increase. If the power supplied from the known source is then reduced until the barretter resistance returns to 50 ohms, the amount of power reduction from the known source will equal the power supplied by the unknown source. The unknown power level is thus measured by metering the decrease in the known power source.

The known power source can be adjusted automatically to maintain constant barretter resistance by using a bridge circuit in a closed loop with an amplifier. In many commercial microwave power meters the closed loop forms a self-balancing audio oscillator so that the known power source is an ac signal (in combination with some dc which is also applied). The oscillator technique has the advantage of eliminating dc offset drift errors in the balancing and metering circuits. In the power meter described here, however, the known power source is pure dc. The dc approach was chosen for ease of calibration and testing, for circuit simplicity, and to allow a wide rf frequency range. (The relatively large rf coupling capacitor required for low-frequency response would introduce excess phase shift and upset the balance

By James H. Bowen, WA4ZRP, 6500 Carefree Lane, Apartment B1-21, Roanoke, Virginia 24019 in an ac balanced bridge, depending on the lowfrequency impedance of the rf source.)

circuit description

The circuit diagram of the rf wattmeter is shown in fig. 1. The design philosophy was to explore what useful sensitivity could be achieved in a simple circuit without the use of special low-drift components or special schemes for drift compensation. The experimenter who wishes to build his own version of the wattmeter is encouraged to try his hand at improvements.

The incandescent lamps, I1 and I2, used in the rf sensor are subminiature T-3/4 types obtained at a hamfest flea market. The lamps have wire leads and the glass envelopes are 0.187 inch (5mm) long by 0.094 inch (2.5mm) in diameter. Their dc characteristics indicate they are similar to Chicago Miniature types CM2, CM30, or CM3102. Fig. 2 shows the measured current-voltage (I-V) characteristic of one of the lamps. Note the non-linear nature of the plotted data which indicates changing lamp resistance. This general characteristic is typical of all incandescent lamps with tungsten filaments. Fig. 3 shows the same data plotted as dc resistance, V/I, versus power dissipated, VI. At rf frequencies, the resistance of the lamp during any rf cycle remains constant and equal to the dc resistance because one rf cycle is much shorter than the minimum thermal response time of the lamp filament (skin effect does not appear to seriously alter the resistance of the small diameter, high resistivity filament over the frequency range of interest).



The author's completed power meter.

In order to simultaneously feed dc and rf to the lamps over a wide bandwidth, the lamps are connected in series for dc and in parallel for rf. Chip capacitors C1 and C2 perform the functions of rf coupling and bypassing, respectively, with low impedance over a wide frequency range. If chip capacitors are not available, small ceramic disks with zero lead length may be used. For good uhf measurement accuracy, construction of the rf sensor must be based on good uhf construction practices, with emphasis on minimizing parasitic inductances by keeping all leads short. The rf paths through C1, either lamp, and C2 to ground must be as short as possible.

The rf sensor is built on a small piece of doubleclad glass-epoxy printed-circuit board 1/16 inch (1.5mm) thick as shown in the photograph. Both sides of the board were soldered directly to the rear of a BNC connector, with the connector center pin soldered to a pad approximately 0.105 inch (2.5mm) wide. This pad forms a 50-ohm microstrip transmission line leading to chip capacitor C1. One lead of both 11 and 12 is soldered to a small pad connected to the other end of C1. A small hole is drilled through the board at the ground lead of I2 so this lead can be soldered to the ground plane on both sides of the board. The opposite lead of I1 is soldered to the pad in the upper right hand corner in the photo. A dc feed wire is also soldered to this pad and chip capacitor C2 is soldered from the point of attachment of I1 across a gap to the ground plane.

On the ground side of C2, another hole is drilled through the board and a wire is soldered through this hole to form a direct connection to the ground plane on the back of the board. The use of small filament lamps with low parasitic inductance, and this method of construction ensure good performance into the uhf portion of the spectrum. (Warning: chip capacitor ends must be soldered quickly with minimum heat; otherwise tin-lead solder will rapidly leach away the metallization from the ends of the capacitors.)

The layout of the remaining dc portion of the wattmeter circuit is not particularly critical and was built on a Vector DIP padboard mounted on the meter terminals. The unit is housed in a $4 \times 5 \times 6$ inch (10.2 x 12.7 x 15.2cm) minibox.

The lamps are operated at sufficient dc current to bring their series resistance to 200 ohms. If the lamps are reasonably well matched, the resistance of each lamp will be about 100 ohms, making the parallel rf resistance equal to 50 ohms. If two lamps identical to the one plotted in **fig. 3** are used, each will dissipate about 7 mW at a resistance of 100 ohms, for a total dissipated power of 14 mW. Thus, 14mW is the maximum rf power which can be measured in a 50-

ohm system with two such lamps. A highest scale of 10 mW was therefore chosen for the wattmeter. Random drift establishes a practical limit of 10 µW for the most sensitive scale.

To maintain their series resistance at 200 ohms, the lamps are operated in a bridge circuit consisting of R1, R2, R3, and the rf sensor. For best accuracy, R1, R2, and R3 should all be selected to be as close as possible to 200 ohms with R1 and R2 selected for best match, and R3 selected closest to 200 ohms.

The voltage difference between the two legs of the bridge is sensed and amplified by U1, a μ A741 or similar type op-amp IC. The capacitors in the feedback loop of U1 form an integrator for very high dc gain and good stability. The output of U1 passes through diode CR1 to transistor Q1, the bridge current driver. Q1 is connected as an emitter follower, and supplies the necessary current to bring the bridge to a balanced condition. The 10k resistor across Q1 feeds a small residual positive bias to the bridge to ensure that the bridge will always come to balance with a positive potential, even though U1 may initially turn on with a negative output. Diode CR1 prevents emitter-base breakdown of Q1 if U1 turns on with a negative output.

Following turn-on, the output of U1 will quickly

become positive in response to the residual positive bias on the bridge. The voltage at the output of U1 will continue to increase until enough current flows through the rf sensor to bring its resistance to 200 ohms, at which point equilibrium is achieved. In practice, the bridge comes to balance within a second or two of turn-on, with some overshoot due to the thermal lag of the lamps.

The equilibrium voltage at the top of the bridge, V_{B} , (3.50 volts in the unit shown) is fed to the metering circuit made up of U2A, U2B, and associated components. Range switch S2A selects one of the calibration resistors, R4 through R10. A method for calculating the values of these resistors is covered in the calibration section.

Op-amp U2A compares the voltage selected by S2A to a reference voltage established at pin 3 of its input. Since the full-scale voltage change in V_B is only 1.1 mV for the 10 µW scale, the reference voltage supply must be extremely stable and minutely variable. To establish a stable reference voltage, fet Q2 is connected as a constant-current source feeding zener diode, CR2. Any fet having an IDSS of 3 mA or more could be used for Q2. Alternatively, a 5-volt, three-terminal regulator IC could probably be used instead of O2 and CR2.



C1, C2 0.1 µF chip capacitor or miniature leadless ceramic discap

- 11, 12 subminiature T-3/4 incandescent lamp (Chicago Miniature type CM2, CM30, or CM3102)
- J1 BNC lack, flange mount

fig. 1. Schematic diagram of the rf wattmeter for 1 to 500 MHz. Fixed-value capacitors are disk ceramic except as noted; polarized capacitors are electrolytic or tantalum; resistors are 1/4 or 1/2 watt carbon composition types.

B15 miniature 50k 10-turn pot

2k trimmer R4-R10 (see table 1 of text)

R19

- S1 dpst toggle switch
- **S**2 2-pole, 7-position rotary wafer switch

Resistor network R11 through R16 divides the zener voltage down to the value required to match V_B. To get the required voltage resolution with a reasonable adjustment range, a miniature 10-turn pot was used at R15. If a 10-turn pot is not available, then both a coarse and a fine adjust pot must be used. Resistors R11 through R14 and R16 reduce the adjustment range of R15; this increases resolution. Resistors R11 through R13 are chosen to establish a reference voltage close to V_B with the wiper of R15 disconnected. Resistors R11 through R13 also serve to maintain a fairly low impedance for the reference voltage. Resistors R14 and R16 are selected to reduce the adjustment range of R15, and to establish a residual voltage close to V_B on the wiper of R15 when the wiper is set at mid-range.

Since the specified minimum open-loop gain of a single μ A741 op amp is marginally low for proper operation of the metering circuit, two op amps are con-



Interior of the rf power wattmeter. All active circuits are installed on the perf board mounted on the meter terminals. The two incandescent lamps are mounted on the small section of printed-circuit board soldered to the BNC jack (lower left).

nected in cascade. Op amp U2B supplies an additional gain of 100 to the open-loop gain of U2A. A dual op amp, the MC1458CP, was used for U2A and U2B, though two μ A741s could have been used or a quad 741 could have been used for the entire unit.

The meter, M1, is connected in the feedback path of U2. Meter M1 is a 200 μ A meter removed from an old vacuum-tube voltmeter. The action of U2 is to supply enough current through the feedback path to maintain the voltage at pin 2 of U2A equal to the



fig. 2. Current-voltage (I-V) characteristic of an incandescent lamp of the type used in the rf power meter.

reference voltage at pin 3. Since the current flowing in pin 2 of U2A is negligible, the current in the feedback circuit continues through the calibration resistor, R_{CAL}, selected by S2A. This current has no effect on V_B because it is automatically compensated for by U1. By Ohm's law, the feedback current is equal to $\Delta V/R_{CAL}$, where ΔV is the difference between the reference voltage and V_B. On all scales except the 10 mW scale, all feedback current normally passes through meter M1. Diode CR3 conducts when the feedback current is negative, preventing M1 from pinning hard in the negative direction when the circuit is negatively unbalanced. Resistor R17 prevents M1 from being severely overloaded in the positive direction when the circuit is unbalanced positively. Resistor R17 is selected so that M1 reaches full scale somewhat before the output of U2B saturates in the positive direction. Resistor R18 and diode CR4 shunt some feedback current past M1 on the high end of the 10 mW scale to linearize the reading.

To allow portable operation, the unit is powered by two 9-volt batteries. Battery voltage *sag* following turn-on contributes some additional drift to the circuit. The miniature transistor radio batteries shown in the photograph sagged excessively and have been replaced by larger 9-volt batteries (Eveready 246). For enhanced stability, somewhat higher battery voltage could be used followed by electronic regulators to 9 or 12 volts. If it is desired to power the unit from the ac line, regulated dc supplies are a must.

The value of calibration resistance, R_{CAL}, for any scale is determined by calculating ΔV , the change in V_B for a given applied rf power level. The total dc power dissipated in the bridge is given by V_B² divided by 200 ohms, the series-parallel combination bridge resistance. Since each leg of the bridge has

equal resistance, the dc power dissipated in the rf sensor is 1/4 the total dc power dissipated in the bridge. The rf power applied to the sensor, P_{rf} , is equal to the difference in dc power dissipated in the sensor with no rf applied and the dc power dissipated in the sensor with rf applied, as expressed by

$$P_{rf} = \frac{1}{4} \frac{V_{BE}^2}{200} - (\frac{1}{4}) \frac{(V_{BE} - \Delta V)^2}{200}$$
(1)

where V_{BE} is the equilibrium voltage at the top of the bridge with no rf applied and ΔV is the change in



Construction of the rf sensor showing the two incandescent lamps and chip capacitors C1, C2. Components are mounted on a small section of double-clad PC board which is soldered to the rear flange of the BNC connector.

bridge voltage following application of rf. Solving the above equation algebraically for ΔV results in the following solution:

$$\Delta V = V_{BE} - \sqrt{V_{BE}^2 - 800P_{rf}}$$
(2)

A given desired full-scale rf power is used in eq. 2 to determine a corresponding ΔV . The required value

table 1. Calculated values for calibration resistors for the rf power meter (V_{BE} = 3.5 volts, $I_{FS} = 200~\mu A$).

Prt	ΔV	RCAL
10 µW	1.143 mV	R4 = 5.715 ohms
30 µW	3.430 mV	R5 = 17.15 ohms
100 µW	11.450 mV	R6 = 57.25 ohms
300 µW	34.460 mV	R7 = 172.30 ohms
1 mW	116.200 mV	R8 = 581.10 ohms
3 mW	361.500 mV	R9 = 1808.00 ohms
10 mW	1.438 V	7192 ohms (see text)

of R_{CAL} for proper full-scale reading is determined by dividing ΔV by I_{FS}, the full-scale value of meter current. **Table 1** shows the calculated values for the meter shown. Similar calculations should be made when duplicating the wattmeter, using the measured values of V_{BE} and I_{FS}.

The equation for ΔV is the equation of a parabola. Thus, the meter current varies parabolically instead of linearly with rf power. On the low-power scales, however, the voltage varies over such a small sector of the parabola that for all practical purposes it is linear. On the highest scale, the deviation from linear becomes significant, and is such that when the meter is calibrated for an accurate full-scale reading, the indicated power will be less than the actual applied power at levels below full scale. Table 1 shows that a resistance value of 7192 ohms is needed for proper full-scale calibration of the meter on the 10 mW scale. For accurate calibration near the bottom of the 10 mW scale, a resistance 1000 times the value of the calibration resistor for the 10 µW scale, or 5715 ohms, would be required. Therefore, without some form of compensation, readings made near the bottom of the 10 mW scale will be only 79 per cent of the actual value, or 1 dB low.

To avoid lettering a special nonlinear 10 mW scale on the meter face, I used a compensation network. A compromise value of the calibration resistor R10 was selected at about 6000 ohms to reduce the error at the low end of the 10 mW scale. On the same scale, switch S2B connects the series combination of CR4 and R18 across M1 and R19. Toward the high end of the 10 mW scale, CR4 begins to conduct, shunting the excess current past M1. Variable resistor R18 determines the amount of current shunted away from M1, and variable resistor R19 determines the point at which diode CR4 begins conducting.

Before adjusting R18 and R19, an accurate voltmeter is connected from the top of the bridge (at V_B) to the reference voltage at pin 3 of U2A to read ΔV . A value of ΔV corresponding to a full-scale reading of 10 mW (1.438 volt in the meter shown) is artificially established by adjusting the reference voltage level. Then R18 is adjusted for a full-scale reading of the power meter. A ΔV corresponding to a reading of 6 mW (0.7705 volt in the meter shown) is then set and R19 is adjusted for a reading of 6 mW.

Since these two adjustments interact, they should be repeated several times until the meter reads both 10 mW and 6 mW. Linearization is now complete and the meter should be found to be quite accurate at all power levels.

The adjustment of R19 has no effect on the calibration of the other scales, provided the output of U2B is not at saturation for full-scale deflection of



fig. 3. Plot of lamp resistance vs power dissipated in the lamp.

M1. Since the nonlinearity on the 3 mW scale is such that readings on the low end of this scale are only 5 per cent low (-0.23 dB), no linearization was deemed necessary for this and lower scales.

For most accurate results, the values of R_{CAL} used for the lower scales should be as close to the calculated values as possible. Junk box resistors within a per cent or two of the desired values were selected using a digital ohmmeter. Where a proper value could not be found, a series or parallel combination was used.

