the TEMPO SYNCOM S1 DOES IT ALL AND GIVES YOU 800 CHANNELS

This amazing pocket sized radio represents the year's biggest breakthrough in 2-meter communications. Other units that are larger, heavier and are similarly priced can offer only 6 channels. The SYNCOM'S price includes the battery pack, charger, and a telescoping antenna. But, far more important is the 800 channels offered by the S1.

The optional touch tone pad adds greatly to its convenience and the addition of a Tempo solid state amplifier adds tremendously to its power.

**SPECIFICATIONS**

<table>
<thead>
<tr>
<th>Frequency Coverage</th>
<th>144 to 148 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Channel Spacing</td>
<td>Receive every 5 kHz</td>
</tr>
<tr>
<td></td>
<td>transmit Simplex or</td>
</tr>
<tr>
<td></td>
<td>-600 MHz</td>
</tr>
<tr>
<td>Power Requirements</td>
<td>9.6 VDC</td>
</tr>
<tr>
<td>Current Drain</td>
<td>17 ma-standby</td>
</tr>
<tr>
<td>Batteries</td>
<td>8 pieces Ni-Cad battery included</td>
</tr>
<tr>
<td>Antenna Impedance</td>
<td>50 ohms</td>
</tr>
<tr>
<td>Dimensions</td>
<td>40 mm x 62 mm x 165 mm (1 6&quot; x 2.5&quot; x 6.5&quot;)</td>
</tr>
<tr>
<td>RF Output</td>
<td>Better than 1.5 watts</td>
</tr>
<tr>
<td>Sensitivity</td>
<td>Better than 5 microvolts</td>
</tr>
</tbody>
</table>

**SUPPLIED ACCESSORIES**

- Telescoping whip antenna, Ni-Cad battery pack, charger
- Optional ACCESSORIES
  - Touch tone pad: $55
  - Tone burst generator: $29.95
  - CTCSS subaudible tone control: $29.95
  - Rubber flex antenna: $6
  - Leather holster: $16
  - Cigarette lighter plug mobile charging unit: $6
  - Matching 30 watt output 13.8 VDC power amplifier (530): $89
  - Matching 80 watt output power amplifier (580): $539.00

**TEMPO VHF & UHF SOLID STATE POWER AMPLIFIERS**

Boost your signal... give it the range and clarity of a high powered base station. VHF (136 to 175 MHz)

<table>
<thead>
<tr>
<th>Drive Power</th>
<th>Output</th>
<th>Model No</th>
<th>Price</th>
</tr>
</thead>
<tbody>
<tr>
<td>2W</td>
<td>130W</td>
<td>130A02</td>
<td>$209</td>
</tr>
<tr>
<td>10W</td>
<td>130W</td>
<td>130A10</td>
<td>$169</td>
</tr>
<tr>
<td>30W</td>
<td>130W</td>
<td>130A30</td>
<td>$199</td>
</tr>
<tr>
<td>2W</td>
<td>80W</td>
<td>80A02</td>
<td>$169</td>
</tr>
<tr>
<td>10W</td>
<td>80W</td>
<td>80A10</td>
<td>$149</td>
</tr>
<tr>
<td>30W</td>
<td>80W</td>
<td>80A30</td>
<td>$159</td>
</tr>
<tr>
<td>2W</td>
<td>50W</td>
<td>50A02</td>
<td>$129</td>
</tr>
<tr>
<td>30W</td>
<td>30A02</td>
<td>$5</td>
<td></td>
</tr>
</tbody>
</table>

UHF (400 to 512 MHz) models, lower power and FCC type accepted models also available.

The Tempo line also features a fine line of extremely compact UHF and VHF pocket receivers. They're low priced, dependable, and available with CTCSS and 2-tone decoders. The Tempo FMT-2 & FMT-42 (UHF) provides excellent mobile communications and features a remote control head for hide-away mounting.

The Tempo FMH-2, FMH-5 & FMH-42 (UHF) hand held transceivers provide 6 channel capability, dependability and many worthwhile features at a low price. FCC type accepted models also available. Please call or write for complete information. Also available from Tempo dealers throughout the U.S. and abroad.

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- DS3000 KSR Version 2 1st Microprocessor Controlled Amateur Equipment with Editing for Baudot & ASCII 1973
- DS3000 KSR Version 3 1st 3-mode Amateur Send-Receive Terminal for Baudot, ASCII, and MORSE 1974
- ST5000 Low-cost, Big-performance RTTY Demodulator
- ST6000 High-performance RTTY Demodulator 1975

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- 24 line screen
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- WRU answer-back
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GRANDMASTER

MEMORY KEYERS

At $139.95 this MFJ-484 GRANDMASTER memory keyer gives you more features per dollar than any other memory keyer available — and Here's Why . . .

WEIGHT CONTROL TO PENETRATE ORM. PULL TO COMBINE MEMORIES A AND B FOR 1, 2, OR 3 FIFTY CHARACTER MESSAGES.

MESSAGE BUTTONS SELECT DESIRED 25 CHARACTER MESSAGES.

SPEED CONTROL, 8 TO 50 WPM. PULL TO RECORD.

TONE CONTROL. PULL TO TUNE.

VOLUME CONTROL. POWER ON-OFF.

DELAY REPEAT CONTROL (0 TO 2 MINUTES). PULL FOR AUTO REPEAT.

LED INDICATES DELAY REPEAT MODE.

RESETTING MEMORY TO USE AT BEGINNING.

MEMORY SELECT: POSITIONS 1, 2, 3 ARE EACH SPLIT INTO MEMORY SECTIONS A, B, C, D (UP TO TWELVE 25 CHARACTER MESSAGES). SWITCH COMBINES A AND B. POSITION K GIVES YOU 100, 75, 50, OR 25 CHARACTERS BY PRESSING BUTTONS A, B, C, OR D.

NOW YOU CAN CALL CO, SEND YOUR QTH, NAME, ETC., ALL AUTOMATICALLY.

And only MFJ offers you the MFJ-484 Grandmaster memory keyer with this much flexibility at this price.

Up to twelve 25 character messages plus a 100, 75, 50, or 25 character message (4066 bits total).

A switch combines 25 character messages for up to three 50 character messages.

To record, pull out the speed control, touch a message button and send. To playback, push in the speed control, select your message and touch the button. That's all there is to it!

You can repeat any message continuously and even leave a pause between repeats (up to 2 minutes). Example: Call CO. Pause. Listen. If no answer, it repeats CO again. To answer, simply start sending. LED indicates Delay Repeat Mode.

INSTANTLY INSERT message or make changes in any playing message by simply sending. Continue by touching another button.

Memory resets to beginning with button, or by tapping paddle when playing. Touching message button restarts message.

LEDs show which 25 character memory is in use and when it ends.

Built-in memory saver. Uses 9 volt battery, no drain when power is on. Saves messages in memory when power loss occurs or when transporting keyer. Ultra compact, 8x2x6 inches.

PLUS A MFJ DELUXE FULL FEATURE KEYER. Lamarc operation with squeeze key. Dot-dash insertion.

Dot-dash memories, self-completing dots and dashes, jamproof spacing, instant start (except when recording). All controls are on front panel: speed, weight, tone, volume. Smooth linear speed control. 8 to 50 WPM.

Weight control lets you adjust dot-dash-space ratio; makes your signal distinctive to penetrate ORM.

Tone control. Room filling volume. Built-in speaker.

Tone functions. Key transmitters for tuning. Ultra reliable solid state keying: grid block, cathode, solid state transmitters (-300 V, 10 ma. max., +300 V, 100 ma. max.). CMOS ICs, MOS memories. Use 110 VAC or 12 to 15 VDC. Automatically switches to external batteries when AC power is lost.

OPTIONAL SQUEEZE KEY for all memory keyers. Dot and dash paddles have fully adjustable tension and spacing for the exact "feel" you like. Heavy base with non-slip rubber feet eliminates "walking". $29.95 plus $2.00 for shipping and handling.

THIS MFJ-482 FEATURES FOUR 25 OR A 50 AND TWO 25 CHARACTER MESSAGES.

• Speed, volume, weight, tone controls
• Combine memory switch
• Repeat, tune functions
• Built-in memory saver

Similar to MFJ-484 but with 1024 bits of memory, less delay repeat, single memory operating LED. Weight and tone controls adjustable from rear panel. 6x2x6 inches. 110 VAC or 12 to 15 VDC.

THIS MFJ-481 GIVES YOU TWO 50 CHARACTER MESSAGES.

• Repeat function
• Tune function
• Built-in memory saver

Similar to MFJ-482 but with two 50 character messages, less weight controls. Internal tone control. Volume control is adjustable from rear panel. 5x2x6 inches. 110 VAC or 12 to 15 VDC.

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MISSISSIPPI STATE, MISSISSIPPI 39762

More Details? CHECK — OFF Page 118
contents

12 small beams
Leslie A. Moxon, G6XN

26 voltage-tuned oscillator
Norman J. Foot, WA9HUV

32 operation characteristics of the 555 timer
Robert J. Marshall, WB6FOC

40 receiver digital display
Frank C. Getz, N3FG

46 1-MHz oscillator
Thomas V. Cefalo, WA1SPI

48 matching grounded-grid tubes
Ulrich L. Rohde, DJ2LR

50 the key-toggle
Donald E. Hilldeth, W6NRW

56 i-f transformers
Joseph J. Carr, K4PV

62 updating the HW-2036
Michael J. Miller, WB6TMH
Edmund B. Fitzgerald, WABODR

68 the dasher
Joseph H. Fenn, KH6JF

70 passive phase-shift networks
Robert B. Wilds, K6ZV

76 cabinet construction
Robert A. Thompson, W7KDM

82 digital circuits: propagation delay and flip-flops
Leonard H. Anderson

4 a second look
6 letters
118 advertisers index
105 flea market
112 ham mart
88 ham notebook
92 new products
8 prestop
118 reader service
56 weekender

march 1979
This is the time of year when many high-school seniors are scurrying around, planning their future education, sending applications off to the college of their choice, and taking entrance exams. Seniors who are also Radio Amateurs are probably considering a career in electronics. If they’re lucky, they will have a knowledgeable guidance counselor who can steer them in the right direction; if not, they’ll probably pick a school with a good reputation and work from there. Sometimes this works out, and sometimes it doesn’t — it depends entirely on what the student is looking for.

Electrical and electronics engineers who graduated more than ten years ago would probably not recognize the engineering curriculum now offered by their old alma mater because, in the past few years, there have been significant changes in engineering education. During the 1960s the classical engineering educational programs tended to become more and more theoretical oriented, with less emphasis on applied engineering. The backgrounds of some electrical engineering staffs changed from being primarily applied electronics to applied mathematics, and attempts to develop practical engineering programs were not all that successful. In recent years, however, some engineering colleges have restructured their curriculums for a better balance between the theoretical and the practical. On the other hand, some colleges have continued to stress the theoretical aspects of engineering science, so the prospective student is faced with a very important, but difficult, choice.

Not too long ago, the prestige of an engineering school was almost always gauged by the theoretical emphasis of its courses; each school tried to outdo the others in the theoretical sophistication of its curriculum. Unfortunately, the majority of jobs within the sphere of electronics engineering does not require such an advanced mathematical sophistication as they do a “gut” understanding of electronics. If you talk to students at a theoretical school, you’ll find that many of them don’t know how to solve a simple steady-state ac problem, although they can invert a matrix and use state-variable techniques.

The difficulty with this type of engineering education is that graduates are not adequately prepared to solve the day-to-day engineering problems they will be presented with in industry. Employers are faced with the prospect of several months of on-the-job training before the newly hired engineer becomes a fully contributing member of the staff. Obviously, a new engineer who can solve problems quickly and practically in the real world is a valuable asset.

In the 1970s several colleges introduced four-year electronic technology programs in an attempt to get back to the old practical engineering concept; the courses at these colleges emphasize electronic hardware and laboratory techniques as well as electrical theory. Although graduates of these Bachelor of Science Technology programs have been pictured as fitting into the occupational spectrum somewhere between the technician and the engineer, many professors see technology graduates as having much wider employment opportunities. In fact, technology graduates now have opportunities in many areas of electronic applications and design traditionally occupied by engineering graduates, jobs vacated because of the change in emphasis in engineering education programs.