The dB scale was added to the meter face so power could be read directly in dBm (dB with respect to a milliwatt) and so that losses and gains could be read out directly in dB. The scale position corresponding to each dB mark is given by

$$P = \frac{1}{antilog_{10}(0.1X)}$$
(3)

where *P* is the relative scale position (with 1 = full scale) and *X* is the number of dB below full scale.

procedure for use

Following turn-on, the meter is allowed to stabilize and the desired scale is selected. In the meter shown, stabilization is almost immediate on the higher power scales; several minutes are required on the 10 μ W scale before warm-up drift ceases. Once the meter has sufficiently stabilized, the zero adjust pot, R15, is adjusted for zero reading. The rf power is then applied and readings are made. Provided the lamps are not burned out, the meter will not be damaged by exceeding the maximum power for the scale selected. Since the lamps can safely dissipate 200 mW, a considerable margin of safety exists. Random drift is significant on the 10 μ W scale; thus the meter zero should be checked between readings for greatest accuracy when using that scale.

measured performance

Following calibration as described, the rf wattmeter was connected through one foot (30cm) of RG-58/U coaxial cable to the calibrated output of a Wavetek 3001 rf generator. Over the frequency range from 1 to 500 MHz, the generator power setting agreed to within 0.3 dB of the wattmeter reading at full scale on all wattmeter scales. The good agreement cannot be taken as a claim for wattmeter accuracy, however, because the specified worst-case generator power error on the most accurate power range is only 1.25 dB.

At 432 MHz, the input swr of the wattmeter was measured at 1.6:1. When measuring power from a 50-ohm source at 432 MHz, the resulting reading is calculated to be 0.24 dB low, due to reflected power. If the impedance of the source is adjusted to conjugately match the load presented by the wattmeter and interconnecting low-loss cable, this source of error is eliminated. On lower frequencies, the swr and resulting mismatch loss are expected to be even less because the parasitic reactance of the lamps and fixtures will be lower.

The wattmeter sees nearly constant use in testing rf circuits and devices of all types. Used directly, or with attenuators, it measures gains and losses. Used with directional couplers, hybrids, or rf bridges, it measures reflected power, return loss, and standing wave ratio. Since the lamps are a high temperature 50-ohm load, the wattmeter is also used as a noise generator for receiver rf amplifier tuneup and testing. In the few months since its construction, the wattmeter has become a virtually indispensable addition to my test bench.

references

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drift-correction circuit

for free-running oscillators

If you're bothered by warmup drift in your transceiver, here's a circuit that provides automatic compensation and uses readily available components

The principle of drift correction of an oscillator can be used in receivers or transmitters to compensate for warmup drift. The principle can also be used in new designs using simple free-running oscillators instead of the more complex types that use heterodyne mixing or phase-locked loops.

The idea is simple and straightforward. It can be best explained if you consider the operation of a frequency counter in which an oscillator frequency is measured. If the counter gate time is one second, and if sufficient displays are present, a 14-MHz signal could be displayed as 14.012.345 MHz. If, after the next measuring period, the least-significant digit changes from 5 to 7, for example, the oscillator frequency will have drifted 2 Hz high during that period. To counteract the drift, you could manually tune the oscillator back to its original frequency after each measurement. But there's a better way — read on. In the system described here, oscillator drift is compensated automatically. Only the last digit of the counter display is inspected after a measurement period. It is checked if the number is above or below a fixed value (5 in the example above). For values of 6, 7, 8, or 9, a voltage on a varicap in the oscillator reduces the frequency; for values of 0, 1, 2, 3, and 4, the reverse action occurs.

From this simple example it can be seen that:

1. The oscillator frequency always varies at a slow rate around a fixed value.

2. Stable points occur within 10 Hz from each other over the vfo tuning range.

3. Drift and short-term stability of the oscillator must be within limits. In the example cited, the drift must not exceed a few Hertz per second, otherwise the circuit can't compensate for the drift.

4. The automatic correction should be *very light*. If, after one correction period, the frequency overshoots too much, the remedy is worse than without the system.

For proper operation the correction-circuit time constant must be rather long (but also short enough to counteract the "natural" drift). Because of the long time constant, tuning feels quite normal. After a manual frequency adjustment, the frequency will creep to its nearest "stable" point (actually an unstable point) and will remain there. Because these points are closely spaced you don't notice the operation of the system by listening to a CW or ssb signal.

Note that, for correct operation of the system, the time base frequency doesn't have to be exactly 1 Hz, but the time base must be very stable. Thus the time base must be derived from a crystal oscillator. Counting can be in binary instead of binary-coded decimal format.

circuit description

The circuit is shown in **fig. 1**. Only one stage of a counter is required. A 74LS93 binary counter (U1) counts the oscillator frequency that is to be stabilized. This stage is preceded by a 2N709 transistor

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(Q1) to obtain sufficient sensitivity. About 100 mV of input signal is required.

After each counting period, the value of the 2^3 output (Q_c , pin 8, of U1) is stored in a D-type flip-flop, U2, (half of a CD4013) at the rising edge of the time base signal. The flip-flop output drives an integrator (U3) up or down, which in turn drives a varicap in the oscillator to correct the frequency.

The time base frequency that actually determines system stability is derived by dividing the frequency of a crystal oscillator. A 1-MHz crystal oscillates with one input gate of a CD4060, (U4), which also contains 14 binary dividers. In combination with a CD4020, (U5), these two circuits divide the 1-MHz frequency by 2¹⁸ to about 3.81 Hz, so the stabilization points are spaced at 3.81 times 8 Hz, or 30.5 Hz.

I found that FT241 crystals between 400 and 500 kHz oscillate very well in this circuit. The total dividing factor should be 2^{17} in that case, which can be obtained by using output pin 2 of U4 instead of pin 3, as shown in **fig. 1**.

The counter counts almost continuously. Just after the transfer of the state of the Q_c output to the D-type flip-flop (U2), a short reset pulse is generated by the other half of the flip-flop (U6). To achieve this action, the clock input signal of U6 is delayed by R1C1. After the Q output is set, the flip-flop resets itself because the Q output is connected through R2C2 to its own reset input. The resulting positive-going pulse is about 0.5 microsecond duration (line 3, fig. 2). This pulse resets the 74LS93 counter to zero which starts counting again immediately thereafter.

Worth mentioning is the long time constant of the integrator, which is formed by R3 and C3 (fig. 1). Capacitor C3 must be a low-leakage type, not an

electrolytic. A polystyrene or polycarbonate type will do.

The switches labeled UP and DOWN (fig. 1) serve a dual purpose. First, after circuit switch-on, the integrator output can be brought into its range manually; but also, small frequency variations can be made by pushing the UP or DOWN button. So a push-button-controlled fine tuning is obtained, which is convenient if, for example, a CW signal slowly drifts out of a narrow CW-filter passband. (With this system installed you can be sure it's the other station that drifts.)

The CA3140, a very convenient operational amplifier, is used because of its high fet input impedance. The integrator output signal can be monitored on a meter to verify that it's still within its operating range. The action of the varicap in the oscillator must be such that a 10-volt output variation of the integrator shifts the frequency about 3 kHz.

construction

The circuit was built onto a piece of *Vero* board and installed in my CW transceiver. A doublebalanced diode mixer is used in my rig, so a highlevel oscillator signal was available.

The UP and DOWN pushbuttons were mounted on the transceiver front panel. The control signal was monitored in a particular position of the transceiver meter switch.

Several prototype circuits were built using different construction methods, such as mounting all components on a copper-clad board with the ICs in sockets, but mounted upside down so that the socket pins could be wired directly. All these prototype circuits worked well, so the layout shown shouldn't be too critical. Just make sure that you avoid long wires between the ICs.



fig. 1. Circuit for vfo stabilization.

The circuit shown has been used for quite some time in my transceiver, which has a free-running oscillator on all bands. The highest frequency is 21 MHz, but the circuit has been used experimentally with oscillators operating to 40 MHz.

Within one minute after switch-on, the transceiver has crystal-quality stability on all bands. The 30-Hz frequency spacing between stabilization ponts is more than adequate for CW and ssb work. Also, during transmission, with about 200 watts to the anten-



fig. 2. Timing sequency of signals in the circuit of fig. 1. The time base is 3.8 Hz.

na, a jump to another stabilization point has never occurred.

A kind of proportional control system was tried instead of the constant-speed system described above. In this system, the corrective action depended on the offset value. Although the control could be measured (to be more effective), I believe this idea is really not worth the more complex electronic circuitry. Reason: with both systems a vfo becomes virtually drift-free, and both systems are not noticed during operation.

conclusion

The system described here doesn't turn a bad vfo into a good one but helps to make a good one even better. Especially where a low-noise oscillator is important, as for local oscillators in high dynamic-range front ends for receivers to obtain low reciprocal mixing, I believe this technique could be applied successfully at least for hf-band applications.

Synthesized oscillators appear to be noisier than good free-runnng types so if this system is used in combination with a digital frequency read out, on a well-designed, free-running oscillator, a much simpler system results than is possible with fully synthesized oscillators, giving at least the same or better results.

ham radio



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active bandpass filters -

some staggering thoughts

Here's a rundown on stagger-tuned filters using op amps as active devices great idea for many amateur applications

Applications for active bandpass filters in the audio-frequency range can be found in every part of amateur radio. Audio selectivity for CW, speech processing for ssb, tone-detector filters for RTTY, and control-tone separation for fm repeaters are only a few of the uses. In this article you'll learn an easy way to design and build stagger-tuned operationalamplifier active filters to fit your requirements. All you need is one of the readily available hand-held scientific calculators (or some other method for calculating square roots and logarithms).

Perhaps you've seen other types of active filters or filter designs using LC components. Why use stagger-tuned filters, and why use active filters? It's easy to build very narrowband audio filters by cascading, one after another, several identical simple filter sections. This may be adequate for some tasks but can often leave a lot to be desired in terms of transient response (*ringing*), peaked or *narrow-nosed* amplitude response, and poor skirt selectivity (*shape factor*). Conventional circuits using inductors can give excellent performance if well designed, which is often done with complex computer-aided design programs. But inductors are often large and hard to tune. Many amateurs have been discouraged by the need to add or remove turns from the 88-mH toroidal inductors common in RTTY use.

features

The filters described here offer many advantages. They give amplitude response with flat or slightly rippled characteristics in-band. Out of band, they have excellent skirt selectivity and a shape factor that improves directly as more filter sections are added. As a bonus, the transient response is usually much better than narrow-nosed filters. Best of all, each stage can be tuned separately with no measurable interaction or detuning of the other stages — this is a real plus for experimenters.



fig. 1. Typical stagger-tuned response. By choosing the correct peak frequency, f_n , and Q for each stage we get the response shown.

Now the bad news (which isn't really too hard to take). Stagger tuning requires that each stage provide enough gain so that the sum of the stage gains is greater than that of the overall filter. This is because of staggering loss, of which you'll see more shortly. With op amp ICs and their large open-loop (no feedback applied) gain, this parameter turns out to be of little concern. Another problem is that if one

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of the stages is out of tune, the filter response can be poorer than in designs that purposely introduce interaction between filter sections, as in most LC designs or in *leapfrog* active filters, which are much more difficult to design.

description

The stagger-tuned filter is made of two or more stages, each having a different peak frequency, f_n , with an associated Q (which may be the same as the Q of one of the other stages), and a certain amount of gain, G. The sum, in dB, of the gains versus frequency can be arranged to give a flat response over the band of interest. Fig. 1 shows how this happens. In the area between the two peaks one response rises as the other falls. By choosing the right f_n and Q for each stage, we get the response shown. Note that, at the center frequency of the overall response f_{O} (where the stages have equal loss), the net loss is twice as much (in dB). This is the stagger loss, S, which must be made up by the sum of the individual stage gains to give unity gain overall. Compare figs. 1 and 2. Fig. 2 shows a two-stage nonstaggered or "synchronously tuned" filter response with the same 3-dB bandwidth. Note the poorer skirts and the much rounder passband.



fig. 2. Response of a two-stage synchronously tuned filter (compare with the response in fig. 1).

In the stagger-tuned filter, the shape of each stage response is of the classic single-resonator shape (the same as that generated by a single parallel LC circuit with a shunt resistance to define the Q). The amplitude response is defined mathematically (for those of you itching to use your HP-25) as follows:

$$\frac{V_{out}}{V_{in}} = -10 \log_{10} \left[1 + Q^2 \left(\frac{f}{f_n} - \frac{f_n}{f} \right) \right]^2$$
(1)

Eq. 1 is of interest only and is not necessary for



fig. 3. A two-stage Butterworth filter showing α or d as functions of fractional bandwidth, δ .

designing a filter. It can be used for analysis, however. You can find the response of each stage then add all responses together to find the overall filter response. One thing that's important to note is that the curve has *geometric symmetry*. All this means is that if the upper (x) dB-down point is two times the center frequency, then the lower (x) dB point will be at one-half f_n . This relationship is expressed by

$$f_n = \sqrt{f_L f_H} \tag{2}$$

where f_L is a frequency below f_n with the same attenuation as f_H , which is higher than f_n . Note that f_n is *not* the arithmetic average of f_L and f_H . The resultant overall filter response will exhibit the same type of symmetry as the stages of which it is composed. So **eq. 2** holds for the complete filter, where n is zero.

Design procedure. In designing the filter, the first thing is to decide what type of filter is wanted. The Butterworth, or maximally flat filter, provides the flattest passband and a good skirt shape. The Chebychev or equal-ripple filter gives ripples in the passband (1 dB in the designs to follow), but in turn, it has very rapid cutoff of the band. Many other filter types are in use, but these two will serve you well.

Next you must determine how many stages you want. This requirement is determined by the required shape factor, with the other consideration being how much circuitry you want to build. High *Q*s and more precise tuning of the stages are also requirements of the higher-performance designs.

Shape factor. To refresh your memory, shape

table 1. Shape factors for Butterworth and Chebychev filter designs.

number of	butterworth		1-dB chebychev	
stages	3/30 dB	6/60 dB	3/30 dB	6/60 dB
1	31.60	577.00	31.60	577.0
2	5.62	24.00	4.70	20.8
3	3.16	8.33	2.30	6.6
4	2.37	4.90	1.75	3.5
5	1.99	3.57	1.60	2.5
6	1.78	2.89	1.30	2.0

factor (also called *selectivity ratio*) is the ratio of bandwidth at a higher attenuation to the bandwidth at a lower attenuation. Most common is the 6 - 60-dB



fig. 4. Three-stage Butterworth filter showing α or d versus fractional bandwidth, δ .

shape factor. **Table 1** shows this shape factor versus the number of stages for Butterworth and 1-dB-ripple Chebychev filters. Also given is the 3 - 30-dB shape factor.

After deciding the type and complexity of the filter, specify the lower 3-dB point, $f_L(3 dB)$, and the

table 2. Approximations for fractional bandwidth, δ_i equal to or less than 0.3 for Butterworth and Chebychev active filters.

Butterworth

two-stage	
$\alpha_1 = 1 + 0.365\delta$	$d_1 = 0.707\delta$
three-stage	
$\alpha_1 = 1 + 0.450\delta$	d ₁ = 0.500δ
four-stage	
$\alpha_1 = 1 + 0.485\delta$	d ₁ =0.380δ
α ₃ = 1 + 0.195δ	d ₃ = 0.920δ

Chebychev (1-dB ripple)

two-stage	
$\alpha_1 = 1 + 0.365\delta$	$d_1 = 0.433\delta$
three-stage	
$\alpha_1 = 1 + 0.450\delta$	d ₁ =0.220δ

upper 3-dB point, $f_H(3 dB)$. Then use **eq. 2** to find f_O . Next find δ , the fractional bandwidth.

$$\delta = \frac{(f_H - f_L)}{f_0} \tag{3}$$

This parameter, δ , is the main design factor. It's used to find tuning data for each stage. Refer to **figs. 3, 4**, or **5** for Butterworth filters of two, three, or four stages respectively. For a 1-dB Chebychev filter of

table 3. Design equations for two-, three-, and four-stage filters. Parameter α is the ratio of resonant to filter center frequency.

for two-stage filters

$f_1 = (f_0)(\alpha_1)$	$f_2 = f_0 / \alpha_1$
$Q_1 = 1/d_1$	$Q_2 = 1/d_1$
$G_1 = G_2 = (S + G_2)/2$	

for three-stage filters

 $\begin{array}{ll} f_1 = (f_0)(\alpha_1) & f_2 \equiv f_0 & f_3 = f_0/\alpha_1 \\ Q_1 = 1/d_1 & Q_2 \equiv 1/\delta & Q_3 = 1/d_1 \\ G_1 = G_2 = G_3 = (S + G_0)/3 & \end{array}$

for four-stage filters

 $\begin{array}{ll} f_1 = (f_0)(\alpha_1) & f_2 = (f_0)(\alpha_3) & f_3 = f_0/\alpha_3 & f_4 = f_0/\alpha_1 \\ \Omega_1 = 1/d_1 & \Omega_2 = 1/d_3 & \Omega_3 = 1/d_3 & \Omega_4 = 1/d_1 \\ G_1 = G_2 = G_3 = G_4 = (S + G_0)/4 \end{array}$

two or three stages, see **fig. 6** or **7** respectively. From the appropriate figure, obtain α_1 and d_1 (and α_3 and d_3 for a four stage Butterworth). If your filter has a $\delta \quad 0.3$, **table 2** offers approximations for α and d, which usually give better accuracy than reading from the graph. Decide what overall gain, G_0 , in dB you want from the filter, then use **table 3** to find the tuning frequency, the Q, and the gain for each stage.

It's a good idea to organize the stages as given, with the highest-frequency stage first. (This mini-



fig. 5. Four-stage Butterworth, with α or d as functions of fractional bandwidth, δ .



fig. 6. A Chebychev two-stage filter (1-dB ripple) showing α or d as functions of fractional bandwidth, δ .

mizes harmonic distortion for the overall filter.) The higher-frequency stages have the lowest open-loop gain, which means that feedback will be less effective in reducing the distortion in these circuits than in the lower-frequency stages. Putting the low-frequency stages last gives maximum attenuation to any harmonics generated by the higher frequency stages.

Multiple-feedback circuit. Now that you know what the stages must do, the only thing remaining is to design circuits with the required f_n , Q, and G. For stages with low Q (less than 10), the multiple feedback (MFB) circuit in **fig. 9** performs well. Almost any op amp will work here, but depending on its bandwidth, limitations exist on maximum Q and maximum f_n .

The upper limit on Q for the MFB circuit is given by the smaller of

$$Q_{max} \cong \sqrt{f_T/(5f_n)},$$
 (4)
 $Q_{max} \cong 10$

where f_T is the frequency at which the op-amp gain equals zero dB (unity gain). The frequency, f_n , should be limited to about 1 per cent of f_T (10 kHz for a 1-MHz f_T amplifier, such as the type 741).

These restrictions minimize the effects of amplifier gain on f_n and Q, which ensures accurate calculation of these parameters and freedom from drift because of amplifier gain changes with temperature.

The component values in the MFB circuit can be found easily. Choose convenient value of capacitor, *C*. The resistors are:

$$R3 = \frac{Q}{\pi f_n C} \tag{5}$$

$$R1 = \frac{R3}{2 \cdot 10^{(G/20)}}$$
(6)



fig. 7. Chebychev filter with three stages (1-dB ripple) showing α or d versus fractional bandwidth, δ .

$$R2 = \frac{1}{\left[(2\pi f_n C)^2 R3 - (1/R)\right]}$$
(7)

Note that $[10^{(G/20)}$ equals $antilog_{10}(G/20)]$, where G is the gain, as described previously.

State-variable design. The limitations of the MFB circuit require that a higher-performance circuit be used in some cases. The state-variable circuit in **fig. 10** can do some amazing things. It can provide very high *Q*s (over 100) and is hard to beat for stability and lack of sensitivity to passive component drift. However, it does take two more op amps and four more resistors than the MFB design.

There are several degrees of freedom in this design. Choose C, R2, and R4 for convenience.* The remaining resistors are found from

$$R1 = R_2 Q / (10^{G/20})$$
 (8)

$$R3 = \frac{1 - \frac{J_n}{f_T}}{2\pi f_n C}$$
(9)

$$R5 = \frac{R4}{\left[\frac{2Q + (10^{G/20})}{\left[1 + \frac{4Q + (10^{G/20})}{\frac{f_T}{f_n}}\right]} - 1\right]}$$
(10)

*A "convenient" capacitor is one as small as possible that doesn't require overly large resistors. Choosing resistors too much above 100k (for 741s or similar op amps) can lead to excessive dc offsets because of input-bias currents. Fet input op amps have extremely small bias currents and will tolerate resistors in the tens of megohms. For the MFB circuit, capacitors with about 10 kilohms of reactance are in the ballpark. For instance, at 1500 Hz, a 0.01 μ F capacitor is suitable. In the state-variable circuit, capacitors of about 100k ohms of reactance can be used, such as 0.001 μ F at 1500 Hz. A reasonable value for *R2* or *R4* is between 10k - 100k.

Both circuits can be impedance-scaled if the calculations of component values reveal one or more values that are out of the desirable range. This means that all resistor values may be changed so long as all change by the *same ratio*, and the capacitors change by the *reciprocal* of that ratio. For example, if you find a 300k resistor where you'd like to have 100k, you can change it by making all the resistors one-third of their original value and by making the capacitors three times as large. In the state-variable circuit, R4 and R5 may be changed independently of the other resistors so long as the ratio R4:R5 is constant.

design example

The design procedure is used to create an input prelimiter filter for an RTTY demodulator (TU). We'll



fig. 8. Loss due to staggering, S_i as functions of fractional bandwidth, δ_i for various active filters.

choose an overall gain of 30 dB to provide adequate drive to the limiter from normal speaker signal levels. To give flat response in-band and reasonable delay distortion (associated with the transient response), we'll choose a Butterworth design. For good selectivity a four-stage configuration will be used. For 170-Hz shift and 45.45 Baud (standard 60 wpm), the



fig. 9. Schematic showing the MFB, or multiple-feedback circuit.

CCIR formula shows the bandwidth to be 246 Hz. To allow for tuning error and drift, a 300-Hz bandwidth at the 3-dB points will be used. The mark frequency is 2125 Hz; the space frequency is 2295 Hz. Thus the passband should be from $f_L = 2060$ Hz to $f_H = 2360$ Hz. Eq. 2 gives

$$f_o = \sqrt{(2060)(2360)} = 2205 \, \text{Hz}$$

From eq. 3 we obtain the fractional bandwidth

$$\delta = \frac{(2360 - 2060)}{2205} = 0.1361$$

Since δ is less than 0.3, use the approximations in table 2.

 $\begin{aligned} \alpha_1 &= 1 + (0.485) \ (0.1361) = 1.066 \\ d_1 &= (0.38) \ (0.1361) = 0.0517 \\ \alpha_3 &= 1 + (0.195) \ (0.1361) = 1.0265 \\ d_3 &= (0.92) \ (0.1361) = 0.1252 \end{aligned}$

From fig. 8 the loss due to staggering, $S_{,} = 18.2 \text{ dB}$ and from table 3 we have

$f_1 = (2205)(1.066)$	= 2351 Hz	$Q_1 = 1/0.0517 = 19.$	3
$f_2 = (2205)(1.0265)$	= 2263 Hz	$Q_2 = 1/0.1252 = 8.$	0
$f_3 = 2200/1.0265$	= 2148 Hz	$Q_3 = 1/0.1252 = 8.$	0
$f_4 = 2205/1.066$	= 2068 Hz	$Q_4 = 1/0.0517 = 19.$	3
and $G = (18.2 + 30)$)/4 = 12.05 dB	(per stage).	

It's apparent that the state-variable circuit must be used for the first and fourth stages (Q > than 10). At 2351 Hz a 741-type op amp is capable of

$$Q_{max} \approx \sqrt{\frac{10^6}{5(2351)}} = 9.2$$

Since the second and third stages have *Qs* less than this, the MFB circuit is usable.

Let the capacitors in the state-variable stages be 0.001 μ F and the capacitors in the MFB stages be 0.01 μ F. Let R2 and R4 be 100k in stages 1 and 4. From **eqs. 8**, **9**, and **10** we obtain the values for the first state-variable stage:

$$R1 = (100k) (19.3/10^{12.05/20}) = 482k$$



fig. 10. Schematic of the state-variable stage.

$$R3 = \frac{1 - \left[\frac{2351}{10^6}\right]}{(2\pi)(2351)(10^{-9})} = 67.54k$$

$$R5 = \frac{10^{5}}{\left[\frac{(2)(19.3 + 10^{12.05/20})}{1 + (4)(19.3 + 10^{12.05/20})}\right]} = 2686 \text{ ohms}$$
$$= 2686 \text{ ohms}$$
$$= 1$$

Similarly, for stage four, we find

Now for stage two, using eqs. 5, 6, and 7,

 $R3 = 8/\pi(2263) (0.01 \times 10^{-6}) = 112.5k \text{ ohms}$ $R1 = 112.5k/(2) (10^{12.05/20}) = 14.05k \text{ ohms}$

$$R2 = \frac{1}{\left[2\pi \left(2263\right) \left(0.01 \times 10^{-6}\right)\right]^2 \left[112.5k - \left(\frac{1}{14.05k}\right)\right]} = 439.6 \text{ ohms}$$

And in the same fashion, for stage three, we have

construction

The filter was constructed using two MC3303 quad op amps. Combinations of one per cent resistors were used to give the calculated values within 0.5 per cent or less, nominally. Polystyrene capacitors, 1 per cent tolerance, were used in all sections. The measured response of the filter before tuning is shown in **figs. 11** and **12**. The calculated response which is given for comparison, was generated using **eq. 1** for each stage and then adding the four responses.

Normally, filter sections will need trimming for frequency and/or Q. In many low Q filters ($\delta = 0.3$), 5per cent tolerance resistors will give quite satisfactory results without trimming. The only penalty may be slight center frequency error and perhaps a small amount of skew in the passband frequency response.

For the narrowband filters, and especially those with three or four stages, an audio generator, ac voltmeter, and frequency counter will help in trimming each stage independently to the required parameters. In the state-variable circuit, adjust both R3 values to set the center frequency, then use R5 to fix the Q. Remember Q is the 3-dB bandwidth divided by f_n . For an MFB stage, adjust R3 to give the desired 3-dB bandwidth. Then adjust R2 to set f_n . Varying R2 has virtually no effect on the bandwidth, which means the Q changes at the same rate as f_n . After tuning the RTTY demodulator input filter, the overall response was essentially indistinguishable from the calculated response.

components

Generally, components should be the best you can get. Metal-film resistors and polystyrene or mylar capacitors are hard to beat, but may be *overkill*. Stay away from capacitors designed for bypass or coupling use; their tolerance is poor, as is their stability. Carbon resistors are usually adequate in all but the narrowest filters. For op amps, 741s are suitable (as are the 1458 dual versions and the quads like the



fig. 11. Response as a function of frequency for an RTTY input filter. Solid line: calculated; dots: measured data before tuning.

3303), when used within the limitations given above. The LM318 op amp gives much greater freedom from Q drift (in the state-variable circuit) and f_n drift (in the MFB circuit). Some of the new wideband fet-input op amps, such as the LF356, should be excellent performers. When external frequency compensation is required, use the values specified for unity gain amplifiers.



fig. 12. Passband response of an RTTY input filter.

When interfacing active filters, take care that the source impedance driving the filter is very low, i.e., less than 1 per cent of R1 in either circuit. Another op amp or a voltage follower provides an excellent driver. If the requirement for a low impedance can't be met, deduct the source resistance from the value of R1 in the first stage.

You've seen an easy-to-use method for designing stagger-tuned active filters to your own needs, and have learned to avoid some of the possible pitfalls. Now you can replace that filter you borrowed from someone else's circuit that never did work exactly the way you wanted.

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Over the years many amateurs have traded their old, general-coverage receivers for shiny new "hamband-only" models. We've gained in stability, sensitivity, selectivity, dial accuracy, and many other attributes; but we've lost on frequency coverage. Except for a few narrow windows to the outside world. we can listen only to each other. Those with interests outside the amateur bands have had to use a second receiver (frequently an old general-coverage job) and put up with drift, bulk, and lack of accurate calibration. The XPL Converter, described here, is designed to work with a modern receiver to give the best of both worlds - extremely broad frequency coverage together with crystal stability and calibration accuracy. The converter receives all frequencies between 0 and 28 MHz when used with a receiver tuning 28 to 29 MHz. Construction is simple, straightforward, and inexpensive thanks to integrated circuits.

description

The XPL Converter consists of a wide-range tuned input circuit, 60 kHz to 28 MHz; a local oscillator with injection frequency switch-selectable in 1-MHz steps from 29 to 56 MHz; and a mixer circuit with output 28 to 29 MHz feeding the receiver as a tunable i-f amplifier. A block diagram is shown in **fig. 1**. Local oscillator output is taken from a vfo (vco), which is phase locked to a 100-kHz reference oscillator through a counter chain preset by thumbwheel switches for band selection. The various sections are described in more detail later.

An example may help clarify the frequency conversion technique employed in the *XPL*. If the vco is set at, say, 38 MHz, the tunable i-f range of 28 to 29 MHz will allow reception of signals from 38-29 = 9 MHz to 38-28 = 10 MHz. A 9330-kHz signal in this range would be received at 38-9.33 = 28.67 MHz. The receiver tunes *backwards*, in that the low-frequency end of each range will be received at 29 MHz and the high end at 28 MHz. This turns out to be only a minor operating annoyance, however. Low-side injection could be used for forward tuning but only at the sacrifice of tuning range at the upper end.

The vco is phase locked to the reference crystal, so the local oscillator is of crystal quality as far as accuracy and stability are concerned. Any input frequency can be precisely located and will be stable within the accuracy and stability of the receiver on the 10meter range. For most modern receivers, this means 1-2 kHz accuracy and a few hundred hertz drift on warmup. What a difference from the old generalcoverage boat anchors!

input circuitry

Input-circuit details are shown in **fig. 2**. A singletuned circuit provides input selectivity for the *XPL*. Six switch positions cover 60-150 kHz, 150-450 kHz, 450-1400 kHz, 1.4-4.5 MHz, 4.5-10 MHz, and 10-30 MHz. The four high-frequency ranges use a commercially available coil set having high-impedance balanced antenna windings. On the two low-frequency ranges pi-section single-ended input circuits are used with rf chokes for the inductors. Tuning is by a miniature broadcast superhet variable capacitor having a total capacitance of about 560 pF with the two sections in parallel.

Two input traps are used: a balanced lowpass filter to eliminate TV/fm pickup and a series-resonant

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trap to eliminate overload from any one broadcast station. Additional suppression measures may be required in unusual situations.

There's no need to conform to the input circuit shown. In fact, the antenna tuning section from a scrapped general-coverage receiver could be used to



fig. 1. Block diagram of the XPL converter.

handle the high-frequency end of the range. The low end can be extended with larger inductors, but the tuning range for each band will be quite limited because of distributed capacitance in the coils. Two additional coils, however, will allow tuning to about 15 kHz.

local-oscillator system

The heart of *XPL* is the local oscillator. This circuit consists of a voltage-controlled oscillator (vco), programmable divider chain, crystal-reference oscillator, and phase comparator. A block diagram is shown in **fig. 3**.

Phase-locked-loop operation has been well described in the literature, but a quick review may be worthwhile. A phase-locked loop is a feedback control system that measures the phase difference between two frequency sources and generates an error voltage that changes the frequency of one frequency source until the two sources are in phase synchronism. For continuing phase errors, the phase detector will function on frequency difference and steer the system into phase lock.