Students who are interested in this type of engineering program should be aware that there is a wide difference in B.S. programs parading under the “Technology” banner. Some curriculums are managerially oriented, others are slanted toward applications and design, while still others are little more than two-year electronic technician training programs with added courses in the arts and humanities to fill out four years. Students who wish to enter this area should obviously choose a school carefully to be sure they get exactly what they want.

Jim Fisk, W1HR
editor-in-chief
IC-701, Your Synthesized Passport

Enter the exciting world of HF DX with ICOM’s outstanding, fully synthesized IC-701. Globe-spanning QSO’s are as easy as hook-up and tune-in. Complete installation requires only a good 50 Ohm antenna and an AC power plug-in. Your IC-701 comes with everything else you need for beginning DX transmissions, including the matching IC-701PS external speaker and power supply, the fine SM-2 base microphone, and even two built-in VFO’s.

Turn on the power, and the world’s at your single fingertip. The IC-701 lets you scan all the Amateur HF bands from 160M to 10M (plus some MARS coverage above and below some of the Ham bands) with one finger. No more fooling around with two or more tuning knobs, and no complicated retuning when you QSY.

When talking on your IC-701, you get a 200 watt PEP input signal whose punch is significantly increased by the high quality built-in RF speech processor. This makes your 200 watts sound like so much more that we recommend you leave the speech processor on all the time.

For adding on frequency memory and remote frequency control, the IC-701’s synthesizer is completely compatible with ICOM’s RM2 remote computer controller and with ICOM’s optional EX1 extension, you can operate with the RM2 and a linear amplifier at the same time.

Nothing else matches the value and ease of the IC-701. Plunge into the excitement of HF DX now, and get the whole HF world with ICOM’s IC-701 LSI system.
metric dimensions

Dear HR:

The meter is a widely used unit of length abroad. It is convenient only in that fractions and multiples are decimal. As a specific unit of length it is as arbitrary as the foot; for reference, see the NBS history book Measures for Progress. Navigators use the international unit of distance, the nautical mile (based on the earth’s equator), rather than the result of two French surveyors’ efforts in measuring the distance from Paris to Marseilles.

There are times when metric measurements are useful, other times when other standards are better. Having used so many different standards for so long, I am not really partial to all-metric or all-other. For construction and items used therein it would seem advisable to state the local measurement (depending on the author’s location and supply) first, then the conversion. For example, the W7DI antenna in the November issue used tubing from a U.S. supplier. It comes in standard diameters in inches. The same for plumbing pipe. One is hard pressed to make a supplier understand millimeters or centimeters. It would seem reasonable to state such dimensions in inches first with cm or mm in parentheses. If DK1AG had written the article it would seem reasonable to expect the local German dimensions first, then inches. I am referring mainly to those dimensions where specific material is widely available.

A sheet of plywood in the United States is going to be 4 by 8 feet for a long while to come. Three-quarter-inch plumbing is going to remain ¾” for a long time — in fact, any material used in the building trades can be expected to remain in established dimensions for the next ten years — perhaps longer. Many electronic items also remain in the “English” standard. Most numerous are the ICs. They remain rooted on the 0.06-inch (1.27-mm) grid for a simple reason: The United States started it and uses millions of them every year. And many are fabricated in metric countries.

There are a host of other standards. Why the 1¾” increment on the height of rack panels? Think of how long those have persisted. Some rack panel heights have varied, but the 19-inch (42.3-cm) width is still here. Quarter-inch (6.4-mm) shafts are common. Fasteners are described in threads per inch with roughly arbitrary diameters; both U.S. and foreign types should be differentiated. Consideration should also be given to the enormous number of United States types in use and made each year.

The scientific community remains in both camps and is not certain on a few items. There is increasing use in optics of the nanometer instead of the Angstrom, for example. Seimens has yet to replace mhos. Parsecs, kilometers (why not megameters?), and light-years seem to be interchangeable in astronomical distances. At least to NASA.

We have made the transition from tubes to transistors but tubes are still here. I believe that a double standard can continue, and should do so if certain material is in common supply in the author’s country; contemporary literature should reflect that fact.

It is difficult for all to go completely metric, both for editor and author. We will continue to write on 21.6 by 27.9 cm paper using 3.94 pitch, double-spaced. I am awaiting delivery of a new IBM typewriter with a 38.1-cm carriage and dual pitch (4.72 for correspondence). If it bothers me, I’ll just pour a glass of milk from the 1.89-liter carton.

Leonard H. Anderson
Sun Valley, California

Mr. Anderson’s view of metricization is reasonable and well stated. The response we received to our boxed “metrics only” editorial which appeared with W7DI’s article in the November issue was both immediate and loud. Based on the letters which have crossed my desk during recent weeks, ham radio will be using both metric and English dimensions in our magazine articles for the foreseeable future.

W1HR

zip-cord feedlines

Dear HR:

The article on zip-cord feedlines in the April, 1978, issue and the follow-up comments in October brought back some old memories. I first heard of lamp-cord around 1930 (twisted-pair in those days) from the Globe Wireless operators who used it for transmission line to their “noise reducing” receiving antennas. Later there were a number of articles in the Amateur magazines on how to use lamp-cord transmission lines for both receiving and transmitting dipoles; it provided a reasonable match to the feedpoint impedance. Lamp-cord may have been a little lossy, but the early forms of coax which came into use a few years later was not all that good either!

Wayne W. Cooper, AG4R
Miami Shores, Florida
Instant recall.
(and with 10 or 25 watts of power)

Kenwood offers you a choice.
The TR-7600 (10-watt) or TR-7625 (25-watt) with optional RM-76 Microprocessor Control Unit.

TR-7600 and TR-7625...one of them is sure to fit the needs of today's Amateur Operator who's looking for optimum versatility in a 2-meter FM transceiver. And, when either rig is combined with the RM-76, a whole new dimension unfolds in channel memory and scanning capability. Here's what you get:

TR-7600 AND TR-7625 (ONLY)
- Memory channel with simplex or repeater (+600 kHz transmitter offset) operation.
- Mode switch for operating simplex or for switching the transmit frequency up or down...or for switching the transmitter to the frequency you have stored in the TR-7600 or TR-7625's memory (while the receiver remains on the frequency you've selected).
- Digital frequency display (large, bright, orange LEDs).
- Dual concentric knobs for fast, easy selection of any 2-meter frequency, in 100-kHz and 10-kHz steps.
- Full 4-MHz coverage (144.000-147.995) on 2-meters; 600 channels; 5-kHz offset switch, and MHz selector switch...for desired band (144, 145, 146, or 147 MHz).
- UNLOCK indicator...an LED that indicates transceiver protection when the frequency selector switches are improperly positioned or the PLL is not locked.

TR-7600 AND TR-7625 WITH RM-76
- Store frequencies in six memories (simplex/repeater).
- Scan all memory channels.
- Automatically scan the band in 5-kHz steps.
- Manually scan up or down in 5-kHz steps (or last tune).
- Set lower and upper scan frequency limits.
- Clear scan (for transmitting).
- Stop scan (with HOLD button).
- Scan for busy or open channel.
- Select repeater mode (simplex, transmit frequency offset (+600 kHz or +1 MHz), or one memory transmit frequency).
- Operates on 143.95 MHz simplex (MARS).
- Adaptable to all MARS frequencies.
- Display indicates frequency (even while scanning) and functions (such as auto-scan, lower scan frequency limit, upper scan limit, etc...transmitting out of band).

See the exciting TR 7600, TR 7625 and RM 76 now at any Authorized KENWOOD Dealer!
JIM FISK, WHR and Editor-in-Chief of ham radio and Ham Radio Horizons magazines, suffered a heart attack in late January. The following week Jim was out of intensive care, with vital signs stabilized and all indications good. He will, however, be in the hospital for two to three weeks and resting at home for another six to eight weeks after that.

NOVICE USE OF 220-225 MHZ was proposed by the ARRL in a Petition for Rule Making submitted to the FCC. In its petition, the League suggests that Novices be authorized A1, A2, and F3 privileges across 220-225 MHz, with a maximum input power of 50 watts. A Novice would not be able to set up and operate a repeater or auxiliary station, although he would be able to use one.

In The Petition, the ARRL noted that the Commission had recently changed the character of the Novice license from a short-term, nonrenewable "learner's permit" to a full-term renewable license. This, they noted, reduced one of the most serious former objections to Novice phone privileges, the tendency of some Novices to operate mostly or entirely on phone until their license expires, putting them off the air — often for good.

FM In The 52.0-52.5 MHz segment of six meters was proposed in mid January by another ARRL Petition for Rule Making. This recommendation stems from a resolution to promote better use of the six-meter repeater sub-band. The resolution was passed unanimously at the board of directors July meeting.

Responding In An unusually short time, the FCC assigned RM-3314 to the 220-MHz Novice proposal and RM-3313 to the six-meter FM petition. Comments on both RMs must reach the FCC before March 1.

BROADCAST INTERESTS MAY THREATEN the U.S. Amateur Radio WARC position far more than the question of sharing 220 MHz with Maritime services. U.S. international broadcasters, not satisfied with the 865 kHz of new hf spectrum proposed for them in the Commission's carefully worked out WARC Report and Order, are now waging a behind-the-scenes battle to double that amount at the expense of other hf users. If they succeed, it will mean cuts for the other services, whose hf expansion was proposed in the U.S. position, and, in the process, it will destroy much of the carefully worked out agreement achieved by various government agencies and industry advisory groups during the past several years.

The Broadcasters Position is supported in a joint separate statement of commissioners Washburn and Quello that's included in the WARC Report and Order. In it, the two commissioners note that the international broadcasters — specifically the International Communications Agency (Voice of America) and the Board for International Broadcasting (Radio Free Europe and Radio Liberty) — want international broadcasting to have another 800 kHz over and above the new 865 kHz already proposed for it. They also point out that the Executive Branch of the government has yet to make its final decision on the potentially explosive issue.

WILSON ELECTRONICS HAS BEEN bought by Regency. The move puts Regency back into the Amateur Radio business it left early last year, though Regency itself plans to remain in scanners and marine radio while Wilson, operating under its present management, will continue in Amateur and CB products.

FCC'S PERSONAL RADIO DIVISION was abolished as part of a sweeping reorganization of the Safety and Special Services Bureau. In the reorganization, Safety and Special Services has been arranged functionally, rather than by service, into four divisions: Policy Development, Rules, Licensing, and Compliance; Safety and Special Services previously had three "service" divisions, Personal Radio, Aviation, and Marine; Industrial Safety; and Public Service, all of which went out of existence in the reorganization. In addition, there had also been an Industrial and Public Safety Rules Division, plus a Legal, Advisory, and Enforcement Division.

The Result Of The Change is expected to be an improvement in Safety and Special Services' overall efficiency, with much more flexibility in shifting workloads. What it will mean for the FCC licensees such as Amateurs is less clear. Improvement in such routine matters as license processing seems certain, but getting a question answered or discussing a problem might be more confusing than under the old setup.

THE CHICAGO FM CLUB has just announced its First Annual "Call for Technical Papers Competition." The winning entry will receive $500 plus expenses (less transportation) for the best Amateur Radio oriented paper. To receive the prize, the winner must be present at Radio Expo '79 to present his paper and for the prize presentation ceremony. In addition, ham radio will consider buying and publishing the winning paper. The names of the winner and runner ups will be published in the September issue of ham radio. All entries must be received by June 1, 1979. A detailed set of rules and instructions and application forms are available from Radio Expo '79, Post Office Box 305, Maywood, Illinois 60153, or phone (312) 345-5252.
OMNI HAS IT ALL. All the advantages and capabilities, all the new convenience and new levels of performance you need, whatever your HF operating specialty. All built-in, ready to use.

ALL SOLID-STATE. All the advantages of total solid-state from the pioneer of HF solid-state technology. Reliable, cool, stable — from receiver front-end to transmitter final.

ALL HF BANDS. From 160 through 10 meters (and all the crystals) plus convertible 10 MHz and "AUX" band positions for possible future needs.

ALL BROADBAND. Band changing without tuneup — without danger to the final amp.

ALL READOUTS. Choose OMNI-A for analog dial (1 kHz markings) or OMNI-D for six 0.43" LED digits (100 Hz readability.)