Two basic systems can be used to generate a selectable series of integrally related frequencies. If the phase comparator is sensitive to reference-oscillator harmonics, the controlled oscillator can be directly locked to a selected harmonic by first tuning it manually to a nearby frequency, then allowing the phase detector to lock up. This is the system used in several commercial receivers. The only objection from a construction point of view is that it requires a manually variable oscillator with dial calibration sufficient to resolve adjacent harmonics. A lock indication is also useful in identifying the proper harmonic.

A more direct way of generating the integrally

related frequencies is to divide the controlledoscillator frequency by programmable digital dividers before phase comparison to the reference frequency. If the oscillator frequency is divided by, say, 24 before the comparison is made, the effect is to lock the oscillator to the 24th harmonic of the reference frequency. In *XPL*, a fixed divide-by-ten and two programmable divide-by-n counters are used to enable lock from the 290th harmonic to the 560th harmonic of the 100-kHz reference frequency in steps of 1 MHz.

oscillator and phase comparator

Fig. 4 is the schematic for the reference oscillator and phase comparator. A 7400 quad NAND gate is used with a 100-kHz crystal to generate the reference frequency. There's no special merit to this scheme other than simplicity, and any convenient oscillator circuit could be used so long as it provides TTL output levels. In the circuit shown, the 0.0047 μ F capacitor at the input to the last gate was necessary to eliminate a double-pulsed output to the phase comparator.

A Motorola MC4044P phase-lock chip was chosen because it offers TTL logic, a nonharmonic-sensitive



fig. 2. Input-circuit schematic.

comparator, and some internal auxiliary transistors. Also, its use in synthesizers has been described in recent articles.

Output from the MC4044P is buffered by an external 2N5457 fet follower and the internal emitter followers. The comparator has unity gain from the phase detector to the output. An active filter is backed up by two poles of rolloff for loop stability and high 100-kHz ripple attenuation. The reference oscillator and phase comparator are supplied from an on-board regulator that provides both isolation and filtering.

voltage-controlled oscillator

The voltage-controlled oscillator in the *XPL* (fig. 5) uses a Motorola MC1648L ECL chip designed for this service. Spectral purity requirements preclude a voltage-controlled multivibrator, so this chip was used with an external high-Q toroidal inductor and a Motorola MV1401 variable-capacitance diode or varicap. Since ECL has a very low logic swing, an



fig. 3. Local-oscillator block diagram.

output translator, 2N4403, is used to regenerate the TTL signal level. At this point you might ask whether the ECL chip is worth the effort. The answer is a qualified "Yes," since it functions from a two-terminal tank circuit and eliminates the need for fussing with feedback in a transistor oscillator.

The MV1401 varicap is rather expensive (in the \$9.00 range), but it has a guaranteed 3:1 tuning range and high Q. This application requires only a 2:1 range, but allowances for temperature variation component tolerances, and other considerations make it necessary to have some overrange. Less-expensive limited-range diodes could be used, but they would require changing fixed capacitors to cover the tuning range for the vco. The inductor is a T-25 mix 6 toroid with four turns of no. 22-28 AWG (0.6-0.3mm) enameled wire.

An output to the mixer is taken directly from the 50 ohm vco output at pin 3. For the TTL counters,

however, the swing is wrong. The sum of one diode drop and a base-emitter drop from the 5-volt-supply rail places the 2N4403 base voltage in the ECL logic voltage range. The 2N4403 collector voltage swings from 0.5V to about 3.5V to drive the counter. A 22ohm base-emitter resistor aids junction recovery and cleans up the output waveform.

The entire vco section is quite susceptible to hum and modulation disturbances. For this reason, a separate voltage regulator is again used. The vco should be located well away from transformer fields or ac power wiring.

programmable counters and translators

This circuit is shown in **fig. 6**. Before getting into counter details, a related matter must be considered. The count set into the preset counters must always be 29 (MHz) higher than the bottom end of the input tuning range, so that the switches can read input range directly. This requirement leads to the necessity of translating switch settings to the counters. **Table 1** summarizes the required relationships. If decimal switches are used, the offset of minus 1 in the units digit can be provided by simply rewiring into the decimal-to-BCD diode matrix, as shown in **fig. 6**. Note, for example, that a switch indication of 4 is translated to BCD 1+2=3, which is 4 minus the required one unit. The **390**-ohm resistors establish a TTL logic zero for open-switch positions.

The tens digit is somewhat more messy. Switch indications must be translated up by 3 *except* when the units position is zero, which requires an uptranslation of only 2. Thus, 00 goes to 29, 01 goes to 30, 10 goes to 39, 11 goes to 40, and so on. Since a zero-units digit is translated to a 9 in the output, the presence of this 9 can be used to change the tens digit to an output lower by one integer.

Two sections of a 7400 quad NAND are used to accomplish this magic. A decimal-to-BCD diode matrix with an offset of +3 is used in conjunction with a second matrix with offset of +2. The proper matrix

table 1	. Relationship	between switch	settings and	l counters.
---------	----------------	----------------	--------------	-------------

input	sw rea	switch readings		counter presets	
range	ten	units	ten	units	
0-1	0	0	2	9	
1-2	0	1	3	0	
2-3	0	2	3	1	
9-10	0	9	3	8	
10-11	1	0	3	9	
11-12	1	1	4	0	
12-13	1	2	4	1	
19-20	1	9	4	8	
20-21	2	0	4	9	
21-22	2	1	5	0	
27-28	2	7	5	6	



Looking down on the XPL converter. Components and wiring are shown on top of the chassis.

is chosen by clamping diodes from the two 7400 outputs. If a units 9 is present, the 9 bus is high and the 9 bus is low. Under this condition, the +3 matrix diodes are clamped low, and the +2 matrix diodes are released. For units digits other than 9, the situation is reversed.

Those familiar with counter techniques may immediately conclude that this is the long way around the barn, and so it is. A simpler solution to the translation would be to use a precounter set to 29 to delay activation of the programmable counters until the first 29 counts have passed. However, this requires two more counter chips, involves a clock gate, and is more difficult to troubleshoot if problems develop. The approach shown was devised with the less-experienced builder in mind, since troubleshooting of the diode matrices can be done with a vtvm. Returning now to the counters, a high-speed 74196/8290 is used as a divide-by-10 prescaler to get the signal into TTL frequency range. This unit has a typical toggle frequency of 75 MHz and handles the 56-MHz maximum input with little effort. The programmable counters, 74192s, are preset by their respective diode matrices. Both are operated in the countdown mode and are cascaded and loaded through the borrow outputs.

Counter operation is as follows: The C_o output of the 74196/8290, pin 2, goes high once every ten input pulses from the vco. Each output pulse causes the units 74192 to count down by one count. When the count reaches zero, the borrow output goes low between pulses and causes the tens 74192 to count down by one count. It, too, generates a borrow pulse after reaching zero, and it is this pulse that's used to reset the system. The borrow pulse from the tens counter is used to load the preset number into each counter.

As an example of operation, suppose the thumbwheel switches are set to 08 (8-9 MHz range). The units counter will be preset to a count of 7 (8-1) and the tens counter to 3 (0+3). Following a reset pulse, the units counter will count down one count every ten vco cycles and first generate a borrow pulse after 70 vco cycles.

After this first 70 cycles from the reset pulse, the counter will again generate a borrow pulse every 100 vco cycles. Each of these borrow pulses causes the tens counter to count down by one count from its preset of 3 and to generate its own borrow pulse after 70 + 100 + 100 + 100 vco cycles. The tens counter borrow and reset pulse thus occurs 370 vco cycles *after* the first reset and then immediately resets the counters again.

The tens counter borrow output frequency is equal



fig. 4. Reference oscillator and phase-comparator schematic.



fig. 5. Voltage-controlled oscillator schematic.

to the vco frequency divided by (10xU + 100xT), where U and T are the units and tens presets respectively (7 and 3 in our example).

The tens-counter borrow output is also used to feed the phase comparator so that the vco frequency is locked to 10x7 + 100x3 = 370 times the reference frequency of 100 kHz. The vco is thus locked at 37 MHz, which is the required local-oscillator injection frequency for receiving 8 MHz with a 29-MHz i-f.

Resistor values shown for the diode matrices are fairly critical. Germanium diodes would provide more margin, but the circuit works well as shown. If the same nominal values are used, no problems should be experienced. Power for the counters and translators is provided by still another 5-volt regulator. This system draws several hundred milliamperes and may need a regulator heat sink.

mixer

A dual-gate, diode-protected mosfet is used for the mixer (fig. 7). The 40673 has good intermodulation characteristics and is simple to use. Output from the drain is taken through an output transformer, broadly resonant at 28.5 MHz, which provides a low impedance output to the receiver. This stage is powered directly from the 9-volt power supply, since decoupling is not a problem. An output switch pole allows the receiver input to be connected to the converter or to a high-frequency antenna. A second pole is used to ground the high-frequency antenna to minimize pickup when the *XPL Converter* is in use.

power supply

All operating power for the *XPL* is derived from a 12-volt transformer and bridge rectifier at about 10 volts (**fig. 8**). A 2N3055 is used as an active filter to reduce ripple. This transistor is much larger than required, but it's cheap, readily available, and needs no

heatsink. If the 9-volt rail is not reasonably clean, the received signal may be hum modulated. Ac input is switched by a third pole of the IN-OUT switch, and a pair of 0.02 μF capacitors are used for line bypassing. Note that these capacitors should have 600-volt ratings.

construction

Each circuit section was built on a separate printed-circuit board for easy testing and debugging. There's no real need to do this, however, and a single PC board might be easier to handle mechanically. The layout shown is also more compact than necessary. The entire unit could be built on perf board if generous ground conductors are used.

Coax cable should not be used for interconnecting circuits except for the vco output to the mixer, mixer output to the receiver, and input from the hf antenna. Other leads should be run in twisted pairs of no. 22-26 (0.6-0.4mm) hookup wire to minimize shunt capacitive loading on the TTL gates and to reduce inductive pickup in the phase-comparator circuitry. Coax cable should not be used in the input circuit, since the high capacitance of this cable could appreciably decrease the tuning range.

The 28.5-MHz output coil was a junk-box relic of unknown parentage. Any coil with a turns ratio of about 5:1 with a slug capable of resonating at 28.5 MHz will do. An inductance of about 3μ H is required.

Most of the parts for the *XPL* are available from surplus houses or other *ham radio* and *QST* advertisers. The MC4044P, MC1648L, and MV1401, however, will probably have to be ordered from a franchised Motorola distributor. Total cost is about \$20 for these items.

The individual regulators were Motorola types, but various National LM-series are equally satisfactory and widely available. The 2N4403 transistor can be

replaced with almost any high-frequency pnp transistor. Similarly, the 2N5457 can be replaced by other N-channel jfet devices, such as the MPF102 series.

All signal diodes should be 1N4148/1N914 or similar silicon computer diodes. As mentioned earlier, germanium diodes can be used in the diode matrices if desired. Power diodes are low voltage, plastic-lead-mounted types.

Capacitors can be ceramic units except for the antenna input capacitor (560 pF) and the vco 1500-pF capacitors which should be of low-loss polystyrene or mica construction.

TRANSLATORS

The entire counter and phase-lock system could be designed around CMOS circuitry except for the vco and the prescaler. CMOS chips are not widely available in surplus outlets but are rather inexpensive when purchased new. Power supply current could be considerably reduced by shifting to CMOS, and the diode translators could be run at a much higher impedance level.

adjustments and troubleshooting

DIVIDERS

The power supply forms a logical first item if the *XPL* is built in steps. Output voltage must be at least



fig. 6. Schematic of the XPL programmable counters and translators.



fig. 7. Mixer schematic.

7 volts to allow the individual regulators their required 2-volt input margin over 5-volt regulated output voltage. Frequency calibration of the reference oscillator can be done by zero beating with WWV or with WCFL, Chicago, on 1000 kHz. The vco should be checked for range by coupling a grid-dip oscillator to the toroidal coil. An input of 0.5 to 3.5-volts positive, derived from a separate source, should drive the vco from 25 to 60 MHz or so. If the frequency range is off, toroid turns may be trimmed or the 1500-pF capacitor value changed to suit. Input-coil slugs should be adjusted to allow coverage of all input frequencies with a bit of overlap.

Operation of the translators can be checked with a vtvm. Logic zero must be 0.8 volt or less, and logic 1 must be 2.4 volts or more — standard TTL levels. Preset counter operation can be checked with a triggered oscilloscope and a low capacitance (10X) probe. The 74196 output pulses will be visible on most inexpensive scopes.

operation

For best results, a good antenna system should be used with the *XPL*. One of the best is an 80-meter inverted V or dipole with open-wire feeders. Except for those few frequencies at which the antenna happens to be an odd number of quarter wavelengths long, its impedance will be quite high. Thus, the high capacitance of a grounded coax antenna feeder would result in serious signal attenuation. A simple long-wire antenna can be used if the end is brought directly to the *XPL* input terminals. If a separate



fig. 8. Power-supply schematic.

antenna system isn't available, any ungrounded antenna feeder system can be used by connecting either lead to the **A1** antenna terminal. Terminal **A2** should be grounded for single ended inputs.

The input circuit should be calibrated at least roughly so you can be sure the desired signal is being peaked. The input circuit provides the only rejection for 10-meter signals present on the antenna. Above about 15 MHz, this rejection may be inadequate to prevent strong 10-meter signals from coming directly through the mixer to the i-f. A resonant trap or a loosely coupled input circuit can be added if this problem proves troublesome. A balanced mixer would reduce the feedthrough, but the added complication seemed unnecessary. I suggest this as an alternative approach for those interested in experimentation.

final remarks

The XPL Converter has been fun to use. Broadcast stations pop up exactly where they are supposed to be. The Selected Cities Weather Summary, broadcast from Miami on RTTY has been interesting to print and peruse. WWV is available on all frequencies for calibration or a check on propagation conditions.

Aviation weather and general information is broadcast on the local low-frequency range station. Every international shortwave band can be received. International air-route traffic control from Miami and New York can also be monitored. And, near the top end, you can even listen to CB operations.

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short circuit

RTTY time/date printout

An important point was missed in **table 1** of the RTTY printout article which appeared in June, 1976, *ham radio*. Pin 4 of U15 should not be grounded but should have the appropriate BCD information for the tens of minutes digit.

As shown, the ten, minutes digit will only display up to 39 minutes instead of 59 minutes. In **fig. 4A**, pins 6 and 7 of the 7490s must be grounded; otherwise the circuit will only print the 19th as the date.

Advancing the date by moving the clock is a very tedious process. Overshooting will mean doing the entire thirty days over again. The diagram below shows a circuit that will permit



you to advance the date by one day with the flip of the switch. When the date resets at the end of the month, flipping the switch will advance the clock to 01, much easier than advancing the digital clock a complete 24hour period. Note that this advance circuit is designed to work with a low input so the date advance must be done before 2000 hours.

pi network design and analysis

Eq. 8 in the pi network design article, September, 1977, *ham radio*, should not have the radical sign on the right hand side of the expression; it should read



direct output synthesizer for two meters

In fig. 4 of the direct output synthesizer in August, 1977, ham radio, the lines connected to pins 10 and 8 of U3C have been transposed. For correct operation, pin 10 is connected to the pin 9s of the 74161s, and pin 8 of U3C is connected to pin 1 of U2B. On U1, pin 2 is the input from U3D and pin 1 should be connected to the junction of the 100 and 360 ohm resistors. U1 may exhibit some temperature and voltage sensitivity at times causing the divide-by-21 function to become a divide-by-22. This problem can be cured by either of two methods: putting a 330 pF capacitor from pin 2 of U1 to ground or replacing U3 with a 74L00 instead of the 7400. U8 is a 7483, not a 7473. In fig. 6, the 0.1 µF capacitor connected to pin 2 of U18 should be a 0.01 µF disc capacitor. Also, the 40kohm resistor on the output of U18B should be 10k.

serial converter for 8-level teleprinters

The serial converter in August, 1977, *ham radio*, uses a 74121 for U16, not a 7474.

R1 (at minimum point of
$$X_{C1}$$
 curve) = $R_{1B} = \frac{2X_L^2}{R_2}$ (8)
Also, eq. 12 should read as follows:

$$X_{L} = (R1 + R2) \frac{Q_{o}^{2} + \sqrt{Q_{o}^{2} - (Q_{o}^{2} + 4) \left(\frac{R2 - R1}{R1 + R2}\right)^{2}}}{Q_{o}^{2} + 4}$$
(12)

audio frequency speech processing

The circuit board layout for the audio speech processor in August, 1977, *ham radio* was missing several connections. The diagram above shows the correct circuit board layout. The output is taken from the center of R13 and not as shown in **fig. 5** in the article. The numbering for the pins of the ICs in the schematic diagram should be changed to correspond with the 8-pin mini DIPs used on the finished board.

fig. 3	change to
5	3
4	2
6	4
10	6
11	7

phasing-type single-signal detector

In fig. 2, page 72 of October, 1976, ham radio, the two 180-ohm resistors should be connected between gate 2 and the source of the dual-gate mosfet as shown below. Also, gate number 1 is not connected to the source.



spectrum analyzer

There are several errors in the spectrum analyzer construction article which appeared in the June, 1977, issue. The 75.1 ohm resistor in the rf attenuator should be 71.5 ohms; the six 69.1 ohm resistors should be 61.9 ohms (fig. 10). The i-f attenuator should have three, *not two*, 20 dB sections (like the rf attenuator).

The mixer diodes used by the author are Hewlett-Packard part number 5082-2900; most any hotcarrier diodes should work if they are all the same type.

The crystal in the second local oscillator is 150 MHz ± 2 MHz; the crystal in the third local oscillator is 39.3 ± 1 MHz. The 10k resistor associated with CR401 should go to switch S601A, the 250 kHz position; the same for the 10k resistor associated with the second crystal filter, Y401 (fig. 11). The 2.4k resistor in series with CR402 should go to switch S601A, the 10 kHz position. The coil located near CR403, and the switch contacts near R402, are parts of the same relay.

Large size Xerox copies of the top and bottom chassis photographs are available from *ham radio*, and will be sent to interested readers upon receipt of a self-addressed, stamped envelope.

reducing IMD in highfrequency receivers

The 3-dB pad between the local oscillator input and the balanced mixer, in **fig. 6** on page 30 of the March, 1977, issue of *ham radio*, should have the values transposed (the series resistor should be 18 ohms, the shunt resistor 300 ohms.)

bandspreading techniques for resonant circuits

In eq. 19 on page 49 of the February, 1977, issue of ham radio, the term C_r should not be included under the radical sign. The equation should read:

$$C_p = \frac{\sqrt{C_q + C_r^2}}{2V} - C_r$$





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then be refined with the HP-25 program. My first step was to rewrite Sobol's equation as

$$Z_{o} = \frac{120\pi}{\sqrt{\epsilon_{r} \frac{w}{h}}} \left[\frac{1}{1 + 1.735 \epsilon_{r}^{-0.0724} w/h^{-0.836}} \right]$$

simple formula for microstrip impedance

In many amateur vhf and uhf applications strip transmission lines etched on printed-circuit board are used for impedance matching and as components in tuned resonant circuits. Although several methods are available for calculating the characteristic impedance of microstrip transmission line, the formula derived by Sobol¹ is the most popular. It has been widely publicized in Motorola Semiconductor's application notes and appeared recently in QST^2 . Sobol's equation:

$$Z_o = \frac{120\pi h}{\sqrt{\epsilon_r} \ w(1+1.735\epsilon_r^{-0.0724} \ w/h^{-0.836})}$$

where w is strip width, h is the dielectric thickness, and ϵ_r is the relative dielectric constant of the substrate.

Sobol's equation gives Z_o as a function of microstrip geometry, but in practical applications you usually need to know what size microstrip is required for a given impedance. Since the equation can't be solved directly for w/h, an interactive trial-and-error solution is necessary. This can be done rather quickly with a high-speed computer, but an iterative solution with a programmable calculator such as the HP-25 may require a minute or more — an iterative solu-

tion with a non-programmable calculator is impractical.

Some time ago N6TX (ex WA6UAM) sent me an iterative HP-25 program for Sobol's microstrip By inspection, to a first approximation Z_o is equal to the first term on the right-hand side of the equal sign; the term inside the parenthesis is a modification term which is a function of both ϵ_r and w/h. Designating the



fig. 1. Microstrip impedance calculated with simple formulas developed by W1HR (dashed lines), as compared to actual impedance (solid line). For $\epsilon_r > 4$, accuracy is very good for w/h>0.2.

equation which provided acceptable accuracy for most design work. This program begins at w/h = 1 and iterates out to the required value. Therefore, for low and high values of Z_o a solution requires considerable calculation time. To reduce calculation time I decided to see if I could develop a simple equation for an approximate value of w/h which could quantity $(1.735\epsilon_r^{-0.724}w/h^{-0.836})$ as *K*, eq. 1 was rewritten as

$$\frac{120\pi}{Z_o\sqrt{\epsilon_r}} = \frac{w}{h} (1+K) = \frac{w}{h} + K \cdot \frac{w}{h}$$

All that remained was to find a value for $K \cdot w/h$ which satisfied varying values of ϵ_r and w/h. After calculating several tables of values, it was ap-

H. Sobol, "Extending IC Technology to Microwave Equipment," *Electronics*, March 20, 1967, page 112.
 R. Olsen, N6NR, "Designing Solid-State RF Power Circuits," *QST*, September, 1977, page 15.

parent that $K \cdot w/h = 1$ would give the desired results. Substituting and rearranging terms yielded the expression

$$\frac{w}{h} \approx \frac{120\pi}{Z_o \sqrt{\epsilon_r}} - 1 \tag{3}$$

When this equation was plotted on graph paper and compared to a graph of Sobol's equation, the similarity was much closer than I expected the curve had essentially the correct shape, but all values were slightly larger than those given by Sobol's formula. This was the desired result; rewriting the HP-25 program around eq. 3 considerably reduced calculation time.

Later it occurred to me that it might be possible to further factor **eq. 3** to obtain a more accurate formula for microstrip impedance. After calculating numerous tables of Z_o vs w/h and ϵ_{τ} , and inspecting the values, I found that the impedance of microstrip etched on a substrate with $e_{\tau} > 4.0$ could be approximated within a few per cent by the following equations:

$$\frac{w}{h} \approx \frac{120\pi}{Z_o \sqrt{\epsilon_r + \sqrt{\epsilon_r}}} - 1$$

$$Z_o \approx \frac{120\pi}{(\frac{w}{h} + 1)\sqrt{\epsilon_r + \sqrt{\epsilon_r}}}$$
(5)

For microstrip etched on glass-epoxy circuit board ($\epsilon_r = 4.8$), these equations can be reduced to

$$\frac{w}{h} \approx \frac{142.6}{Z_o} - 1 \qquad Z_o \approx \frac{142.6}{\frac{w}{h} + 1}$$

- - - -

. . . .

For Teflon-fiberglass circuit board ($\epsilon_r = 2.55$) the simplified expressions are

$$\frac{w}{h} \approx \frac{185.1}{Z_o} - 1 \qquad Z_o \approx \frac{185.1}{\frac{w}{h} + 1}$$

The dielectric constant of Teflonfiberglass is below the value recommended for these equations, but accuracy is still acceptable for many applications. These formulas can be solved quickly by hand (or with a simple four-function calculator), and should be a big help to amateurs who want to design their own microstrip circuits. They can also be used to determine the approximate impedance of circuit traces for digital logic boards (for best results the V_{cc} and ground lines for TTL should have low impedance).

The accuracy of these simplified equations is surprisingly good. As shown in fig. 1, for w/h > 0.2, the simplified formulas are within a few per cent of the impedance calculated with more accurate equations; this covers the microstrip impedance range most commonly used in radio communications work. With fiberglass-epoxy board the formulas are within about 1 ohm of the exact expression for all values of Z_{0} below 60 ohms. The values for Teflonfiberglass board are somewhat less accurate, but are still acceptable for most amateur work.

James R. Fisk, W1HR

improved (vfo) stability for the Atlas 180

Early versions of the Atlas 180 transceiver have exhibited poor vfo stability with a varying dc supply voltage. In some cases, the vfo will actually be frequency modulated at dc input voltages below 13 volts. Atlas owners can check for this condition by listening to a signal or the calibrator beat note and adjusting the dc supply from about 11.5 volts to 14.5 volts. A 500-milliampere supply is more than ample to operate the receiver. The test can also be made in the car by first setting up the beat note with the engine off and then starting the engine. After a few moments the battery system will come up to full-charge voltage of 14.5 volts. Any change in pitch during this time indicates poor vfo power supply regulation. The units in which this is most likely to occur are those which use a 10-volt regulator circuit consisting of a transistor with a 10-volt Zener on the base.

The solution to the problem is to remove the 27-ohm decoupling resistor (R401 in my Atlas 180) on the vfo board (PC-400), and replace it with a 78L08ACP low-power 8-volt regulator. The wire that previously connected to the 10-volt bus is then reconnected to the 13-volt bus. After making this change, retuning is unnecessary for dc inputs of 11.5 volts to 14.5 volts, and there are no reports of frequency modulation when operating mobile without the engine running. There is no other noticeable change in the operation of the vfo due to the 8-volt rather than 10-volt supply.

Dave Sargent, K6KLO



fig. 1. Modification to the Atlas 180 vfo power supply to prevent any frequency modulation due to voltage changes. The 78L08ACP voltage regulator is used to prevent voltage changes. It is fed from the normal 12.6-volt dc supply.









The TS-520S combines all of the fine, field-proven characteristics of the original TS-520 together with many of the ideas and suggestions for improvement from amateurs worldwide.

FULL COVERAGE TRANSCEIVER

The TS-520S provides full coverage on all amateur bands from 1.8 to 29.7 MHz. Kenwood gives you 160 meter capability, WWV on 15.000 MHz., and an auxiliary band position for maximum flexibility. And with the addition of the TV-506 transverter, your TS-520S can cover 160 meters to 6 meters on SSB and CW.

DIGITAL DISPLAY DG-5 (option)

The Kenwood DG-5 provides easy, accurate readout of your operating frequency while transmitting *and* receiving.

OUTSTANDING RECEIVER SENSITIVITY AND MINIMUM CROSS MODULATION

The TS-520S incorporates a 3SK35 dual gate MOSFET for outstanding cross modulation and spurious response characteristics. The 3SK35 has a low noise figure (3.5 dB typ.) and high gain (18 dB typ.) for excellent sensitivity.

NEW IMPROVED SPEECH PROCESSOR

An audio compression amplifier gives you extra punch in the pile

ups and when the going gets rough.

VERNIER TUNING FOR FINAL PLATE CONTROL

A vernier tuning mechanism allows easy and accurate adjustment of the plate control during tune-up.

FINAL AMPLIFIER

The TS-520S is completely solid state except for the driver (12B-Y7A) and the final tubes. Rather than subsitute TV sweep tubes as final amplifier tubes in a state of the arc amateur transceiver, Kenwood has employed two husky S-2001A (equivalent to 6146B) tubes. These rugged, time-proven tubes are known for their long life and superb linearity.

HIGHLY EFFECTIVE NOISE BLANKER

An effective noise blanking cricuit developed by Kenwood that virtually eliminates ignition noise is built into the TS-520S.

RF ATTENUATOR

The TS-520S has a built-in 20 dB attentuator that can be activated by a push button swich conveniently located on the front panel.

PROVISION FOR

A special jack on the rear panel of the TS-520S provides receiver signals to an external receiver for increased station versitility. A switch on the rear panel determines the signal path ... the receiver in the TS-820 or any external receiver.

The VF0-520 remote VF0 matches the styling of the TS-520S and provides maximum operating flexibility on the band selected on your TS-520S.

The TS-520S is completely selfcontained with a rugged AC power supply built-in. The addition of the DS-1A DC-DC converter (optional) allows for mobile operation of the TS-520S.

The TS-520S has 2 convenient RCA phono jacks on the rear panel for PHONE PATCH IN and PHONE PATCH OUT.

The CW-520-500 Hz filter can be easily installed and will provide improved operation on CW.

The AGC circuit has 3 positions (OFF, FAST, SLOW) to enable the TS-520S to be operated in the optimum condition at all times whether operating CW or SSB.

The TS-520S retains all of the features of the original TS-520 that made it tops in its class: RIT control • 8-pole crystal filter • Built-in 25 KHz calibrator • Front panel carrier level control • Semibreak-in CW with sidetone • VOX/PTT/MOX • TUNE position for low power tune up • Built-in speaker • Built-in Cooling Fan • Provisions for 4 fixed frequency channels • Heater switch.

pecifications

Amateur Bands: 160-10 meters plus WWV (receive only) Modes: USB, LSB, CW Antenna Impedance: 50-75 Ohms Frequency Stability: Within ± 1 kHz during one hour after one minute of warm-up, and within 100 Hz during any 30 minute period thereafter Tubes & Semiconductors: Tubes (S2001A x 2, 128Y7A) Transistors 52 FFTs 19 Diodes. 101 Power Requirements: 120/220 V AC, 50/60 Hz, 13.8 V DC (with optional DS-IA) Power Consumption: Transmit 280 Watts Receive: 26 Watts (with heater off) Dimension: 333(13%) W x 153 (6-0) H x 335(13 (13-3/16) D mm(inch) Weight: 16.0 kg(35.2 lbs) TRANSMITTER RF Input Power: SSB: 200 Watts 'PEP CW: 160 Watts DC Carrier Suppression: Better than -40 dB Sideband Suppression: Better than -50 dB Spurious Radiation: Better than 40 dB Microphone Impedance: 50k Ohms AF Response: 400 to 2,600 Hz RECEIVER Sensitivity: 0.25 uV for 10 dB (S+N)/NSelectivity: SSB:2.4 kHz/-6 dB. 4.4 kHz/-60 dB Selectivity: CW: 0.5 kHz/-6 dB. 1.5 kHz/-60 dB (with optional CW-520 filter) Image Ratio: Better than 50 dB IF Rejection: Better than 50 dB AF Output Power: 10 Watt (8 Ohm load, with less than 10% distortion) AF Output Impedance: 4 to 16 Ohms DG-5 SPECIFICATIONS Measuring Range: 100 Hz to

40 MHz Input Impedance: 5 k Ohms Gate Time: 0.1 Sec. Input Sensitivity: 100 Hz to 40 MHz ... 200 mV rms or over, 10

kHz to 10 MHz., 50 mV or over Measuring Accuracy: Internal time base accuracy ±0.1 count Time Base; 10 MHz

Operating Temperature: -10° to 50° C/14° 122° F

Power Requirement: Supplied from TS-520S or 12 to 16 VDC (nominal 13.8 VDC) Dimensions: 167(6-9/16) W x 43(1-11/16) H x 268(10-9/16) D

mm(inch) Weight: 1.3 kg(2.9 lbs)



DG-5

The locally of digited reaction is evenled is on the TS.5208 by connecting the ploth readom (option) which that just the average readout circuit this counter induces the partial VFC and networking requencies to give you your and throughout. This handsometristy for accessory can be set almost any place in your shack for easy to read operation. In set it on the dashcoard during mobile operation for set all conventence. Six bold digits display your operating frequency while you transmit and takenty. Complete with DH (display hold) switch for nequency memory and 2 position intensity selection. The DG 5 cen also be used as a normal frequency counter up to 40 with a the rough of a switch (input cable provided.)