ALL VOX AND PTT FACILITIES built-in; 3 VOX controls plus PTT control at front and rear jacks for external PTT switch.

ALL SQUELCH NEEDS for tuning and monitoring are built-in.

ALL FILTERS INCLUDED: CW/SSB filter (150 Hz bandwidth with 3 selectable skirt contours) plus 8-pole Crystal filter (2.4 kHz bandwidth, 1.8 shape factor.)

ALL MODE SWITCH puts all filters to work in any mode.

ALL BREAK-IN: Instant or delayed receiver muting to fit any band condition or mobile operation.

ALL-VERSATILE OFFSET TUNING; dual ranges, ±5 kHz range for off-frequency DX or ±0.5 kHz range for fine tuning.

ALL-SENSITIVE RECEIVER: from 2 µV on 160 m to 0.3 µV on 10 m (10 dB S+N/N) for ideal balance between dynamic range and sensitivity.

ALL OVERLOADS HANDLED; dynamic range typically exceeds 90 dB and PIN diode switched 18 dB attenuator also included for extra overload protection.

ALL LINEAR/ANTENNA BANDSWITCHING FROM FRONT PANEL; auxiliary bandswitch terminals on back panel for external relays or circuits are controlled simultaneously by the OMNI bandswitch.

ALL INTERFACE JACKS FOR PHONE PATCH; access to speaker and microphone signals.

ALL-LEVEL ADJUSTABLE ALC; set output from low power to full, retain low distortion at desired drive to power amp.

ALL SIDETONE ADJUSTMENTS; pitch and volume

ALL-POWERFUL, ALL-WARRANTED FINAL AMPLIFIER. 200 watts input to final. Proven design with full warranty for first year and pro-rata warranty for additional 5 years.

ALL 100% DUTY CYCLE. For RTTY, SSTV or sustained hard usage.

ALL-MODE POWER: basic 12 VDC for easy mobile use, external supplies for 117/220 VAC operation.

ALL FRONT PANEL MICROPHONE AND PHONE JACKS. Convenient.

PLUS ALL THE OTHER HANDY BUILT-INS: "Timed" 25 kHz crystal calibrator in OMNI-A with automatic 5-10 sec. "on" time for easy 2-hand dial skirt adjustment... Zero-Beat switch for placing your signal exactly on CW listening frequencies... SWR bridge switches "S" meter to read SWR each time you transmit for continuous antenna monitoring... Separate receive antenna capability... Dual speakers for greater sound at lower distortion... Plug-in circuit boards for fast, easy field service.

ALL-FUNCTIONAL STYLING. "Clamshell" aluminum case clad in textured black vinyl with complementary nonreflective warm dark metal front panel and extruded aluminum bezel and bail. Convenient controls. Complete shielding. And easier-to-use size: 5 3/4"h x 4 1/4"w x 14"d.


Model 545 OMNI-A $899 Model 546 OMNI-D $1069

Experience the all-encompassing HF world of OMNI. See your TEN-TEC dealer or write for all the details.

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April 27, 28, 29, 1979
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high performance small beams

For antenna sizes up to that of the quad or three-element Yagi, there is no direct connection between size and gain. This article shows how to design small beams without sacrificing gain.

In view of the urgent need for small beams, the number of them in use is remarkably small. In a recent sample of 14-MHz contacts with Australian amateurs, I found that eighty per cent of the stations worked were using quads or three-element yagis; the remainder, the ones with the biggest signals, had even larger beams, beyond the resources of most of us. There are, nevertheless, some smaller beams, such as the VK2ABQ and the capacitively loaded quad,\textsuperscript{1,2} which usually give a good account of themselves. Of particular importance, in the present context, is the fact that some of the most outstanding signals observed during my 50 years as a licensed amateur have originated from stations using driven arrays with only two elements!

The significance of this lies in the ease with which the performance of such beams can be calculated,\textsuperscript{3} and the fact that size comes into the calculations only when estimating the efficiency, which remains high for element sizes and spacings down to approximately 3 meters (10 feet). Translating this fact into practice has brought to light some interesting problems, in particular the problem of overcoupling. That is so basic a problem that the absence of references to overcoupling in the literature suggests that no serious attempt has, until this time, been made to make beams as small as possible.

To achieve this reduction in size, elements must be capacitively end-loaded, without resorting to lossy inductances. This involves large concentrations of metal, which, besides tuning the elements, also couple them very tightly together. Following the normal behavior of coupled circuits, a large secondary (parasitic element) current and a small primary (driven element) current are generated, and there is virtually no beam action. Fortunately, I have found it quite easy to overcome this problem by neutralization similar to that used in push-pull amplifiers.

Another inevitable consequence, when small size is combined with high efficiency, is narrow bandwidth; this makes it essential to use separate feeders for each element so that fine tuning can be carried out in the shack. Also, this arrangement brings with it other important advantages. Since the beam is instantly reversible, less time is wasted in beam rotation, and, because less than 180 degree rotation is required, you can use low-loss, open-wire feed lines as well as simpler and cheaper methods of beam rotation. The direct relationship between the size and performance of large beams is well known, and some readers may find it difficult to accept that a small beam can be as good as a big one, particularly if those readers have experience with typical small beams using large loading inductances. The claim is, on the face of it, improbable, and one might think that it could be dismissed by invoking some general scientific principle, as in the case of perpetual motion. Looking for such a principle we come instead to the surprising discovery that, although the gain of big beams is limited, there is absolutely no limit to the theoretical gain from an antenna, provided it is small enough! You would be justified in some skepticism at this point, since it turns out that gains much in excess of 6 dB are impracticable unless the boom length is increased to half a wavelength or more. It is

By Leslie A. Moxon, G6XN, 1 Stoner Hill House, Froxfield, Petersfield, Hants, England
possible, though, to go down in size to about 3 meters square (10 feet square) without dropping below about 4.5 dB gain.

If such statements are found puzzling, it is probably because of the failure in most of the literature to distinguish between two completely different methods of beam formation, additive and subtractive. The subtractive method is typified by the W8JK array, and, as I have shown elsewhere, most amateur high-frequency beams can be regarded as derived from or related to this array. The radiation patterns are calculable without using any variable other than the direction. Therefore, the gain is independent of size, provided the efficiency remains high enough to ensure that most of the power is radiated. It is this constraint, allied with the need for adequate bandwidth, which in practice limits the gain of subtractive beams to about 6 dB.

fig. 1. In (A), the elements are fed 180 degrees out of phase so that the fields add in phase along the line at right angles to the elements. A receiver located 45 degrees off the main beam sees a phase difference corresponding to AC-AB, i.e., 54 degrees, causing a drop of 4 dB in signal level in addition to the 3 dB which would be expected for a single element. Thus, the use of two elements has produced a narrower beam by virtue of the wide spacing. For subtractive gain, as shown in (B), the elements are closely spaced. Radiation would be cancelled except for a small phase shift of which the maximum value corresponds to the distance AC, or 36 degrees for a spacing of λ/10. At 45 degrees to the beam, the phase shift is reduced by the ratio of AB to AC (11 degrees), which translates into a 3 dB drop in signal level. If the fields are additive in the wanted direction, the 11-degree shift is of no consequence and the radiation pattern becomes that of a single dipole.

fig. 2. Method for estimating gain from the radiation resistance. Diagram (A), which is drawn to scale, shows how the W8JK element currents of 2.12 amps "add" to give the equivalent of 1.61 amps in a single element. The currents are correct for a radiated power of 72 watts. This would produce a current of 1 amp in a dipole, so that the voltage gain is 1.61, i.e., 4.2 dB. (B) shows how if one element is given a phase lead equal to the spacing, only about half as much current is needed to produce the same field strength. In each case, the current is obtained from

The following discussion is intended to provide further insight into the problems of designing small beams, so that you can make your own choice from the available options and then design the "best-possible" beams for the space and facilities you have available. Practical details and some performance data are given for a number of designs, but these are intended as guidelines and not as blueprints. Although the feasibility of the 3 meter square (ten foot square) design has been proved, my own preference, given the space, is a 5.2 meter square (17 foot square) 3-element design, which, besides providing slightly more gain and greater bandwidth, lends itself to a particularly light form of construction and easier methods of achieving multiband operation.

It is a strange paradox that the main advantages of increased size relate to operation at the higher frequencies, to the extent that considerable extra gain can be obtained at 28 MHz by using full-size, 14-MHz elements. For no reason that I can discover, this advantage is usually thrown away by using traps or nesting to "cut the elements down to size" at the higher frequencies.

gain variation vs antenna size

Gain results from concentrating the radiation in one particular direction at the expense of other directions. This can be achieved by arranging more-effective addition or, as with the W8JK, less-
effective cancellation of radiation for the wanted direction. Addition requires wide spacing, as in fig. 1A, where the elements are arranged so that their fields add in phase for the wanted direction. Viewed at an angle of 45 degrees to the line of fire, the elements are closer together by 0.29 wavelength, which translates into a phase shift of 104 degrees and a corresponding drop of 4 dB in signal level. This drop of course, is in addition to the \( \cos \theta \), or “angle to the wire” effect, which applies equally to beams and dipoles and amounts to \(-3\) dB at 45 degrees. But for the wide separation of the elements, there would be no phase shift, no narrowing of the radiation pattern, and hence no gain.

On the other hand, given an array of \( n \) elements sufficiently far apart for mutual interaction to be ignored, we can provide each \( \frac{1}{n} \) of the power. Since the voltages at the receiver add in phase, there is a power gain equal to the number of elements. Because the antennas have to be separated by at least \( \lambda/2 \), it turns out, not surprisingly, that gain is proportional to size. During reception, such an antenna collects most of the energy contained in the volume of space which it occupies, thus giving rise to the concept of aperture, for which it is sometimes claimed “there is no substitute.” It is perhaps difficult to conceive that a tiny beam can collect as much energy as a big one, but the explanation lies in the high \( Q \) of the smaller antenna. Just as tuned circuits couple together more tightly when their \( Q \) is increased, so reduction in the radiation resistance is accompanied by “tighter coupling” of the antenna into the surrounding space.

To understand the mechanism of small beams, the W8JK system shown in fig. 1B is the easiest starting point. Radiation would be completely cancelled in all directions except for that caused by the phase shift, which results from the elements not being exactly the same distance from the receiver. Field strength is proportional to the apparent separation \( \Delta B \), which, like the radiation from a single element, varies as \( \cos \theta \). In simpler language, the field strength, at a given distance and direction, is proportional to the apparent length multiplied by the apparent separation of the elements. The separation factor applies equally in a plane at right angles to the diagram, so that the radiation pattern of a horizontal W8JK array is given by \( \cos^2 \theta \) in the horizontal and \( \cos \theta \) in the vertical plane. For a single horizontal wire, the pattern is \( \cos \theta \) in the horizontal plane and an omni-directional pattern in the vertical plane. The important point to note is that none of these patterns contain any reference to the dimensions, though we are of course assuming them to be small. In fact, the conclusion that gain is independent of size turns out to be accurate within about 0.5 dB for spacing up to \( \lambda/4 \) and element lengths up to just over \( \lambda/2 \). Up to these sizes, each element behaves in the same way, i.e., as a “point source” of energy. It is only when an element is very large that appreciable extra gain, or directivity, can
be expected, as for example when a 14-MHz element is used as two half-waves in phase on 28 MHz. It should be obvious from inspection that quad loops, even full-sized ones, do not meet this condition.

To prove the point, there is no harm in treating the loops as a pair of stacked dipoles. Various handbooks provide data indicating that the stacking gain for $\lambda/4$ spacing is only 1 dB. However, some of the 1 dB is lost by bending over the ends, since a half-wave dipole has only a small gain (up to 0.4 dB) when compared with shorter dipoles; a further small amount is lost due to radiation off the ends. Even more is lost when parasitic elements are added, since the stacked dipoles then become stacked yagis, and, according to the usual rules, the higher the gain of individual antennas the further apart they have to be placed in order to achieve an appreciable stacking gain.

It is unfortunate that many wild claims have been made for the quad, some of them involving professional journals and computer studies. It needs to be stressed that measurements are very difficult and computers need to be asked the right questions. The habit of accepting figures without checking them against ordinary common sense is not confined to the Novice! In fact, as I have found, better low-angle gain is obtained by omitting the lower halves of quad loops and using the upper halves as inverted V or U elements. This increases the mean height by 2.4 meters (8 feet) at 14 MHz; low-angle gain for a flat, unobstructed site being proportional to antenna height, this more than offsets any slight loss of free-space gain for heights up to about 21 meters (70 feet)!

The real advantage of the quad is the large amount of extra gain (3-4 dB) obtainable by using the 14-MHz elements at 28 MHz with suitable resonators so that they become a “Bi-Square,” but this is rarely exploited. Another frequently made claim, that the quad provides better DX signals by “lowering the angle of radiation”, is also without foundation; the lobes of practical antennas are too broad to discriminate between direct and ground reflected waves which must always interact in the same way, resulting in a loss at low angles unless the antenna is very high or the ground sloping. To obtain the effective gain at a low angle of radiation, the free-space gain is multiplied by the ground reflection factor, which, for horizontal antennas, depends only on antenna heights and radiation angle, being the same for a quad, yagi, dipole, small rhombic, or even a minibeam!

The 4-dB gain of the W8JK can be estimated very roughly from the 3-dB widths of the radiation patterns discussed earlier, or, more accurately, from mutual-impedance data. For $\lambda/2$ dipoles spaced at $\lambda/8$, mutual impedance is 64 ohms, which has to be subtracted from the self-resistance of 72 ohms so that for each element the radiation resistance is 8 ohms. If the power available is 72 watts, you would have a current of 1 ampere in an ordinary dipole, or 2.12 amps in each of the W8JK dipoles. The phase difference due to spacing is 45 degrees, and, by “completing the triangle” as in Fig. 2, you’ll see that the distant field is the same as that which would result from 1.61 amps in the dipole, i.e., there is a voltage gain of $1.61/1$ or 4.2 dB.

Now, if the phase of the current in one element is advanced by an amount corresponding to the spacing, the total phase shift becomes zero for one direction and is doubled for the other direction. The resultant unidirectional (cardiod) pattern would require only about half the current to produce a given field strength. It happens (by coincidence) that the gain is unchanged, so that the effective radiation resistance has been multiplied by about four, proving that the beam can be made a lot smaller with the same results. Fig. 3, adapted from references 3-6, shows how radiation resistance varies with element length, spacing, and phase angle, from which it can be seen that provided you shorten the dipoles by sacrificing the ends and not the middle, the length can be reduced to 37 per cent of a half-wavelength before the radiation resistance drops to its W8JK value, giving a length of only 37 meters (12 feet) at 14 MHz. This reduction in size could be carried out by taking the W8JK beam and folding its ends to fit the available space. Whether or not the example can be improved upon, it does demonstrate that if the W8JK beam works properly you could expect to be
equally successful at 14 MHz with a 3.7 x 2.4 meter (12 x 8 foot) miniature beam which has been adjusted to give a large front/back ratio.

Thus, changing the size does not alter the gain, but, obviously, if the length or spacing is halved, the current must be doubled to maintain a constant signal, so that the radiation resistance has been divided by four. In the same way, you could alter the shape to just under 3 meters square (10 foot square) without affecting matters in any way, save that the diameter of turning circle is slightly decreased and it becomes easier to handle the ends.

Whatever shape is selected, radiation resistance may be obtained by multiplying the resistances plotted in fig. 3A by the factors obtained from figs. 3B and 3C. This is not quite the full story, since, as demonstrated by fig. 3 and 4, an extra dB of gain results from using an intermediate value of phase shift. However, this halves the radiation resistance and doubles any power loss. From the lower gain curves in fig. 4, it can be deduced that for a 3-meter-square (10-foot-square) 14-MHz beam, using no. 12 AWG (2.1-mm) wire elements with bent ends, a net gain of 4.5 dB should be obtained after allowing for resistance losses. There may be a slight additional loss due to the "short dipole" effect mentioned, but this is caused by a small difference in the amount of endwise radiation and is much less in the case of a beam.

To put this into perspective, the maximum gain theoretically possible for a three-element parasitic array on a quarter-wavelength boom is 7.5 dB, but in this case, the radiation resistance is only 4 ohms (even less than that of the small beam), and the maximum gain normally found in practice is less than 6 dB. Theory goes on to predict a possible gain equal to the square of the number of elements, but before getting far along this road you're faced with huge currents and voltages, infinitesimal radiation resistances, and microscopic tolerances to the extent that the 6-dB figure is unlikely to be exceeded in practical rotary beams for 14 MHz without increasing the boom length to at least $\lambda/2$, which is sufficient to provide some "additive" gain. Reducing the size below 3 meters square (10 feet square) leads to a similar rapid increase of practical difficulties. And, as I shall point out later, there are substantial advantages, particularly for multiband operation, in an increase to about 5.2 meters square (17 feet square).

Limitations on the reductions possible in antenna size result from the following:

1. Drop in efficiency, i.e., the radiation resistance becomes comparable with the losses. This is basic and imposes a well-defined, practical limit.
2. Narrow bandwidth — the acceptable lower limit depends on the skill of the designer and the skill plus patience of the operator.

Short elements have to be loaded to bring them to resonance. The aim must be to keep the radiation resistance high, which requires end loading, and the loss resistance low, which rules out the use of lossy devices such as coils or long resonant feeders. Fig. 5 shows ten methods that have been used, with only 5A through 5D meeting the radiation resistance requirement. The others tend to be more convenient, but use the triangular tips instead of the center of a sine-wave current distribution, thus halving the average current and dividing the radiation resistance by four, besides requiring an inductive and therefore relatively lossy loading device in the center. Of these methods, 5F is open to least objection, since stubs have less resistance than coils; 5H is particularly bad, since it multiplies the losses of 5F by the total number of half-wavelength in the resonant system. Methods 5I and 5J are used for multibanding,8 which necessarily adds to the losses because of circulating currents in the resonators. Methods 5E to 5H are therefore applicable only for very modest reductions in antenna size.

Of the remainder, 5C makes use of small end-loading capacitances which have their effective values greatly inflated by near resonance with the inductors. In one typical design, the total inductance is about 14 μH, so that for a Q of 200 the loss resistance would be 6 ohms and more or less comparable with the radiation resistance, the element lengths being 3.7 to 4.3 meters (12 to 14 feet). Fig. 5A is ideal insofar as there is negligible loss in the capacitance, but it is difficult to achieve enough loading, so that some inductance on the lines of 5C is likely to be required; in this case, however, it could take the form of short stubs and the losses would be minimal. Arrangement 5B, consisting of a half-wavelength dipole with its ends bent over, is very similar, but simpler and equally efficient. The bent ends are not, of course, pure capacitances and there are losses, but, as has been seen, this method allows high efficiency to be maintained down to very small beam sizes. Efficiency can be estimated with the help of fig. 4.

Another efficient type of small element is the capacitively loaded loop shown in fig. 5D. Loops with 3.2-meter (10.5-foot) sides have been used by G3YDX for a 14-MHz, two-element beam,2 and a design patented by G3IMX achieves two-band operation by using traps to remove the loading at the higher frequencies. For a single 3.7-meter (12 foot) loop, the radiation resistance is 75 ohms. About 20 ohms could be expected for a 1.8-meter (6-foot) loop, the loss resistance (referred to the feedpoint) probably being about 2 ohms. My own attempt to compress a 14-MHz quad into a 1.8-meter (6-foot) cube was unsuccessful, but I still think it might be achieved with rigid, all-metal construction and inductive loading stubs, plus sufficient ingenuity. The problem is to dispose of enough loading in the space available. In this case, neutralization will certainly be required as the lower physical limit is approached. Neutralization will not be required with 3-meter (10-foot) loops, as sufficient loading can be achieved within the plane of the loops without bringing ends into proximity.

It is important in all cases to arrange the loading so that wires carrying appreciable current are not doubled back in a direction parallel to the driven element, in which case they subtract from the wanted radiation, bringing the radiation resistance "down with a bump!" This constitutes one of the main problems of construction.

**Bandwidth**

The bandwidth of a dipole may be roughly estimated from the radiation resistance (R) and the characteristic impedance (Z), which depends on the ratio of length to diameter. Each per cent of detuning produces a reactance of 0.015×Z ohms; and the 3-dB bandwidth points (corresponding to an SWR of 2:1) may be found by equating this to R. Bandwidths are plotted in fig. 6 for 14-MHz dipoles of various wire diameters. The figures should be valid, as an approxi-
fig. 7. The top illustration shows a practical arrangement for a 14-MHz phasing control. The beam reversing switch, though not shown, interchanges the feeders. This circuit may be usable to 28 MHz, but the inductance should be reduced to about 0.8 pH; some experimentation with values may be needed. The principle of operation is demonstrated by (B), as shown the load currents are equal and the ratio of X/R determines the phase shift. If, as is more usual, the loads are unequal, current equality may be achieved by adjustment of the reactances provided in (A). The components of (A) can also be rearranged as a tuner for a parasitic type of beam (see (C)).

mation, even when the dipoles are folded as in figs. 5B or 5F. When two elements are assembled to form a beam, the bandwidth is reduced because of the drop in R, though, for a given value of R it is increased and could be almost doubled by virtue of the “coupled circuit effect.”

It is important to distinguish between three kinds of bandwidth:

1. The pattern bandwidth — the bandwidth over which the radiation pattern and gain are satisfactory
2. The bandwidth for satisfactory SWR
3. The bandwidth over which the antenna is usable subject to adjustments that can be carried out without leaving the shack.

To illustrate, one element of the first version of my small beam had an intrinsic bandwidth (as calculated) of about 450 kHz. The completed minibeam, using a parasitic reflector, had a useful bandwidth of only about 200 kHz; at one edge the front/back ratio dropped below 8 dB and at the other edge the gain had dropped by 1 dB. In wet weather, due to an inadequately treated bamboo spider, the tuning shifted 200 kHz low, outside the desired range. In contrast, with the second version the useful bandwidth was only 130 kHz. However, it was desirable to keep within about 20 kHz of the tune-up frequency. But, by using separate feeders to each element, the beam was tunable to any frequency in the 14-MHz band either as a driven array or a driven element plus reflector. So, for practical purposes, despite the smaller intrinsic bandwidth, the useful bandwidth was greater.

With the parasitic reflector, the phase-shift is related to the ratio of reactance to the total resistance, R, of the reflector, and, in addition, must meet the requirement for reasonable gain in accordance with fig. 4. Although the actual relationship is much more complex, a rough idea of the bandwidth for useful performance can be obtained as follows. Referring to fig. 4, we might decide that performance is acceptable for values of phase shift from one half up to one and one half times the spacing, which means from 25 to 75 degrees in the case of a spacing of 3 meters (10 feet). At 25 degrees, the front-to-back ratio is down to 10 dB and the radiation resistance (fig. 3) is getting very low, while at 75 degrees the gain is down to 3 dB. Phase shift is partly due to X, and partly to the mutual reactance between the elements, which, in this case, takes care (more or less) of the required mean value of 50 degrees. The maximum allowable shift of ±25 degrees occurs when X/R is roughly equal to this angle (measured in radians), i.e., $X = R/2$. From fig. 3, $R = 15$ ohms and for $Z = 1000$ ohms (a typical value) a reactance $R/2$ results from ±0.5 per cent detuning, i.e., the bandwidth is 1 per cent or 140 kHz, in good agreement with the observed figure of 175 kHz (i.e., 14,300-14,125) from fig. 10.

The first version of the small beam was 3.7 meters square (12 feet square), which would be expected to double the radiation resistance and bandwidth; in this case, too, the calculations are in line with the observed performance and there was good agreement in regard to the observed bandwidth for a 20-dB front-to-back ratio. From fig. 4, the theoretical 20 dB points correspond to a change of ±18 per cent in phase angle, or 36 per cent in reactance, i.e., 6 ohms, which translates into a bandwidth of 0.4 per
cent, or 56 kHz in comparison with a measured figure of 55 kHz; however, this must be regarded as coincidental, since both figures are very rough.

**phasing**

It is essential for the currents in the elements to be equal (or nearly so), as well as correctly phased. With parasitic arrays there is usually no independent control of these quantities, though equality tends to be achieved in the case of quad reflectors. With \( \lambda/2 \) dipole elements the inequality, though it degrades the front/back ratio, does not upset gain to a serious extent. In contrast, the small-beam elements behave as overcoupled circuits resulting in large ratios of reflector to driven element currents.