NOTE: TS-520 owners can use the DG-5 with a DK-520 adaptar kit.





We told you that the TS-820 would be best. In little more than a year our promise has become a fact. Now, in response to bundreds of requests from PLL • The TS-820S employs

PLL • The TS-820S employs the latest phase lock loop circuitry. The single conversion receiver section performance offers superb protection against unwanted cross-modulation. And now PLL allows the frequency to remain the same when switching sidebands (USB, LSB, CW) and eliminates having to recalibrate each time.

DIGITAL READOUT • The digital counter display is employed as an integral part of the VFO readout system. Counter mixes the carrier VFO, and first heterodyne frequencies to give *exact* frequency. Figures the frequency down to 10 Hz and digital display receive and transmit frequencies are displayed in easy to read, Kenwood Blue digits. SPEECH PROCESSOR • An RF circuit provides quick time constant compression using a true RF compressor as opposed to an AF clipper. Amount of compression is adjustable to the desired level by a convenient front panel control.

IF SHIFT • The IF SHIFT control varies the IF passband without changing the receive frequency. Enables the operator to eliminate unwanted signals by moving them out of the passband of the receiver. This feature alone makes the TS-820S a pacesetter.

The TS-820 and DG-1 are still available separately.

best. In little more than a year our promise has become a fact. Now, in response to hundreds of requests from amateurs, Kenwood offers the TS-820S"... the same superb transceiver, but with the digital readout factory installed. As an owner of this beautiful rig, you will have at your fingertips the combination of controls and features that even under the toughest operating conditions make the TS-820S the Pacesetter that it is.



Experience the excitement of 6 meters. The TS-600 all mode transceiver lets you experience the fun of 6 meter band openings. This 10 watt, solid state rig covers 50.0-54.0 MHz. The VFO tunes the band in 1 MHz segments. It also has provisions for fixed frequency operation on NETS or to listen for beacons. State of the art features such as an effective noise blanker and the RIT (Receiver Incremental Tuning) circuit make the TS-600 another Kenwood "Pacesetter".



An easy way to get on the 6 meter band with your TS-520/ 520S, TS-820/820S and most other transceivers. Simply plug it in and you're on ... full band coverage with 10 watts output on SSB and CW.



Experience the luxury of 450 MHz at an economical price.

The TR-8300 offers high quality and superb performance as a result of many years of improving VHF/ UHF design techniques. The transceiver is capable of F₃ emission on 23 crystal-controlled channels (3 supplied). The transmitter output is 10 watts.

The TR-8300 incorporates a 5 section helical resonator and a

two-pole crystal filter in the IF section of the receiver for improved intermodulation characteristics. Receiver sensitivity, spurious response, and temperature characteristics are excellent.





Check out the new "built-ins": digital readout, receiver pre-amp, VOX, semi-break in, and CW sidetonel Of course, it's still all mode, 144-148 MHz and VFO controlled. Features: Digital readout with "Kenwood Blue" digits • High gain receiver pre-amp • 1 watt lower power switch • Built in VOX • Semi-break in on CW • CW sidetone • Operates all modes: SSB (upper & lower), FM, AM and CW • Completely solid state circuitry provides stable, long lasting, trouble-free operation • AC and DC capability (operate from your car, boat, or as a base station through its built-in power supply) • 4 MHz band coverage (144 to 148 MHz) • Automatically switches transmit frequency 600 KHz for repeater operation. Simply dial in your receive frequency and the radio does the rest... simplex, repeater, reverse • Or accomplish the same by plugging a single crystal into one of the 11 crystal positions for your favorite channel • Transmit/Receive capability on 44 channels with 11 crystals.



Handsomely styled and a perfect companion to the TS-700S. This unit provides you with the extra versatility and the luxury of having a second VFO in your shack. Great for split frequency operation and for tuning off frequency to check the band. The function switch on the VFO-700S selects the VFO in use and the appropriate frequency is displayed on the digital readout in the TS-700S. In addition a momentary contact "frequency check" switch allows you to spot check the frequency of the VFO not in use.





TR-7400A

Features Kenwood's unique Continuous Tone Coded Squelch system, 4 MHz band coverage, 25 watt output and fully synthesized 800 channel operation. This compact package gives you the kind of performance specifications you've always wanted in a 2-meter amateur rig.

Outstanding sensitivity, large-sized helical resonators with High Q to minimize undesirable out-of-band interferance, and give a 2-pole 10.7 MHz monolithic crystal filter combine to give your TR-7400A outstanding receiver performance. Intermodulation characteristics (Better than 66dB), spurious (Better than ~60dB), image rejection (Better than ~70dB), and a versatile squelch system make the TR-7400A tops in its class. Shown with the PS-8 power supply

(Active filters and Tone Burst Modules optional)



This 100 channel PLL synthesized 146-148 MHz transceiver comes with 88 pre-programmed channels for use on all standard repeater frequencies (as per ARRL Band Plan) and most simplex channels. For added flexibility, there are 6 diode-programmable switch positions. The 15 KHz shift function makes these 6 positions into 12 channels. 10 watt output, ± 600 KHz offset and LED digital frequency display are just a few of the many fine features of the TR-7500. The PS-6 is the handsomely styled, matching power supply for the TR-7500. Its 3.5 amp current capacity and built-in speaker make it the perfect companion for home use of the TR-7500.

The high performance portable 2-meter FM transceiver. 146-148 MHz, 12 channels (6 supplied), 2 watts or 400 mW RF output. Everything you need is included: Ni-Cad battery pack, charger, carrying case and microphone.



Kenwood developed the T-599D transmitter and R-599D receiver for the most discriminating amateur.

The R-599D is the most complete receiver ever offered. It is entirely solid-state, superbly reliable and compact. It covers the full amateur band, 10 through 160 meters, CW, LSB, USB, AM and FM.

The T-599D is solid-state with the exception of only three tubes, has built-in power supply and full metering. It operates CW, LSB, USB and AM and, of course, is a perfect match to the R-599D receiver.

If you have never considered the advantages of operating a receiver/transmitter combinationmaybe you should. Because of the larger number of controls and dual VFOs the combination offers flexibility impossible to duplicate with a transceiver.

Compare the specs of the R-599D and the T-599D with any other brand. Remember, the R-599D is all solid state (and includes four filters). Your choice will obviously be the Kenwood.





Dependable operation, superior specifications and excellent features make the R-300 an unexcelled value for the shortwave listener. It offers full band coverage with a frequency range of 170 KHz to 30.0 MHz • Receives AM, SSB and CW • Features large, easy to read drum dials with fast smooth dial action • Band spread is calibrated for the 10 foreign broadcast bands, easily tuned with the use of a built-in 500 KHz calibrator • Automatic noise limiter • 3-way power supply system (AC/Batteries/External DC) ... take it anyplace • Automatically switches to battery power in the event of AC power failure.



Fine equipment that belongs in every well equipped station

OZU Jeries	
TS-820S	TS-820 with Digital Installed
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DG-1	Digital Frequency Display
VFO-820	Deluxe Remote VFO for for TS-820 / 820S
CW-820	500 Hz CW Filter for TS-820/820S
DS-1A	DC-DC Converter for 520/820 Series
520 Series	0207020 001100
TS-520S	160-10 M Transceiver
DG-5	Digital Frequency Display for TS-520 Series
VFO-520.	Remote VFO for TS-520 and TS-520S
SP-520	External Speaker for 520/820 Series
CW-520	500 Hz CW Filter for TS-520/520S
DK-520	Digital Adaptor Kit for TS-520
599D Serie	s
R-599D	160-10 M Solid State Receiver
T-599D	80-10 M Matching Transmitter
S-599	External Speaker for 5990 Series

CC-29A	2 Meter	Converter	tor
	R-599D		
CC-69	6 Meter R-599D	Converter	for
M-599A	FM Filte	r for R-599	D

smort wave listening

R-300 General Coverage SWL Receiver

VHE LINES

TS-600	6 M All Mode Transceiver
TS-700S	.2 M All Mode Digital Transceiver
VFO-700S .	Remote VFO for TS-700S
SP-70	Matching Speaker for TS-600 / 700 Series
TR-2200A.	2 M Portable FM Transceiver
TR-7400A	2 M Synthesized Deluxe FM Transceiver

MORE ACCESSORIES:

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RA-1 T90-0082-05 PB-15 E07-0403-05 See Service Manual Specify Model Specify Model TR-2200A TR-2200A TR-2200A All Models TR-7400A TS-700A; TR-7400A All Models All Models

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TR-8300

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MC-50....

PS-5.....

PS-8.....

for all Kenwood models.

For use with

PS-6



The Kenwood HS-4 headphone set adds versatility to any Kenwood station. For extended periods of wear, the HS-4 is comfortably padded and is completely adjustable. The frequency response of the HS-4 is tailored specifically for amateur communication use. (300 to 3000 Hz, 8 ohms).



Model #

The MC-50 dynamic microphone has been designed expressly for amateur radio operation as a splendid addition to any Kenwood shack. Complete with PTT and LOCK switches, and a microphone plug for instant hook-up to any Kenwood rig. Easily converted to high or low impedance. (600 or 50k ohm).

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For literature on any of the new products, use our *Check-Off* service on page 150.

crystal filters

Sherwood Engineering has announced two new additions to their crystal filter line. As complements to the CF-600/6, the new CF-2.6K/8 or CF-2.3K/8 crystal filter sets will replace the normal 8-kHz wide first i-f filter in the Drake R-4C. Each set has two filters, USB and LSB, that must be switched for the correct sideband. The individual filters are 8-pole crystal-ladder filters.

The CF-2.6K/8 is a set of ssbbandwidth filters that are approximately 200 Hz wider than the normal second i-f phone filter. This allows a limited amount of passband tuning, while still reducing the second i-f bandwidth from 32 kHz, at -60 dB, to approximately 4 kHz. The other phone filter pair is the CF-2.3K/8, which is slightly narrower (100 Hz nominally) than the second i-f filter. Having the new filter sharper than the normal filter produces the equivalent of a 2 to 2.1 kHz filter, with 16 poles distributed over two frequencies. The passband tuning is then used to align the center frequencies, of the two filters, for proper cascading. This narrow combination offers the ultimate in phone selectivity. The bandwidth using the CF-2.6K/8, with the normal phone filter, is 2.3 kHz, at -6 dB, and 3.1 kHz at -60 dB; the bandwidth for the CF-2,3K/8 is 2.1 kHz and 2.9 kHz, at the 6 and 60 dB points. The additional advantages gained by distributing selectivity over two i-f frequencies are: virtual elimination of the chance of overloading the second mixer, and elimination of off-frequency signals that leak around the normal second i-f filter.

In addition to offering the basic filters, Sherwood Engineering also sells switching kits for the first i-f filters. The simplest arrangement is for the operator who wants to switch only between the two ssb bandwidth filters (CF-2.3K/8 or CF-2.6K/8). Custom-designed kits are also available to permit switching of all first i-f filters, 8 kHz, 2.6/2.3 kHz, or 600 Hz. Prices for the new filters are \$120. The basic switching kit is \$29.00 with the cost increasing approximately \$25.00 per additional filter switched. Exact price quotes are given based on an individual's needs. For more information, contact Sherwood Engineering, Incorporated, 1268 South Ogden Street, Denver, Colorado 80210.

two-meter preamplifier



A new two-meter preamp has been introduced by Janel Labs. This preamp is specially designed to improve the sensitivity of transceivers and includes bypass circuitry for carrying transmit power through the unit. The preamp has a low noise figure, which gives excellent sensitivity for weak signals. An adjustable delay circuit (similar to that used in VOX circuits) allows for its use on all modes — f-m, ssb, am and CW.

The gain of the QSA 5 has been optimized for transceivers. It has a 15-dB gain level, which is sufficient to improve the sensitivity as much as practical but low enough to avoid creating overload problems.

A front-panel switch on the QSA 5

disables the preamp from the antenna line. This switch allows you to reduce gain on local signals and also allows experimentation on weak signals. A LED pilot light indicates when the preamp is in the line. This same LED also indicates when transmit power is being sensed.

The QSA 5 preamp is available from Janel Laboratories, 3312 S.E. Van Buren Blvd., Corvallis, Oregon 97330. The QSA 5 is available from stock at \$39.95 plus postage. A full one-year warranty is provided. Specifications are available upon request.

multiband antenna coils (40 through 10 meters)



Microwave Filter Company announces a set of antenna coils that will convert an amateur antenna from a single-frequency band of limited operation to operation on all amateur hf bands (40-10 meters).

Known as Reyco antenna coils, they are designed to shorten the overall physical length of an original single-frequency-band antenna. Model numbers are KW-40, 20, 15, and 10. Used in pairs, the model KW-40 coils will give flexibility of operation on all five hf amateur bands. Ideal performance is obtained by using all four coil pairs (KW-40 through KW-10).

In today's crowded apartment and suburban communities, the shortened antenna using Reyco multiband coils provides flexibility in minimum space. For additional information, write Microwave Filter Company, 6743 Kinne Street, East Syracuse, New York 13057.

All in the family.

Feather Touch Keyer \$69.95



No moving parts! The **Kantronics Feather Touch Keyer** responds to the lightest touch. No more slapping or sloshing! No moving parts also means the end of adjusting and readjusting before each QSO.

The **Feather Touch** sends self completing dots and dashes, adjustable from 7½ WPM, and gives you a great fist on the air. Attractive design and compact size make the **Feather Touch** a professional addition to the sharpest ham station. Design features keep the keyer from creeping away as you send.

This **battery powered** unit is great for portable use or home operation with the aid of any DC power supply from 5-15 volts. Pick up a motionless keyer today!



Notcher CW Filter \$34.95

Make your CW receiver selectivity razor sharp with the **Kantronics Notcher Audio CW Filter**. This filter makes sense out of the biggest pileups! The **Notcher** funnels down to 150 Hz @ -3dB to separate signals that appeared to be on top of each other before.

Your **Notcher** will operate portable with a 9 volt internal battery, or from your 5-15 volt DC power supply.

Designed to look sharp too, the Notcher is one in a growing family of Kantronics quality products. Our quality is more than skin deep. One look inside will tell you the Notcher is built to perform! The Standard Frequency Calibrator





Kantronics frequency calibrator is The Standard. Advanced CMOS circuitry checks your frequency with crystal controlled accuracy. Zero-beat your transceiver to The Standard at 50 KHz intervals.

No direct connections are needed, the unit transmits to your receiver. Internal jumpers adjust **The Standard** for a choice of 25 KHz, 50 KHz or 100 KHz intervals.

Powered by battery for portable operation, or 5-15 volt DC power supply. The Standard is a handsome station accessory that looks sharp, inside and out.

Be confident of your frequency.





Magnetic Mount



The Kantronics Mobile 2 Antenna offers a reasonable alternative

to the high priced VHF antenna! The **Mobile 2** is a high-quality, quarterwavelength antenna that is quickly installed.

Choose between **magnetic or trunk** mounting bases. Both include 18 feet of RG-58/U coax cable and standard PL-259 connector. Specify 147 MHz or 220 MHz whip and coil assembly. All these features . . . for a low, low price!

Trunk Mount





Features:

Custom computer grade commercial components, capacitors, and tube sockets manufactured especially for high power use—heavy duty 10Kw silver plated ceramic band switches • Silver plated copper tubing tank coil • Huge 4" easy to read meters-measure plate current, high voltage, grid current, and relative RF output • Continuous duty power supply built in • State of the art zener diode standby and operating bias provides reduced idling current and greater output efficiency • Built in hum free DC heavy duty antenna change-over relays • AC input 110V or 220V AC. 50-60Hz . Tuned input circuits . ALC-rear panel connections for ALC output to exciter and for relay control . Double internal shielding of all RF enclosures . Heavy duty chassis and cabinet construction and much, much more.