Neutralization allows the coupling to be reduced to any desired extent, so that it can be adjusted in conjunction with the tuning of the reflector to obtain equal currents with any required value of phase shift. In this way, very deep nulls can be obtained in one or more back directions without resorting to driving both elements. This is subject to a number of assumptions: For example, the neutralization, which has to be adjusted at ground level, must remain "right" when the antenna is raised to its full height, at all frequencies, and in all types of weather.

A driven arrangement with provision for adjustment of phases and amplitudes is obviously more versatile and allows compensation for considerable errors in tuning or neutralization, but the methods usually employed for driven arrays are based on false assumptions that are particularly disastrous in the case of small beams with reactively coupled elements. It is usually assumed, for example, that \( \lambda/8 \) of line provides a 45 degree phase shift — which is true if the lines are perfectly matched. But often there is little attempt at matching, which will, in any case, be upset if the phasing is adjusted. A further difficulty is that for the elements to have equal impedances the mutual coupling between them must be a pure resistance, a condition that applies in practice only for straight \( \lambda/2 \) elements spaced just over \( \lambda/8 \). This, incidentally, is an important special case since it is then easy to calculate gain and radiation resistance.\(^3\)\(^5\) Having established that gain is independent of size, the numbers obtained can be applied to other sizes on the basis of fig. 3.

If there is capacitive reactance coupling, as happens with wider spacing or miniature elements, the radiation resistance of the reflector may be zero or even negative, in which case its feeder returns more power to the transmitter than it receives! Matching is clearly impossible under these conditions. The method I use, which can take various forms,\(^6\)\(^9\) is based on resonance; in fig. 7, the two feeders are connected through a series-resonant circuit and the complete system, minus the connection to the transmitter, is made resonant. Phase-shift is obtained by off-center connection of the feedline from the transmitter. If necessary, element currents can be made the same by detuning the series circuit so that one element or the other is brought closer to resonance. Matching should be carried out to reduce the SWR in the line to the forward element, a high SWR in the reflector line being of no importance as there is very little power transferred along it. Cross-over of the elements as with the W8JK, though normally essential with full-sized elements, is sometimes not required when coupling to highly reactive elements. Using this arrangement, it has been possible to work with different or even unequal lengths of feedline, and to compensate for quite large errors of adjustment, though never to the extent of being able to use it as a substitute for neutralizing. The beam can be reversed either by interchange of feeders or moving the coil tap. It is possible, in principle, to null out from any given direction “off the back.” A curious feature of reactive coupling is that, despite the remarkable effect it has on relative impedances, the sum of the two radiation resistances is not affected; for fig. 3 to have universal application, it was therefore necessary to plot the sum and not the individual resistances. The sum is in any case a more useful figure for the present purpose since it is into this value that the feeder for the front element has to be matched (so that the antenna impedance is “repeated” in the shack). It is often possible, particularly if the feeder length is a whole number of half wavelengths, to use the phasing circuit with very little modification as a remote tuner for driven element and reflector operation (see fig. 7); this reduces the

![Diagram](image-url)
ing signal from the back direction, using the reversing switch as required.

**practical designs**

Several construction methods have been used. Fig. 8 gives the dimensions of my first small beam, the elements being suspended by thin polyethylene cords from a six-armed spider made from bamboo garden canes which extended just beyond the wire. The two central arms were required to counter the inward pull of the folded sections. One element was driven using 300-ohm line with a delta match having series capacitors to tune out the reactance. I tried to tune the other for a null in the back direction, but this was barely perceptible. Current in the reflector was found to be higher than that in the driven element, so the neutralizing capacitance (formed by parallel wires) was increased by reducing the spacing, retuning the reflector, and adjusting the driven element for lowest SWR at each step. Eventually current equality and a good front-to-back ratio was achieved. At a height of 16.8 meters (55 feet) the antenna was compared with a two-element quad at 13.7 meters (45 feet) and appeared to be at least as good.

Mention has already been made of bandwidth and wet weather problems. It quickly became obvious that two feeders would have to be used. To save cost, I used the transparent, plastic-type of 300-ohm line; lacking previous experience with it, I was unprepared for the rapid deterioration, which caused a lot of confusion. Later, good results were briefly obtained using a pair of 600-lines.

However, mechanical problems arose and the next step was a new design using metal construction with 3-meter (10-foot) elements made from 19-mm (0.75-inch) tubing. Short vertical rods at the ends of the elements (see fig. 9) act mainly as brackets for supporting wire zigzags. I found that, because of limited space, almost as much wire was needed as in the original design so that the use of tubing provided no electrical advantage apart from getting rid of “wet weather” effects caused by the bamboo. To avoid the need for heavy-gauge elements, their ends were guyed back with polyethylene cord to the boom extensions, thus countering the inward pull of the loading wires. Two open-wire feedlines were used.

Some typical measured performance figures are shown in fig. 10. On-the-air performance was down compared with earlier results, on average about 2 dB; this might be due to the smaller size (radiation resistance being halved) in addition to closer proximity to tree branches, the mast being slightly shorter due to breakage and splicing! Another uncertainty arose from loss of the quad as a standard of reference; the 3-element beam, though equal to the quad...
in performance, was so placed that it tended to screen the smaller beam. I could not be certain that rotation to an end-on position was removing all this effect. From this, I felt that the best test for the small beam would be as a direct replacement for the larger one. Now, however, I used a new multiband version, the disposition and loading of the elements for 14 MHz being as shown in fig. 11.

For such tests my yardstick for many years has been another station using a TH6 at 11.6 meters (38 feet) and with a slightly better location for VK (long path) over which most of the tests were made. With the small beam at 14.6 meters (48 feet), I was able, for the first time on record, to equal the signals of the other station on three consecutive days. Normally with a quad at 12.1 meters (40 feet) I should be down in most cases by 1/2 to 1 S-unit. Added to the initial results of the first small beams, this would provide substance for a claim that small beams, given a few feet of extra height, are actually better than big ones! Unfortunately, such a claim could not be supported by theory or common sense, but I was confirmed in my belief that the small beam can be made fully competitive.

Any type of feeder can be used, but, because of its lightness, good-quality 300-ohm line can be recommended as an alternative to open-wire line. I have also used 50- and 75-ohm coax with 4:1 baluns and delta matches identical to the one shown in fig. 8.

three-element beam

A starting point for this development was the original VK2ABQ design¹ in which a quad loop lying on its side is converted into a two-element beam by insulators in the sides. The rather wide spacing (λ/4) unfortunately means that performance is slight-

![fig. 10. Typical performance of the 1-meter-square (10-foot-square) beam. All parameters were measured with the beam at a height of 2.1 meters (7 feet) and the reflector tuned for 14.2 MHz. The dotted curve shows the improvement in SWR after adding 10 cm (4 inches) to each end of both elements and retuning the reflector for best front-to-back ratio at each frequency. Note: The gain curve is listed as dB down from the maximum.](image)

ly worse than the best that can be achieved with two elements, and it seemed logical to me to put a third element in the middle. No increase in the diameter of the turning circle dictated a maximum element length of 7.3 meters (24 feet). With some end loading by vertical rods and the use of linear resonators, I was able to design a triband element (see fig. 12) having almost as much radiation resistance as a full half-wavelength at 14 MHz, higher radiation resistance than a normal element at 21 and 28 MHz, a small amount of extra gain at 28 MHz, and no trap losses.

Tubing was used for this element, but retention of bent wire elements for directors and reflectors made engineering sense, since these elements have less current flowing in them and a lot of cost and weight could be saved. The next step was to eliminate the quad spider by using the driven element as one diagonal of a square and making it support the ends of the parasitic elements which were now V shaped instead of U shaped and filled in the sides of the square as shown in fig. 12C. The other diagonal became the “boom” and consisted of two 2.4-meter (8-foot) bamboo garden canes joined by a length of aluminum, supporting the apex of the Vs.

V-shaped elements are not suitable for multiband- ing by the methods used for the driven element, and ordinary traps would be unsuitable for suspension in wires supported by such a light framework, as well as being too lossy. I therefore used separate ele-

ments for each band with their ends strung out to different points on the vertical loading rods of the driven elements in the hope that this way they might not get entangled! Though somewhat haywire, this arrangement has unexpectedly survived two severe storms and, although it needs a lot of tidying up mechanically, comes fairly close to the ultimate in lightness combined with high performance. Details in fig. 12 should assist any reader wishing to experiment on similar lines. Despite the larger size, the overcoupling problem was not completely avoided and there was some slight difficulty in adjusting the 14-MHz reflector, but this was overcome by increasing the spacing slightly as indicated in fig. 12C.

The linear resonators work as follows.⁹,¹¹ The in-

ductance L_AB of the central portion (AB) of the ele-

ment is tuned to resonance by C1 so that it acts as an insulator for 28 MHz, making the element two half-

waves in phase. At 21 MHz, about half of C2 serves to eliminate the inductance by tuning it to parallel resonance, with the remainder used to series reso-

nate the inductance (L_0) of the outer portions of the element, L_AB and L_0 chosen to be roughly equal. At 14 MHz, the capacitors are virtually “not there,” their values at 21 and 28 MHz having been inflated by series resonance with their connections, and consequently AB no longer acts as a tuning inductance.
This is an alternative form of the beam shown in fig. 9. In this case (A), the elements have been folded in a horizontal plane to avoid any vertical projections. The metal supporting strut in the center of the boom also serves as a 21/28-MHz driven element in the multiband version. The detail shown in (B) illustrates the arrangement for multibanding with linear traps. Moving the shorting bar from S to S' results in an improved trap for 28 MHz, but separate parasitic elements must then be used for 21 MHz. For mono-band 14-MHz operation, the capacitors are omitted, but extra wire may be needed to obtain sufficient loading. The wire lengths, a, b, c, d, and e, are all equal and approximately 35.6 cm (14 inches); C1 and C2 use the wire segments bs and sd to form the linear traps for 28 and 21 MHz respectively.

Better front/back ratios and bandwidths were obtainable at the expense of gain, thus emphasizing the desirability of remote tuning; in this way it should be possible to achieve effective gains between 6 and 7 dB. I tried to achieve this by using a single pair of additional feeders, attached to coupling loops placed near the centers of each set of parasitic elements, but achieved only limited success. A practical, but rather cumbersome alternative is to use separate feeders to all elements.

At 21 MHz, the SWR was 1.5-1.6 over the whole band; front/back ratio increased with frequency from 13 to 16 dB, but was accompanied by an apparent 2-dB drop in gain. At 28 MHz, bandwidth for an SWR better than 2:1 was 350-400 kHz. Tuning for maximum gain at 28.5 MHz, there was a drop of 2 dB at 28.25 and 29 MHz. The front/back ratio rose from 11 dB at 28.5 to 15 dB at 28.9, but was down to 6 dB at 28.25 MHz. These figures have been selected as fairly typical from a wide assortment and the variations illustrate the degree of improvement to be expected from remote tuning. The linear resonator
3.8 cm (1 1/4 in.)

FROM TUBE ELEMENT LENGTHS

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Reflector</th>
<th>Director</th>
</tr>
</thead>
<tbody>
<tr>
<td>14 MHz</td>
<td>11.2 meters (440 inches)</td>
<td>10.8 meters (426 inches)</td>
</tr>
<tr>
<td>21 MHz</td>
<td>6.9 meters (272 inches)</td>
<td>6.3 meters (259 inches)</td>
</tr>
<tr>
<td>28 MHz</td>
<td>5.1 meters (202 inches)</td>
<td>4.8 meters (188 inches)</td>
</tr>
</tbody>
</table>

BOOM LENGTH FROM CENTER OF DRIVEN ELEMENT TO APEX OF V'S

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Reflector</th>
<th>Director</th>
</tr>
</thead>
<tbody>
<tr>
<td>14 MHz</td>
<td>4 meters (156 inches)</td>
<td>3.2 meters (126 inches)</td>
</tr>
<tr>
<td>21 MHz</td>
<td>2.6 meters (101 inches)</td>
<td>2.6 meters (101 inches)</td>
</tr>
<tr>
<td>28 MHz</td>
<td>2.2 meters (86 inches)</td>
<td>2.2 meters (86 inches)</td>
</tr>
</tbody>
</table>

details shown in fig. 12 are applicable also to "full-size" elements, a reduction in length of about four per cent being required due to the presence of C2.

Earlier I criticized trapped beams, but the situation is quite different with small elements, including the V-shaped parasitic elements of the beam just described. If the folded ends are left in place, at the higher frequencies they degrade performance because of an inefficient current distribution and by radiation from the ends. The only practical way to remove the excess capacitance is by means of traps, but ordinary traps result in appreciable losses even with full-size elements. And, as the radiation resistance drops, the power loss increases in the same proportion so that large losses can be expected. This problem can be resolved with the help of the linear resonator, which, instead of degrading the performance at 14 MHz, actually has a very slight beneficial effect by increasing the total capacitance.

Another problem is caused by excessive spacing between elements at the higher frequencies. This can be avoided by using a separate, centrally placed, driven element for 21 and 28 MHz, a coax feeder with 4:1 balun, and a delta match into a linear resonator, providing two-band operation. To keep within an area 3 meters square (10 feet square), long, vertical, loading rods would have been required, but I thought it was better to increase the element lengths to 3.4 meters (11 feet) and make full use of the 4.7-meter (15.5-foot) turning circle.
Multiband operation, with good front-to-back ratios, was obtained using traps as shown in fig. 11B, but on 21 MHz the bandwidth was too narrow to be acceptable. The structure was braced with a lot of cord ties and I was able to use them to support separate (wire) parasitic elements for 21 MHz.

Capacitive coupling from these into the driven element was excessive, however, and the reflector had to be neutralized. Bandwidth was still rather narrow and devious means were needed to achieve even enough remote tuning for band coverage (without beam reversal) on 21 MHz. Despite satisfactory performance, the design became too complicated to be recommended in its present form. It has nevertheless proved that the problem is solvable, and guidelines have been established for further experiments. In particular, I found that proximity between the 14 and 21 MHz parasitic elements (average separation less than 0.3 meter (1 foot)) had no effect on 14-MHz performance. One obvious solution would be the use of separate parasitic elements for the higher bands, stacked about 0.6 meter (2 feet) above and below the 14-MHz elements. Linear traps remain the neatest method, and from inspection of fig. 11B, it is obviously possible to increase the trap length (and hence the bandwidth) by using the entire available length for 21 MHz, in which case the 28 MHz traps would have to be accommodated within the 21 MHz traps. Experiments along these lines can be recommended to anyone who feels he has the necessary skill and patience.

Unfortunately, narrow bandwidth at 21 and 28 MHz, in comparison with a monoband antenna of the same size, appears to be a price that has to be paid for the multibanding of a “smallest-possible” beam. (It is also part of the price usually, but mistakenly, paid for the multibanding of full-size elements which, when used without traps, have very much higher radiation resistances at the higher frequencies.)

One practical point that must be stressed is the need for accurately maintaining the shape of the linear resonators. With separate feeders to each element, small changes in resonant frequency can, of course, be corrected from the shack, but any asymmetry causes the current maximum to shift toward one end of the radiator. Probably worse, a voltage will exist in the center causing current to flow in a metal boom or dielectric losses in a wooden boom if wet! Insulating a metal element from a metal boom is not a complete cure, because there is bound to be some capacitance, but it has been found preferable to bonding.

future trends

I have tried to cover the subject in all its aspects and find it difficult to envisage a future for small beams outside the guidelines presented here, but perhaps someone will take this as a challenge and come up with something really new! There is room for plenty of ingenuity in finding ways of folding elements to fit them into small spaces, and it should be possible to improve bandwidth by the somewhat mind-boggling process of folding a folded dipole! The mini-quad also has interesting possibilities and appears to lend itself to a number of options for multibanding, of which I favor use of 20-meter loops for 10 meters, and separate, stacked elements for 15 meters.

Better construction methods are needed to improve reliability and achieve closer tolerances. A completely foolproof tune-up procedure, which can be used in all situations, has yet to be evolved. More data is needed on the range over which remote tuning from the shack can be achieved without penalties; further development is needed to find the best methods for remote tuning and reversal of small three-element beams. Perhaps someone may then get around to applying these features to the big beams — which will otherwise find themselves at a disadvantage! For multibanding of beams such as those described here, I believe the future lies in improved versions of the linear resonator trap, which would ease the problem of remote fine tuning and avoid the complications of stacking.

As an alternative to neutralizing, I have tried using the 21/28-MHz element as an electrostatic screen between the 14-MHz elements, but I failed to achieve a viable system. There are difficulties, for example, in specifying exactly what is “earth,” and I have found it advisable, even with neutralizing, to insulate elements from the boom and preferably to use a wooden boom.

references


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One comment often heard on the amateur bands is that the art of homebrewing is a thing of the past. Nothing could be further from the truth. The amateur fraternity has acquired new vigor and growth, and more hams now than ever before, are "rolling their own." This is happening in spite of the fact that, again as a result of recent amateur growth, commercial manufacturers of ham gear are able to offer greater bargains in terms of performance per dollar. Yet there is a growing number of do-it-yourself amateurs who are dedicated to building all or some of their own equipment. The VTO described here is only one example of this trend. I am admittedly a homebrew addict.

concept

The constantly changing state of the art has provided a diversity of methods and circuits applicable to today's requirements. The VTO described in this article is more than just another of the garden variety. It was deliberately designed to perform well under the most demanding conditions. You may be interested in duplicating the design, or in using the basic concept to build one to suit your own particular needs.

This article is intended to provide sufficient specific information so that the design can be easily duplicated. If you need a VTO possessing superior performance characteristics, scrutinize the following checklist and decide if this construction project is for you.

1. Adjustable tuning range
2. Exceptionally clean spectrum — like that of a VCXO
3. Frequency essentially independent of load
4. A temperature compensation adjustment to cancel drift
5. Power output +10 dBm in a 50-ohm load
6. Remote tuning capability
7. Frequency independent of line voltage
8. Excellent short-term frequency stability
9. Excellent long-term frequency stability
10. Fast warm-up

After reading this article, you may wish to modify the design to fit your own needs. If so, there are certain things you should not do or need not do:

Don't use powdered-iron tuning slugs in the oscillator coil. It is particularly important to stay away from pot cores, bobbins, sleeves, toroids, and other ferromagnetic materials. Not only do these devices generally have relatively high temperature coefficients, but they possess a characteristic called hysteresis, which may cause the VTO frequency to wander from time to time for what appears to be no good reason. The permeability of powdered iron and ferrite materials is relatively sensitive to even weak magnetic fields.

Don't use a clamping diode on the MOSFET oscillator gate. It isn't necessary and only adds to the complexity. Note that the gate is grounded through the coil winding and therefore bias shift on gate 1 cannot occur.

The VTO described here should be operated at the fundamental, not a subharmonic. With the source bias used, the waveform on gate 1 is very nearly a pure sine wave, as is the waveform at the output. The circuit does not work well as a multiplier.

Don't operate the varicaps with too little bias. If the peak rf voltage exceeds the bias, rf rectification may occur which will generally disrupt performance. In the present design, varicap bias varies between 7 and 10 volts, although bias as low as 4 volts may be used if a wider band is to be covered.

Don't use a zener to provide voltage regulation for gate 2 or the drain of the MOSFET oscillator. (See fig. 1.) Use an integrated circuit regulator such as the µA723. Gate 2 voltage should be provided from a resistor divider network fed from the 723 output.

Don't mount the oscillator on a printed circuit board, since flexing or warping can cause serious stability problems. It is much better to assemble the entire oscillator on a brass or copper plate that can be mounted firmly to a shield box. Copper is preferred.

By Norm Foot, WA9HUV, 293 East Madison Avenue, Elmhurst, Illinois 60126
because it has higher heat conductivity, which helps to keep all circuit components at the same temperature. Use a small plate not less than 0.8 mm (0.032 inch) thick.

Don’t use plastic, phenolic, or fiber coil forms. The coil forms used in this VTO are Cambion part number 1536-3-1. Only the amplifier circuit coil uses the tuning slug, which is carbonyl J (green).

Do not use ceramic trimmers in the oscillator circuit. The glass-piston trimmers specified possess a low temperature coefficient and are mechanically rugged.

Do not locate the VTO circuit near heat-generating components such as power supply regulators or transformers. Remember, since the oscillator is voltage tuned, it can be located at any convenient location because it does not require mechanical linkage with the front panel. You may want to take advantage of this feature and locate it next to the mixer where it belongs.

application

This particular VTO was designed as the local oscillator for the second mixer in a communications receiver. The design requirements were very exacting. It had to operate above 30 MHz and also drive a counter to provide a digital readout. The readout constantly monitors the frequency, which means that any drift exceeding 50 Hz would be particularly annoying, not to mention the problem associated with copying sideband signals. Furthermore, the tuning range was to be adjusted to cover exactly 100 kHz plus 100 Hz overlap at each end. The drift after warmup turned out to be considerably less than originally expected, in spite of the higher-than-normal operating frequency and the use of varicaps for tuning.

tuning mechanism

Although the circuit values can be modified so that the VTO covers other frequency ranges, this particular unit tunes 30,250 to 30,350 kHz. A five-turn, linear-taper, Bourns precision potentiometer (part number 35205-417-453) is used for frequency tuning. A 6.4-mm (2.5-inch) diameter Millen fluted knob provides approximately 20-kHz per revolution, permitting ssb signals to be tuned in with relative ease. This amounts to 18 degrees per kHz. A ten-turn potentiometer can be substituted if greater tuning resolution is desired. In any event, avoid the use of a gear train to increase resolution if backlash is to be avoided.

Originally, there was considerable concern regarding the ability of the of the potentiometer to stand up

fig. 1. Schematic diagram of the voltage-tuned oscillator. All fixed resistors are 1/4-watt composition, and all capacitors are dipped silver mica unless otherwise indicated. The 470-pF feedthrough capacitors are from Spectrum Control (part number 54-794-002-741M). The varactors are TRW 4808s. Capacitor C3 is an Erie 0.7-12 pF piston trimmer, while C4 is a 1-18 pF Erie piston trimmer. L1 is wound on the Cambion 1536-3-1 coil form; L2 uses the same form with the carbonyl J tuning slug inserted as shown in the diagram.
to abuse day after day. However, the original unit showed no indication of wear after two years of almost daily and often grueling use.

**circuit details**

The VTO includes a 3N201 dual-gate MOSFET oscillator driving a low-cost 40673 MOSFET buffer amplifier. A high degree of isolation is achieved by use of the 100-ohm resistor in the oscillator drain circuit. As a result, pulling of the oscillator as the buffer coil is tuned through the oscillator frequency is less than 200 Hz. Since the buffer circuit is broadband and therefore fixed tuned, this effect has insignificant effect on drift. If the load terminals are opened or short-circuited, the change in the oscillator frequency is less than 100 Hz. Power output over the 100-kHz range is essentially constant at 10 milliwatts under load.

The coils are wound on ceramic forms with no. 26 (0.4-mm) AWG enameled copper wire and heavily doped with polystyrene Q dope. The 50-ohm output winding on the amplifier coil consists of two turns of no. 26 (0.4-mm) AWG enameled copper wire, pushed up tight against the bottom (cold) end of the tuned circuit coil. It is important that the tuning slug be positioned at the 50-ohm output end of the coil, partly overlapping both the tuned circuit and the output winding.

**power supply considerations**

Both the oscillator and its buffer are powered from a μA723 integrated circuit regulator which accepts current from a regulated 14-volt supply. The 723 reduces the 14 volts to 10 volts. The importance of this additional regulator cannot be overemphasized. Not only does it provide the required tight voltage regulation, but by its nature the output voltage is extremely well filtered. For example, when the spectral purity of the VTO was compared while using a battery supply and then the 723 post regulator, there was no noticeable difference. By contrast, without the post regulator, there was considerable hum modulation. Small changes in supply voltage cannot be tolerated because the same supply is applied to the tuning varactors.

A pair of TRW 4808 voltage-variable, varactor diodes (varicaps) are used to control the VTO frequency. These 27-pF tuning varactors are connected in the conventional back-to-back manner with the tuning voltage fed to their junction through an RC filter. The 1000-ohm tuning range trimpot is set near its midpoint so that the voltage across the 45k-ohm tuning potentiometer is approximately 3 volts. Under these conditions, the tuning potentiometer provides the varicaps with between 7 and 10 volts. Long life and reliability is assured since the potentiometer carries only 64 microamperes while the current carried by the moving arm is essentially zero. Data on the varicap is shown in fig. 2.