- Full band coverage 160-10 meters including mars.
 2000 watts P.E.P. SSB input. 1000 watts input continuous duty, CW, RTTY & SSTV.
- Two Eimac 3-500Z conservatively rated finals.
- · All major HV and other circuit components mounted on single G-10 glass plug in board. Have a service problem? (Very unlikely) Just unplug board and send to us
- · Heavy duty commercial grade quality and construction second to no other unit at any price! • Weight: 90 lbs. Size: 91/2" (h) x 16" (w) x 153/4" (d)



Fully Certifiable for Commercial Use Features:

Extremely stable local oscillator for easy measurement of HF, VHF, and UHF bands Extremely state incar occurate incursor insure extremely high stability • Easy to read, accurate linear scale • Direct off the air signal measurement capability.



Specifications:

Frequency: 1.8MHZ-520MHZ/3 range select (A, B, C, EXT), A range: 26.5 MHZ-40MHZ, B range: 48MHZ-60MHZ, C range: 140MHZ-156MHZ, EXT. range: 1.8MHZ-520MHZ (Need Signal Generator) - Generous overranges - Input level: (1) Through type input level: IW-200W (RF Input Ter-

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All Solid State-CMOS PLL digital synthesized - No Crystals to Buy! 5KHz steps -144 - 149 MHz-LED digital readout PLUS MARS-CAP.*

● 5 MHz Band Coverage - 1000 Channels (instead of the usual 2MHz to 4MHz-400 to 800 Channels) ● 4 CHANNEL RAM IC MEMORY WITH SCANNING ● MULTIPLE FREQUENCY OFFSETS ● ELECTRONIC AUTO TUNING - TRANSMIT AND RECEIVE ● INTERNAL MULTIPURPOSE TONE OSCILLATOR ● RIT ● DISCRIMINATOR METER - 15 Watts Output - Unequaled Receiver Sensitivity and Selectivity - 15 POLE FILTER, MONOLITHIC CRYSTAL FILTER AND AUTOMATIC TUNED RECEIVER FRONT END, COMPARE! ● Superb Engineering and Superior Commercial Avionics Grade Quality and Construction Second to None at ANY PRICE.

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- FREQUENCY RANGE: Receive and Transmit: 144.00 to 148.995 MHz, 5Khz steps (1000 channels) INCLUDING NEW BAND 144.5-145.5MHz + MARS-CAP.*
- . LED DIGITAL READOUT.
- 4 CHANNEL RAM SCANNER WITH IC MEMORY: Program any 4 frequencies and reprogram at any time using the front panel controls-scan all or part of the memory-search for occupied (closed) channel or vacant (open) channels. Internal Ni-Cad included to retain memory (no diode matrix to wire or change).
- MULTIPLE FREQUENCY OFFSETS: Three positions A,B,C, provided for installation of optional crystals: EXAMPLE - 1 MHz offset. Duplex Frequency Offset Built in - 600 Khz PLUS or MINUS 5 KHz steps, plus simplex, any frequency.
- INTERNAL MULTIPURPOSE TONE OSCILLATOR BUILT IN: 1750Hz tone burst for "whistle on operation" and sub-audible tone operation possible by simply adding a capacitor across the terminals provided. Internal 2 position switch for automatic and manual operation, tone burst or sub audible tone PL - adjustable 60-203Hz (100 Hz provided).
- AIRCRAFT TYPE FREQUENCY SELECTOR: Large and small coaxially mounted knobs select 100KHz and 10KHz steps respectively. Switches click-stopped with a home position facilitate frequency changing without need to view LED's while driving and provides the sightless amateur with full Braille dial as standard equipment.
- FULL AUTOMATIC TUNING OF RECEIVER FRONT END AND TRANSMITTER CIRCUITS: DC output of PLL fed to varactor diodes in all front end RF tuned circuits provides full sensitivity and optimum intermodulation rejection over the entire band. APC (AUTO POWER CONTROL) - Keeps RF output constant from band edge to band edge. NO OTHER AMATEUR UNIT AT ANY PRICE has these

features which are found in only the most sophisticated and expensive aircraft and commercial transceivers.

TRUE FM: Not phase modulation - for superb emphasized hi-fi audio quality second to none.
 RIT CONTROL: Used to improve clarity when contacting stations

CM-00150

- MONITOR LAMPS: 2 LED's on front panel indicate (1) incoming
- signal-channel busy, and (2) Transmit. • FULLY REGULATED INTEGRAL POWER SUPPLY: Operating voltage for all 9v circuits independently regulated. Massive
- voltage for all 9v circuits independently regulated. Massive Commercial Hash Filter.
- MODULAR COMMERCIAL GRADE CONSTRUCTION: 6 Unitized modules eliminate stray coupling and facilitate ease of maintenance.
- ACCESSORY SOCKET: Fully wired for touch tone, phone patch, and other accessories. Internal switch connects receiver output to internal speaker when connector is not in use.
- MULTI-PURPOSE METER: Triple Function Meter Provides Discriminator Meter, "S" Reading on receive and Power Out on Transmit.
- RECEIVE: Better than .25uv sensitivity, 15 POLE FILTER as well as monolithic crystal filter and AUTOMATIC TUNED LC circuits provide superior skirt selectivity - COMPAREI
 HIGH/LOW POWER OUTPUT: 15 watts and 1 watt, switch
- HIGH/LOW POWER OUTPUT: 15 watts and 1 watt, switch selected. Low power may be adjusted anywhere between 1 and 15 watts. Fully protected-short or open SWR.
- OTHER FEATURES: Dynamic Microphone, Built In Speaker, mobile mount, external 5 pin accessory jack, speaker jack, and much, much more. Size 2½ x 7 x 7½. All cords, plugs, fuses, microphone hanger, etc. included. Weight 5 lbs.

Manufactured by one of the world's most distinguished Avionics manufacturers, Kyokuto Denshi Kaisha, Ltd. First in the world with an all solid state 2 meter FM transceiver.



More Details? CHECK-OFF Page 150

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QUALITY KENWOOD TRANSCEIVERS

The TS-820 is the rig that is the talk of the Ham Bands. Too many built-in features to list here. What a rig and only \$830.00 ppd. in U.S.A. Many accessories are also available to increase your operating pleasure and station versatility.



TS-700A 2M TRANSCEIVER

Guess which transceiver has made the Kenwood name near and dear to Amateur operators, probably more than any other piece of equipment? That's right, the TS-520. Reliability is the name of this rig in capital letters. 80 thru 10 meters with many, many builtin features for only \$629.00 ppd. in U.S.A.



TR-7400A 2M MOBILE TRANSCEIVER

Send SASE NOW for detailed info on these systems as well as on many other fine lines. Or, better still, visit our store Monday thru Friday from 8:00 a.m. thru 5:00 p.m. The Amateurs at Klaus Radio are here to assist you in the selection of the optimum unit to fullfill your needs.





160-10M TRANSCEIVER

Super 2-meter operating capability is yours with this ultimate design. Operates all modes: SSB (upper & lower), FM, AM and CW. 4 MHz coverage (144 to 148 MHz). The combination of this unit's many exciting features with the quality & reliability that is inherent in Kenwood equipment is yours for only \$599.00 ppd. in U.S.A.

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80-10M TRANSCEIVER

This brand new mobile transceiver (TR-7400A) with the astonishing price tag is causing quite a commotion. Two meters with 25W or 10W output (selectable), digital read-out, 144 through 148 MHz and 800 channels are some of the features that make this such a great buy at \$399.00 ppd. in U.S.A.



Other features include Zener regulated meter circuitry, adjustable brake delay, and handsome up-to-date styling compatible to most Ham gear. Cabinet measures 6"X7½"X7½"

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All band operation (160-10 meters) with any random length of wire. 200 watt **output** power capability—will work with virtually any transceiver. Ideal for portable or home operation. Great for apartments and hotel rooms—simply run a wire inside, out a window, or anyplace available. Toroid inductor for small size: 4-1/4" X 2-3/8" X 3." Built-in neon tune-up indicator. SO-239 connector. Attractive bronze finished enclosure.

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Increases usable bandwidth of any antenna. Tunes out SWR on mobile whips from inside your car.

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HP205AG Lab audio gen .02 20kHz 195
HP212A Pulse gen .06 5kHzPRR
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The Touch. It's the best value available in scanners.

Searching Receiver

Touch SP, then enter the starting frequency of your choice. The Touch will search up through the action radio channels in the search band until it hears an active call. You'll probably discover "live" frequencies you never before knew existed.

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Touch 2., then sit back. Any call coming in over the frequency you choose for channel one will automatically override calls on other channels. You'll never miss a call on your favorite frequency.

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Touch SS to Search the unknown. Touch SC to scan the known. You can either search through all bands for unknown frequencies, or listen to the stored frequencies you've selected for the sixteen scanning channels. There's so much versatility, and it's all at the tip of your finger.

Model	ACT-	F-16K

Selectivity

D	1000	
n	ange.	
2.2	30-50	MHz
	146-174	1MHz
	440-512	2MHz
	Ra	Range:

Sensitivity

(20 DB quieting) Lo VHF 0.5 μ V Hi VHF 0.6 μ V UHF 0.7 μ V ± 15 KHz (max.) @ 60 DB Squelch: (threshold) Lo VHF 0.4 μ V

Lo VHFΟ 4 μ V Hi VHFΟ.5 μ V UHFΟ.6 μ V

± 7 KHz (min.) @ 6 DB

Search Scan Range: (max) Lo VHF 4000 channels Hi VHF 5600 channels UHF 5760 channels

Scanning Receiver

Touch PR, then enter the frequency you want as you watch it appear on the L.E.D. display. Next, touch the channel number you wish to use. Then touch SC, the scanning lights will begin the search for action.

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P SS C

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delight operating 20 meters on a full 26' boom with 4 elements, 4 operational elements on 20-15-10, plus separate reflector element on 10 meters for correct monoband spacing. Featured are the large diameter High-Q Traps, Beta matching system, heavy duty Taper Swaged Elements, rugged Boom to Element mounting . . . and value priced at \$259.95. Additional features: • 10 dB Gain • 20-25 dB Front-to-Back Ratio • SWR less than 1.5 to 1 on all bands.

MODEL SY-1 SPECIFICATIONS:

Matching Method: Band MHz: Maximum Power Input: Legal Limit Gain VSWR (at Resonance) Impedance

Reta 14-21-28 10 dB 1.5 to 1

F/B Ratio 20-25 dB **Boom Length** 26 (2" O.D.) No. of Elements Longest Element 26' 7" **Turning Radius**

2" O.D. Mast Diameter Boom Diameter 2" O.D. 7.3 sq. ft. Surface Area 146 lbs. Windload Area Shipping Weight 50 lbs.



20 METERS





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HEAVY DUTY BOOM TO ELEMENT EXTRUSION





Sift ideas	S F.POPQ	The engineering
R X 28C 28-35 MHz FM receiver with pole 10.7 MHz crystal filter. R X 28C W/T same as above -wired & testec R X 50C Kit 30-60 MHz revr w/2 pole 10.7 MHz crystal filter. R X 50C W/T same as above -wired & testec R X 50C W/T same as above -wired & testec R X 144C Kit 140-170 MHz revr w/2 pole 10.7 MHz crystal filter. R X 144C W/T same as above -wired & tested R X 220C Kit. 10.7 MHz crystal filter. R X 220C W/T same as above -wired & tested R X 432C Kit. 432 MHz revr w/2 pole 10.7 MHz crystal filter. R X 432C Kit. same as above -wired & tested R X 432C Kit. same as above -wired & tested R X 432C W/T same as above -wired & tested	2 	RXCF accessory filter for above receiver kits gives 70 dB adjacent channel rejection 8.50 RF28 Kit 10 mtr RF front end 10.7 MHz out 12.50 8.50 RF50 Kit 6 mtr RF front end 10.7 MHz out 12.50 12.50 RF144D Kit 2 mtr RF front end 10.7 MHz out 17.50 17.50 RF432 Kit 432 MHz RF front end 10.7 MHz out 17.50 17.50 RF432 Kit 432 MHz RF front end 10.7 MHz out 17.50 10.7 MHz 17 module includes 2 IF 10.7F Kit 10.7 MHz 1F module includes 2 27.50 FM455 Kit 455 KHz IF stage plus FM detector 17.50 15.00
TX50 transmitter exciter, 1 watt, 6 TX50 W/T same as above-wired & testec TX144B Kit. transmitter exciter -1 watt-2 TX144B W/T. same as above-wired & testec TX220B Kit. transmitter exciter-1watt-2 MHz	mtr. 39.95 1 59.95 mtrs 29.95 20 29.95	TX 220B W/T · same as above – wired & tested 49.95 TX 432B Kit · transmitter exciter 432 MHz ·
 PA2501H Kit. 2 mtr power amp-kit 1 w in- out with solid state switching case, connectors PA2501H W/T. Same as above-wired & tester PA4010H Kit. 2 mtr power amp-10w in-44 out-relay switching PA4010H W/T. Same as above-wired & tester 6 mtr power amp-1 w in-15w out-less case, connectors & switching PA144/15 Kit. 2 mtr power amp-1 w in-15w out-less case, connectors & switching PA144/25 Kit. same as PA144/15 kit but 25° PA220/15 Kit. PA144/25 Kit. similar to PA144/15 kit but 25° PA432/10 Kit. PA140/10 W/T Nu in-140w out-2 mtr amp PA140/30 W/T Swin -140w out-2 mtr amp 	25w POWER AMPLIFIE 59.95 4.74.95 59.95 4.74.95 Yout, 169.95 W 39.95 W 39.95 MHz 39.95 MHz 39.95 D159.95	Blue Line RF power amp, wired & tested, emission- CW-FM-SSB/AM Model BAND Input Power Output BLC 10/70 144 MHz 10W 70W 139.95 BLC 2/70 144 MHz 2W 70W 159.95 BLC 10/150 144 MHz 10W 150W 259.95 BLC 30/150 144 MHz 30W 150W 239.95 BLC 30/150 144 MHz 2W 60W 159.95 BLC 10/160 220 MHz 10W 60W 159.95 BLD 10/60 220 MHz 10W 60W 139.95 BLD 10/120 220 MHz 10W 40W 139.95 BLE 2/40 420 MHz 10W 40W 139.95 BLE 2/40 420 MHz 10W 80W 259.95 BLE 30/80 420 MHz 10W 80W 259.95 BLE 10/80 420 MHz 10W 80W 289.95
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RPT50 Kit. repeater - 6 meter. RPT50. repeater - 6 meter. wired & te RPT144 Kit. repeater - 2 mtr - 15w - compl (less crystals) repeater - 220 MHz - 15w - compl RPT220 Kit. repeater - 10 watt - 432 MHz (less crystals) repeater - 10 watt - 432 MHz RPT144 W/T repeater - 15 watt - 2 mtr. RPT220 W/T repeater - 15 watt - 2 20 MHz RPT32 W/T repeater - 15 watt - 432 MHz DPLA50 6 mtr close spaced duplexer	465.95 sted 695.95 ete 	DPLA1442 mtr, 600 KHz spaced duplexer, wired and tuned to frequency379.95DPLA220220 MHz duplexer, wired and tuned to frequency379.95DPLA432rack mount duplexer319.95DSC-Udouble shielded duplexer cables with PL259 connectors (pr.)25.00DSC-Nsame as above with type N connectors (pr.)25.00
TRX50 Kit Complete 6 mtr FM transceiv 20w out, 10 channel scan wit (less mike and crystals). TRX144 Kit same as above, but 2 mtr & 15 TRX220 Kit Same as above except for 220 TRX432 Kit same as above except for 220 Same as above except for 220 TRX432 Kit TRC-1 transceiver case only transceiver case and accessorie	er kit, h case 229.95 w out219.95 and 254.95 19.95 es 39.95 TRANSCEIVER:	S OTHER PRODUCTS BY VHF ENGINEERING CD1 Kit 10 channel receive xtal deck w/diode switching
SYN II Kit. 2 mtr synthesizer, transmitt o programmable from 100 KH2 (Mars offsets with optional adapters) SYN II W/T . Same as above-wired & testee MO-1 Kit. 18 MH2 optional tripler	offsets SYNTHESIZERS - 10 MHz, - - 169.95 - - 239.95 - - 2.50 -	S for RX with priority 19.95 Crystals we stock most repeater and simplex pairs from 146.0-147.0 (each) 5.00 CWID Kit 159 bit, field programmable, code identifier with built-in squelch tail and ID timers 39.95 CWID wired and tested, not programmed 54.95 CWID wired and tested, not programmed 59.95 MIC 1 2.000 ohm dynamic mike with 12.95
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nd the angel said unto her, Fear not, Mary: for thou hast found favor with God. And, behold, thou shalt conceive in thy womb, and bring forth a son, and shalt call His name Jesus. Luke 1: 30,31 KJV

Foy to the World...