The varicap tuning sensitivity is 30 μV/Hz. If the spectral purity of the oscillator is to be high enough for the intended application, power supply noise and ripple must be extremely low. The μA723 is specified at 20 μV rms noise typical. This could cause a random peak-to-peak fm of 1.89 Hz. Most of this noise is eliminated by means of the lowpass filter consisting of the 47k-ohm resistor and the 1.0 μF capacitor. What little noise actually remains should consist of very low-frequency components, since the RC filter has a 20 Hz cut-off frequency. In the author's application, there is no detectable reduction in receiver output noise when a battery is substituted for the μA723 regulator.

Ripple from the 723 is specified as typically 74 dB below the input ripple. Even if the input ripple were 100 mV peak-to-peak, the output ripple would only be 2.0 μV, including the filtering action of the varicap RC filter. This translates to less than a tenth of a cycle fm.

The varactor capacity as seen by the oscillator coil varies from 3.8 to 4.4 pF. Since the total circuit capacitance. Small changes in supply voltage cannot be tolerated because the same supply is applied to the tuning varactors.

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The varactor capacity as seen by the oscillator coil varies from 3.8 to 4.4 pF. Since the total circuit capacitance.
city is approximately 90 pF, the ratio of high to low end frequencies is as follows:

\[
\frac{F_H}{F_L} = \sqrt{\frac{C + \Delta C}{C}} = \sqrt{\frac{90.6}{90}} = 1.0033
\]

where

- \(F_H\) = high-end frequency
- \(F_L\) = low-end frequency
- \(C\) = total circuit capacitance

Therefore, if \(F_L = 30,250\) kHz,

\[
F_H = 1.0033 \times 30,250 = 30,350\ kHz.
\]

Of course, the tuning range tripot can be adjusted to provide overlap at each end of the tuning range if that is desired.

The frequency ratio formula can also be used for different frequencies and tuning ranges. I suggest that the circuit capacity be scaled in proportion to wavelength. A VTO operating between 5.0 and 5.1 MHz, for example, could use a circuit capacity of about 500 pF. The varicaps would each be 150 pF, while \(C\) would be approximately 20.2 pF, which requires a tuning voltage range of about 6.4 to 10 volts.

**tuning linearity**

While the tuning potentiometer is linear, the varicap tuning characteristic is not. As a consequence, the frequency tunes slightly faster at the low end of the range than at the high (see fig. 3). However, the departure from linearity is small. Based on the manufacturer’s data, and by measurement, the number of kilohertz per revolution at the low end is 22.4 compared with 18.2 at the high end. This nonlinearity might be objectionable if it were necessary to employ a mechanical frequency indicator, such as a planetary mechanism. In my application, the VTO drives a programmable digital counter in which case the small nonlinearity is of no consequence.

**temperature compensation**

Considerable effort went into developing a variable temperature compensating circuit for this VTO. While the idea is not new, it lends itself well to this particular application. The compensation circuit is simple and easy to adjust.

A piston trimmer is connected in series with a negative temperature coefficient ceramic capacitor. The coefficient was intentionally made larger than needed. The effective circuit coefficient is adjusted by tuning the piston trimmer, \(C_4\). The change in frequency which resulted is compensated for by adjusting \(C_3\). There is a particular setting of these two capacitors which will provide zero temperature coefficient, therefore zero oscillator drift after warm-up.

Based on manufacturer’s data, the voltage-tuned varactors have a temperature coefficient of 250 to 300 ppm/degrees C. Of course, most of the other circuit elements exhibit positive coefficients as well, but their coefficient is generally less — in the order of 50 to 100 ppm/degrees C. Dipped silver-mica capacitors were found to be entirely satisfactory for use in the oscillator and amplifier tuned circuits.

A series of tests was performed using NTC capacitors with various temperature coefficient values. Finally, a 51-pF, N330 capacitor was selected. An additional 24-pF, NTC 1510 capacitor was added in parallel to bring the circuit temperature coefficient within range of \(C_4\). \(C_4\) is a key circuit element because it allows the effective temperature coefficient to be adjusted, and therefore the amount and direction of oscillator drift. The proper setting was found after a number of drift tests were performed over a period of a few hours using different settings of \(C_3\) and \(C_4\). The results of these tests are shown in fig. 4.

After a period of about three months, and again a year later, the drift was checked. It was found that no change in compensation had taken place. The circuit gives you a real good opportunity to check the other fellow’s frequency stability!

After two years of drift-free use, and strictly out of curiosity, the original 3N201 oscillator MOSFET was replaced with another made by a different manufacturer. It was found that the frequency was slightly higher and needed trimming. Subsequent drift tests showed the circuit to be over-compensated. A new set of drift tests were run and the overall temperature coefficient quickly brought back to zero.

If the drift that occurs after turn-on from a cold start is annoying, or if the application requires absolutely zero warmup drift, I recommend that the drift correction circuit described by PA0KSB be used. A new set of drift tests were run and the overall temperature coefficient quickly brought back to zero.

![fig. 3. Chart of the voltage-tuned oscillator's linearity as the oscillator tunes from one end of its range to the other.](image-url)
There was absolutely no warmup drift. It was necessary, however, to include the up-down tuning push buttons recommended to periodically correct for calibration errors inherent in the design. Since this requirement was more of a nuisance than the warmup drift, the circuit was not permanently installed. From a practical standpoint, about the time the 6146s are warmed up and ready to go, the VTO has stabilized! It was also interesting to note that if the VTO design described here were to be modified for operation at 5.0 MHz, the warmup drift would be only about 100 Hz.

Excellent short-time stability was achieved by enclosing the VTO circuit in a metal box. Also, a brass cup heatsinked to the chassis was mounted over the 3N201 transistor and its socket. This protects the transistor from drafts to which it is particularly susceptible, and forces the ambient temperature in the vicinity of the transistor to track that of the entire circuit. This combination was found to be very effective.

One of the most frustrating problems associated with checking out a new oscillator is being able to separate out the various causes of drift. They include the effective temperature coefficient of the overall electrical circuit, the effects of voltage variations, and mechanical effects. If the oscillator varies or jumps when the oscillator chassis is tapped sharply with a screwdriver, you have mechanical problems. Don't try to temperature stabilize the circuit until it is mechanically sound. It is a good idea to run the temperature of the circuit up and down several times using a hair dryer to reduce mechanical strains before checking drift.

If you use the regulator circuit suggested here, you should be able to set the line voltage as low as 100 volts before detecting any oscillator drift. Long-term drift tests should be performed by running the circuit as it would normally be operated. The use of a hair dryer will probably prove only that the temperature compensating circuit has a thermal lag.

Short-term drift and low-frequency shot noise effects can be determined by tuning in WWV (assuming the VTO is used as the local oscillator of the receiver). Using the BFO, set the tuning for a convenient heterodyne (500 Hz for example), and then acoustically zero beat an audio oscillator and speaker combination with the receiver output. Listen for any warbling and jumping. If you use a low-beat note, you should be able to hear and detect frequency variations as small as ±2 Hz by reading frequency changes from the audio oscillator dial. If the oscillator is working properly, random variations should not exceed ±5 Hz at 30 MHz, or ±1 Hz at 5 MHz, and there should not be any jumping. It is interesting that the technique described here allows you to detect variations as small as one part in $10^7$ or better.

references

bibliography
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More Details? CHECK — OFF Page 118
The 555 timer has become one of the most popular ICs used by both industry and hobbyists alike. Since its introduction by Signetics Corporation in 1972 (as the NE555), every major IC manufacturer has produced a 555 equivalent. Variations have been developed by Signetics and others, primarily for specific applications (NE556, NE558 quad, XR2250 programmable).

Circuits using the 555 show up in virtually every electronics-oriented magazine. The popularity of the 555 is partly due to its versatility and partly because of its cost (generally less than 50 cents). These published circuits use the timer to generate ramps and time delays, detect missing pulses, act as oscillators, and a host of other applications. The list of specifics can become quite extensive. Most of these applications now appear as “cookbook” circuits with little detail on how the design was accomplished and what variations a builder of that circuit can expect.

This article was written to help you get a better understanding of the 555 and how its eight pins react to external components. With this information, you can get a feel for how most 555 applications work. For the more creative, I'll outline the basic design rules for using the 555. (Information on the 555 also applies to the 556 dual timer.)

Fig. 1 shows the block diagram of the 555 timer. The basic components consist of an output driver, a control flip-flop, and two voltage comparators. The flip-flop drives the discharge transistor and output driver. The state of the flip-flop is controlled by the reset pin (pin 4) or one of two comparators. One comparator is controlled by the voltage on the trigger pin (pin 2) and the other controlled by the voltage on the threshold pin (pin 6).

These comparators have separate reference points which are controlled by the three-resistor divider from $V_{cc}$ to ground. The resistors are all of equal value (5-k ohms). The reference voltage for the threshold comparator is thus $2/3$ of $V_{cc}$ and is also available on pin 5, the control voltage pin. The reference voltage into the trigger comparator is set at $1/3$ $V_{cc}$ by the divider network (Note: it is also one half of the voltage on the control voltage pin).

The output (pin 3) has the capability of providing 200 mA. Because of the totem-pole structure, this

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current supply is available whether the output is high or low. Additionally, when the output is high, the discharge transistor is turned off.

monostable operation

To help understand the action of the comparators it is best to show the operation of the 555 in one of the basic timing modes. The monostable configuration of the timer requires only two external components. It is shown in fig. 2A (the capacitor on pin 5 is optional).

A resistor (R_A) is connected from V_CC to the discharge pin (pin 7) and threshold pin (pin 6). A capacitor is then connected from pins 6 and 7 to ground. The values of R_A and C will determine for how long the output remains high.

Initially, the output is low and the discharge transistor is turned on. This essentially shorts the timing capacitor C. Therefore, the threshold comparator input is at zero volts. When the trigger comparator receives a trigger that is less than 1/3 V_CC, it causes the flip-flop to switch. The output goes high and the discharge transistor turns off. When this happens, the timing capacitor starts charging to V_CC via R_A. As soon as the charge on the capacitor reaches the threshold voltage level (2/3 V_CC), the threshold comparator triggers the flip-flop. This drives the output stage low and turns on the discharge transistor, which discharges the timing capacitor, and the timer is back to its initial condition. Fig. 2B shows the waveforms on the timer during this time.

triggering

Due to the nature of the trigger circuitry, the trigger comparator will trigger the flip-flop whenever the trigger voltage drops below 1/3 V_CC. For proper timing, the trigger level must return to a voltage level greater than 1/3 V_CC. This is because the trigger comparator has overriding control of the flip-flop. Should the trigger input be held low for a period longer than the timing cycle, the output will remain high, without regard to the voltage on the threshold comparator.

reset function

The 555 has a reset pin for applications that require an abort signal. That is, it may be necessary to interrupt a timing period or to inhibit a trigger. The reset control (pin 4) is normally high. If the reset is held greater than 1.0 volt, the 555 will function normally. However, if pin 4 is held below 0.4 volt the output will be held low. When the timer output is high and the reset goes low, the output will turn off immediately. If the reset pin is between 0.4 and 1.0 volt, the timer is in no man's land; some devices may reset, some may not. When using the reset function, it is important to have the reset less than 0.4 volt. This level is guaranteed to cause the timer to reset. To prevent possible noise spikes from causing an unwanted reset, pin 4 is normally connected to the V_CC pin when not being used.
calculating pulse width

When describing the operation of the 555 timer in the monostable operation, I mentioned that the values of \( R_A \) and \( C \) determined the time the output stayed high. The voltage on \( C \) must charge to \( 2/3 \) \( V_{cc} \). Fig. 3 shows the normal RC charge curve. As shown, the voltage rises from 0 to 100 per cent in five RC time constants \( (\tau) \). The 66.6 per cent \( (2/3) \) point occurs at \( 1.1\tau \), or

\[
T = 1.1 RC
\]

where

\( T \) is in seconds,
\( R \) is in ohms
\( C \) is in farads

changing the pulse width in the monostable mode

The threshold control voltage on pin 5 is an important feature of the NE555 timer. By imposing a voltage on pin 5, the internal comparator reference levels, the timing can be varied. For example, if one-half the supply voltage were placed on pin 5 (normally held at \( 2/3 \) \( V_{cc} \)), the timing capacitor would charge only to one-half \( V_{cc} \) before the threshold trip level is reached. The time required to charge the capacitor can be found by observing how many time constants it takes for the capacitor to charge to 50 per cent using the equation

\[
T = \ln \left(1 - \frac{V_5}{V_{cc}}\right) RC
\]

or by using Fig. 3. With the voltage on pin 5 equal to 0.5 \( V_{cc} \), the new timing equation is

\[
T = (\ln 0.5)RC = 0.693RC.
\]

You must remember that the trigger threshold is also affected by changes on the voltage level at pin 5. In the example above, the trigger threshold is now lowered to one-quarter \( V_{cc} \) (halfway below the voltage on pin 5). In order to trigger, the trigger pulse must now go below one-quarter \( V_{cc} \). This feature of the timer opens a multitude of application possibilities such as pulse width modulators, voltage-controlled oscillators, and so forth.