Then said Mary unto the angel, How shall this be, seeing I know not a man? And the angel answered and said unto her, The Holy Ghost shall come upon thee, and the power of the Highest shall overshadow thee; therefore also that holy thing which shall be born of thee shall be called the Son of God. Luke 1: 34,35 KJV

And she brought forth her firstborn son, and wrapped him in swaddling clothes, and laid him in a manger; because there was no room for them in the inn. Luke 2:7 KJV

For God so loved the world, that he gave his only begotten Son, that whosoever believeth in him should not perish, but have everlasting life. John 3:16 KJV

Nearly 2000 years ago, God reached out and touched the world with his love. He is still reaching out today seeking the lost.

Let his love reach into your heart this Christmas season. Become a member of God's family by turning from your way, and trusting in God's way today.

May you and your family reach out and accept the joy he has for you this Christmas.

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THE EXPERIMENTER OR DESIGNER SPECIFICATIONS: OUTPUT VOLTAGES: +5V, +12V, -12V; USABLE CURRENT: 750mA; % Regulation at 500mA: 0.2%; Short-circuit limited at 1.0 amp; Thermal overload protected. Power requirements: 117VAC, 60HZ, 40 Watts. Function Generator: Frequency range: 1HZ to 100 HZ in 5 bands. Amplitude adjustable from 0 to 10 VPP. DC offset adjustable from 0 to \pm 10V. Waveforms: Sine, square, triangular and TTL Clock. TTL Clock to t + 5V level, 200 ns rise and fall time, Frequency determined by Function Generator. Output impedance 1.2K ohm. Most of all, it's easy to construct and service. PC boards are predrilled, plated thru and solder flowed. Over 1000 units sold to schools.

sold to schools

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MAL POINT, ZERO SUPPRESSION UPON DEMAND. FEATURES TWO INPUTS: ONE FOR LOW FREQUENCY INPUT, AND ONE ON PANEL FOR USE WITH ANY INTERNALLY MOUNTED HAL-TRONIX PRE-SCALER FOR WHICH PROVISIONS HAVE ALREADY BEEN MADE. 1.0 SEC AND .1 SEC TIME GATES. ACCURACY ± .001%. UTILIZES 10-MHz CRYSTAL 5 PPM. COMPLETE KIT \$124.00

HAL-TRONIX BASIC COUNTER KITS STILL AVAILABLE

THE FOLLOWING MATERIAL DOES NOT COME WITH THE BASIC KIT: THE CABINET, TRANSFORMER, SWITCHES, COAX FITTINGS, FILTER LENS, FUSE HOLDER, T-03 SOCKET, POWER CORD AND MOUNT-ING HARDWARE.

(Same	Specifications as HAL-600A)	\$124.00
(Same	Specifications as HAL-300A)	\$99.00
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PRE-SCALER KITS

HAL-0-300PRE (Pre-drilled G10 board and all components) \$19.95 HAL-0-300P/A (Same as above but with preamp) \$29.95 HAL-0-600PRE (Pre-drilled G10 board and all com-\$39.95 ponents) HAL-1GHZ (New Item - Available in December) \$124.95

PRE-BUILT COUNTERS AVAILABLE (HAL-600A - \$229.00) (HAL-300A - \$199.00) HAL-

- \$199.00). ALLOW 4- TO 6-WEEK DELIVERY 50A -ON PRE-BUILT UNITS.



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PREAMPS

HIGH GAIN . LOW NOISE 30 dB power gain, 2.5-3.0 dB N.F. at 150 MHz, 2 stage, R.F. protected, dual-gate MOSFETS. Manual gain AGC. 4-3/8" x 1-7/8" x 1-3/8" aluminum case with power switch and



your choice of BNC or RCA receptacles. Available factory tuned to the frequency of your choice from 5 MHz to 350MHz with approximately 3% bandwidth. Up to 10% B.W. available on special order. Requires 12 VDC @10mA

Model 201 price (5 200 MHz) \$29.95 201-350 MHz \$34.95

CONVERTERS 2 METERS

This converter has a minimum of 20 dB gain and a noise figure of 2.5-3.0 dB which assures you of a sensifivity of .1 microvolt or better. The circuit uses a



dual-gate MOSFET R.F. stage and a dualgate MOSFET mixer (thereby giving you a minimum of cross-modulation products), 6 tuned circuits, a bipolar oscillator and .005% crystal. Covers 144-146 MHz at 28-30 MHz output with one crystal included and 146-148 MHz at 28-30 MHz with an extra crystal (available for \$6.00 more). The glass epoxy circuit board is enclosed in a 16 gauge aluminum case measuring $3-1/2'' \times 2-1/4'' \times 2$ 1-1/4" with your choice of either BNC or RCA receptacles. Also included is a power and antenna switch. Requires 12 VDC @ 15 mA. The converter is also available at other input and output frequencies. Call us for prices. PRICE: Model C-144-A available from stock at \$39.95 with one crystal. Additional crystal

\$6.00 extra. HF & VHF 40 dB GAIN 2.5-3.0 N.F. 150MHz

2 RF stages with transient protected dual-gate MOSFETS give this converter the high gain and low noise you need for receiving very weak signals. The mixer stage is also a dual-



gate MOSFET as it greatly reduces spurious mixing products - some by as much as 100 dB over that obtained with bipolar mixers. A

SYNTHESIZERS

FOR ALL TRANSCEIVERS The STR series synthesizers are available for any transceiver operating from 20 MHz to 475 MHz that uses crystals in the 5 to 85 MHz range. It has a



thumbwheel dial calibrated for your operating frequency plus a selectable transmit offset of plus or minus 600 kHz, plus or minus 1 MHz, and 2 spare offsets that you can add later. Frequency accuracy is .0005% and spurious outputs are 60 to 70 dB down. To process your order we must have the crystal formula of your transmit and receive crystals. If your transceiver uses 1 crystal for both transmitting and receiving (like the Motorola Metrum 11), you can use our receive synthesizer described to the right. Maximum tuning range per synthesizer is 10 MHz above 100 MHz and proportionally less at lower frequencies. Dial increments are in 1 kHz steps from 5 to 30 MHz and 5 kHz steps above.

Model STR synthesizer price: \$259.95 5.150 MHz 151-475 MHz \$279.95



EXTRA LOW NOISE

Excellent for weather satellite reception and recommended by Dr. Ralph E. Taggart in his Weather Satellite Handbook. Less than 2 dB noise figure and



approximately 17 dB gain. Uses a low noise J-FET in a common source neutralized circuit. Available factory tuned to your choice of frequency from 135 MHz to 250 MHz. Bandwidth approximately 4 MHz. Supplied in a 2-1/4" x 1-1/8" x 1-3/8" die cast aluminum weather proof case with a filter for powering it through the antenna. Requires 12 VDC @ 5 mA. Choice of VHF, type "N", or BNC receptacles.

Model 102 PRICE \$36.95

bipolar oscillator using 3rd or 5th overtone plug-in crystals is followed by a harmonic bandpass filter, and where necessary an additional amplifier is used to assure the correct amount of drive to the mixer. Available in your choice of input frequencies from 5-350 MHz and with any output you choose within this range. The usable band-width is approximately 3% of the input frequency with a maximum of 4 MHz. Wider bandwidths are available on special order. Although any frequency combination is possible (including converting up) best results are obtained if you choose an output frequency not more than 1/3 nor less than 1/20 of the input frequency. Enclosed in a $4 \cdot 3/8'' \times 3'' \times 1 \cdot 1/4''$ aluminum case with power and antenna transfer switch and your choice of BNC or RCA receptacles. Requires 12 VDC @ 25 mA. Model 407A price:

5-200 MHz \$54.95 201-350 MHz \$59.95 Prices include .005% crystal. Additional crystals \$8.95 ea.

UHF 20 dB MIN. GAIN 3 TO 5dB MAX N.F. This model is similar in appearance to our Model 407A but uses 2 low noise J-FETS in our specially designed RF stage which is tuned with high-Q miniature



Handbook

trimmers. The mixer is a special dual-gate MOSFET made by RCA to meet our requirements. The oscillator uses 5th overtone

FOR VHF RECEIVERS

This synthesizer has 8000 channels and can tune a continuous 40 MHz segment of your choice from 110-180 MHz in 5 kHz steps. This will satisfy most of your



requirements in the VHF range and can save you hundreds of dollars in crystals plus a lot of time. Stock units are programmed for receivers with the crystal formula Fc = Fs 10.7 divided by 3 but we can program it to almost any other IF at no additional cost at the time of your order. It is supplied with an interface for plugging in to your existing crystal socket. Requires 12 VDC @ 1/2 amp which is easily obtainable from a low cost power supply. The synthesizer has 4 voltage regulators therefore the power supply need not be regulated. Phase noise is not detectable as the VCO is coarse tuned by a DAC thereby easing the requirements of the phase locked loop. Not affected by vibrations encountered in mobile use. Enclosed in an 8" x 3-7/8" x 1-1/2" aluminum case and supplied with a combination tilt stand/mobile mounting bracket.

Price: Model SR 140D-05 \$179.95

NOTE: We can make any synthesizer from audio to 475 MHz. Call us for prices.



HOW TO ORDER: All items on this page are available only from Vanguard Labs. For receivers and converters state model, input and output frequencies, and bandwidth where applicable. For the fastest service call (212) 468-2720 between 9 AM and 4 PM Monday through Friday, except holidays. Your order can be shipped COD by Air Parcel Post.

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A Complete Autopatch facility that requires only a repeater and a telephone line. Features include single-digit access/ disconnect, direct dialing from mobile or hand-held radios, adjustable amplifiers for transmitter and telephone audio, and tone-burst transponder for acknowledgement of patch disconnect

RAP-200 P. C. Card RAP-200R Rack Mount

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The Data Tone to Dial Pulse converter Model DPC-221 pro-vides full compatibility between Touch-Tone* encoders and rotary dial-pulse telephone exchanges. Two separate outputs for the * and # digits provide remote control operation, and a cancel function permits the caller to automatically stop and reset the converter's dialing circuits. DPC-221 P. C. Board \$219.00 DPC-221R Rack Mount



Complete C-MOS keyer, versatile controls allow wide charac-ter-weight variations, speeds from 5 to 50-wpm plus volume and tone control. Solid state output switching transistors are compatible with both grid-block and solid-state transmitters. Unit also available in kit and wired p. c. board only versions. MIGHTY MOS P. C. Card - Wired P. C. Card - Kit \$19.95



UNIVERSAL TOUCH-TONE ENCODERS

The Data Signal TTP Series of keyboard encoders is used to generate the standard 12 or 16 DTMF digits. The encoders provide fully automatic transmitter keying and feature a delayed Transmit Ready light, an interdigit timer, and a built-in audio monitor. Features also include all solid-state, crystal-controlled, digitally-synthesized tones and an optional internal mount Automatic Number Identifier (ANI). TTP-1 (12-digit) \$59.00 TTP-2 (16-digit) \$59.00 *Touch-Tone is a registered trade name of AT&T.



A complete Autopatch facility, similar to the RAP-200, that additionally provides for half- or full-duplex operation and features built-in compression amplifiers plus a long-distance inhibit function on digits 1 and 0. \$749.00 **RAP-400**



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We have DenTron's New MLA-1200

The MLA 1200 is a compact KW designed to fill the gap between your barefoot transceiver or transmitter and a full power 2 KW amplifier. A single 8875 external-anode ceramic/metal triode. (the same revolutionary tubes that power the MLA 2500) yields 1200 Watts PEP SSB and 1000 Watts DC CW with as little as 70 Watts drive. (An autoinatic swamping circuit prevents damage to the final if more than 100 Watts drive is applied to the MLA-1200.) There are scores of features common to both the MLA-1200 and MLA 2500, like forced-air cooling, all-steel chassis construction with tight fitting black wrinkle finish cabinetry, a plug-in PC board for metering, ALC, and mandatory warm-up timing. The MLA-1200 is the



MLA-1200 - \$399.50 AC-1200 - \$159.50 DC-1200 - \$199.50

same size as our Super Tuner (just 10" W x 6¼" H x 10" D), and weighs only 10 pounds! Twin outboard power supplies are available for AC or DC operation, with the MLA-1200's low filament current drain characteristics allowing for standard 6 foot cabling between units. Both supplies are constructed of high quality, high current components, and are designed for a lifetime of trouble-free operation.

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- 1000 Watts DC input on CW, RTTY, or SSTV
- Forced Air Cooling System

- AC or DC Outboard Power Supplies (AC-1200. DC-1200)
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- **GUARANTEED FOR LIFE**

articulate CW that is relaxing to use and a joy to copy. The paddle assembly will delight the CW purist as well as the recent graduate from a bug or hand key. The superlative "feel" is attained by a magnetic return force, instantly adjustable to exactly the right fouch for you. Weighting, the ratio of dit and dah (bits) lengths to the crossing between them is either automa

to the spacing between them, is either automa-tically or manually varied. In the automatic posi-tion, it is programmed to lengthen the bits at slow speed for enhanced smoothness and decrease them as you advance the speed, for highest articulation.

, it can be adjusted to a constant value. The KR50 is versatile. Dit and dah memories are provided for full iambic (squeeze) keying. Either dit or dah, or both, may be turned off for operation as a conventional type keyer. Self-completing

tion as a conventional type keyer, seri-completing characters at all times. A convenient "Straight key" is built-in for QRS sending or tune-up. Also an internal side-tone and 15VAC/12VDC operation is provided.

The KR50 is designed to have a permanent place in your shack for the years, perhaps decades, ahead. An investment in the enjoyment of CW.

PRICE \$110.00



KR20-A KK2U-A Paddie has unique principle with excellent feel for rhyth-mic CW. Characters are self-completing. Bit weighting is optimized for normal speeds. Manual key button conve-niently located for hand sending. Side tone signal, Reed relay. Plug-in circuit boards, 115VAC or 6 to 14 VDC HWD 21-5" X 4145" X 814", WL 215 bs. PRIFC 600 501

KR1-A KKL-A This is the paddle mechanism used in the KR50. Requires 6-14 VDC for adjustable electro-magnetic paddle return force. Adjustable contact spacing. For iambic or conventional keyers. "Straight key" but-ton. Housed in an attractive metal case with cream font bar 27 4 4° X 6°, WH 119, bb: particle space. PRICE \$35.00



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4027	.50	7441	1.15	74141	1.00	74H08	.35	74504	.35	74LS51 .50
4028	.95	7442	45	74150	.85	74H10	.35	74S05	.35	74LS74 .65
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