As can be seen, any variations on pin 5 will cause a change in timing. For that reason it is recommended that a small bypass capacitor (about 0.01 \( \mu F \)) be used on pin 5. This will increase noise immunity of the timer to high frequency trash that could cause timing errors by modulating the threshold trip levels.

astable operation

To configure the 555 timer in the astable or oscillatory mode requires only a slight modification to the monostable configuration. Fig. 4A is the schematic of a 555 timer in the basic astable mode. It requires two resistors and a timing capacitor. Assume that the charge on the timing capacitor is charging toward \( V_{cc} \) through \( R_A \) and \( R_B \). The output is high and the discharge transistor is off. When the capacitor charges to the threshold trip level \((2/3 V_{cc})\) the output goes low and the discharge transistor turns on, shorting pin 7 to ground. The capacitor now begins to discharge through \( R_B \) toward ground. But as soon as the capacitor discharges to \( 1/3 V_{cc} \) (the trip level of the trigger comparator), the output again goes high, the discharge transistor turns off, and the

fig. 4. Schematic diagram of the 555 connected as an astable multivibrator and the waveforms at different points in the circuit (B). In applications where a 50 per cent duty cycle is required, the circuit shown in (C) can be used. In this case, \( CR_1 \) effectively shorts out \( R_B \) during the capacitor charge time.
capacitor starts charging through \( R_A \) and \( R_B \) again, capable of switching high levels of load current. As a result, large current spikes can develop on the supply line. This momentary loading effect can cause a degree of timing error because of changes in charging current. It can also cause noise glitches and false triggering in TTL circuits. To eliminate this phenomenon, it is necessary to bypass the supply line to ground with a capacitor. The size of the bypass required generally depends on the load to the timer. Values range from 0.1 to 10 \( \mu F \), or more. The bypass capacitor should be as close to the device as possible.

**Calculating Frequency**

The time required to charge from 1/3 to 2/3 \( V_{cc} \) is 0.671 \( RC \). The charge path is through \( R_A + R_B \). This is the period of time that the output is high, or

\[
T_1 \text{ (High)} = (R_A + R_B) C \tag{2}
\]

During the discharge time the output is low. The time required to discharge from 2/3 to 1/3 \( V_{cc} \) is also 0.671 \( RC \). But the discharge path is through \( R_B \) only, so

\[
T_2 \text{ (Low)} = 0.671 \ (R_B) C \tag{3}
\]

The total period of oscillation is then:

\[
T_{TOT} = T_1 + T_2 = 0.671 \ (R_A + 2R_B) C \tag{4}
\]

and the frequency

\[
F = \frac{1}{T_{TOT}} = \frac{1}{0.671 \ (R_A + 2R_B) C} = \frac{1.49}{(R_A + 2R_B) C} \tag{5}
\]

The duty cycle is given by:

\[
D = \frac{R_A}{R_A + 2R_B} \tag{6}
\]

Since the charge and discharge paths are different, the duty cycle cannot be less than 50 percent. In some applications, this may cause a problem. With a slight modification to the basic circuit, \( R_A = R_B \). In this circuit, \( CR_1 \) shorts \( R_B \) and the charge time is 0.671 \( R_A C \). The discharge path is still through \( R_B \) and does not change. The series diode, \( CR_2 \) is optional to match the charge and discharge paths. With the diodes in the circuit the formulas for frequency and duty cycle become:

\[
F = \frac{1.49}{(R_A + R_B) C} \tag{7}
\]

\[
\text{Duty Cycle} = \frac{R_B}{R_A + R_B} \tag{8}
\]

With this configuration, the timer is capable of generating duty cycles from 5 to 95 per cent.

**Supply Voltages**

In selecting supply voltages, note that the 555 timer is guaranteed to operate from 4.5 volts to 16 volts. Because the threshold levels work on a ratio, the supply voltage will not change the timing calculations. The timer has a “totem pole” output structure capable of switching high levels of load current.

Bypassing the control voltage pin (pin 5) is generally considered good design practice. As mentioned earlier, the value is not critical, but the capacitor should be as close to the device as possible. Typically, the bypass capacitor is 0.01 \( \mu F \).

**Capacitors**

The timing capacitor size is virtually unlimited; but, the type of capacitor is important. Ceramic disk capacitors are usually unsuitable. They generally are not stable enough to operate in an RC timing circuit. Electrolitics usually have very high leakage rates and would cause drastic timing errors. The Signetics data sheet lists several acceptable types of capacitors, including silver mica, mylar, polystyrene, and tantalum. The smaller the timing capacitor used, the more apparent effect stray capacitance can have on the timing. The larger the capacitor, the more expensive the capacitor is. This is particularly true of the low-leakage types. For long delays, the capacitor has to be very large, but there are ways of getting long delays without using big capacitors.

**Resistors**

There are certain maximum and minimum values
In this example of ac triggering, the trigger pulse must be less than the duration of the output pulse from the timer. The duration of the trigger depends upon the RC time constant of the components.

The duration of the trigger pulse depends upon the RC time constant of the components. The threshold comparator requires 0.25 μA of current to trip the output. Considering worst case, the resistor must be able to supply 0.25 μA of current for the comparator and still charge the capacitor to 2/3 \( V_{CC} \). To calculate the maximum resistance, the IR drop must not exceed 1/3 \( V_{CC} \) with 0.25 μA current flow, or

\[
IR \text{ drop} = V_{cc} - V_{CAP} = V_{cc} - (2/3)V_{cc} = 1/3 V_{cc}
\]

The maximum resistance between \( V_{cc} \) and pin 6 is defined as

\[
R_{MAX} = \frac{V_{cc} - V_{CAP}}{\text{threshold current}} \quad (9)
\]

With a 15-volt supply

\[
R_{MAX} = \frac{15V - 10V (2/3)V_{cc}}{0.25 \times 10^{-6}} = \frac{5}{0.25 \times 10^{-6}} = 20 \text{ megohms}
\]

With a 5-volt supply

\[
R_{MAX} = \frac{5V - 3.33V}{0.25 \times 10^{-6}} = \frac{1.67}{0.25 \times 10^{-6}} = 6.6 \text{ megohms}
\]

When using large resistors, capacitor leakage can cause larger timing errors because it represents a larger percentage of the total charge current available. Excessive capacitor leakage current will also cause an IR drop and not allow the threshold comparator to reach 2/3 \( V_{CC} \).

The minimum value of resistance is determined by the current that the discharge transistor can supply. The discharge transistor is internally current limited to about 35 mA. The discharge transistor must supply two loads. The first is the current through \( R_A \). This should be reduced so that the second path into the capacitor (or through \( R_B \)) carries most of the load. As a general rule, \( R_A \) should not be less than 5k ohms. \( R_B \) should not be less than 3k ohms.

**control voltage changes**

As mentioned earlier, the control voltage pin (pin 5) is normally bypassed to ground by a 0.01-μF capacitor. By imposing a voltage on this pin, it was shown that the timing can be changed since the threshold level is changed. The voltage level on pin 5 can be lowered to about 45 per cent of \( V_{CC} \) when operating the timer in the monostable mode. The trigger level is also changed because it will equal one half the voltage on pin 5, and a certain minimum volt-

---

**fig. 7**

**fig. 8**

Long time delays can be generated by cascading timers with other count-down devices. In this example, the basic timer, \( U_1 \), runs at about 0.028 Hz, is divided down by 100, and then used to trigger another timer configured as a monostable. The output is an eight-second pulse, once every hour.
age level must be maintained on the trigger comparator reference. If the voltage drops below 45 per cent on pin 5, the timer becomes more sensitive to noise because the trigger voltage nears ground. At the other extreme, the control voltage on pin 5 should not exceed about 90 per cent of $V_{cc}$ since the timing capacitor must charge to $V_{cc}$. Exceeding this voltage range in the monostable mode may cause timing error, false triggering, or other problems.

In the astable mode, the voltage levels changes on the control voltage pin can be used to change the oscillating frequency about $\pm 25$ per cent and still remain linear. Changes greater than this will still cause the frequency to change, but linearity decreases due to the RC timing circuit.

Before examining some specific circuits, there are several idiosyncrasies of the timers that can cause problems unless you are aware of them.

**Temperature.** The timer exhibits a small negative temperature coefficient (50 ppm/C°). This can cause small timing changes in the monostable mode and frequency drift in the oscillator mode. In critical applications, R/C values can be selected which have positive coefficients, with the net result a lower drift. Since the astable mode relies on both trigger and threshold levels, the drift from temperature is usually higher than when in the monostable mode. An important point to remember when working with the timer (for that matter any IC) is that power dissipation of the device should never be exceeded. Power causes heat, and excessive heat will destroy the device. The 555 can handle about one-half watt of power at room temperature. This power rating is lower when the device is operated at higher temperature.

**Output.** If a negative voltage (with respect to pin 1) is applied to the output of the timer, the 555 could latch up. This can happen when the 555 is used to drive an inductive load, such as a relay. To prevent this inductive kick back from latching the timer, a diode in series with the output should be used. Fig. 5 shows a schematic of a timer being used to drive an inductive load.

The output drive capability of the timer is 200 mA. Because of the output structure's high current capabilities, and fast rise and fall times, the timer exhibits crossover distortion. This glitch can cause false triggering of TTL circuits. By providing a capacitive load, the timer output is slowed to a point that the glitch does not occur. A capacitor of about 1000 pF from the output to ground will eliminate any false triggering of the TTL circuit (see fig. 6).

**Triggering.** I've stated that in the basic monostable mode the trigger must go below $1/3 V_{cc}$ (or one half
that is a modification of the monostable operation. This circuit provides an output after some predetermined time. Initially, the trigger input to the base of Q₁ is high, causing Q₁ to conduct. Pin 2 of the timer is low and the output (pin 3) is high. When the input goes low Q₁ turns off, allowing the timing capacitor to charge. When C charges to 2/3 Vcc, the output of the timer goes low. The output will stay low as long as Q₁ is turned off. The reset pin can be used to keep the output low if required; otherwise it should be tied high.

Sequential timing. Fig. 10 shows another type of delay circuit. This is a sequential timer. The output of the first timer is used to trigger the second. The control-voltage pin is used to vary the sequence time, but the ratios remain the same. The timing diagram shows how the pulse width is reduced as the control voltage pins are lowered.

Delayed triggering. When power is first applied to the circuit shown in fig. 2 there is a chance that the trigger voltage on pin 2 will be lower than 1/3 Vcc and the timer will trigger. This may not be desirable. To prevent this initial trigger when power is turned on, the reset circuit can be modified as shown in fig. 11.

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**Example circuit**

The following circuits illustrate the use of the 555 timer (or 556 where two timers are used).

**Long time delays.** Because of the limitations of resistor size, long time delays (times greater than one hour) can be difficult to achieve using the basic circuit. One method of getting long delay times is shown in fig. 8. This particular circuit provides a positive 8-second pulse once each hour. U2 receives a trigger once each hour. The output of U2 is set for the desired pulse output by R₄ and C₁₀. R₃ and C₇ provide the ac-coupled trigger. The 7490 counters are set to divide by 100. To provide one cycle per hour, the input must be clocked at a frequency of 1/36 Hz; U₂ is then set to oscillate at 0.028 Hz. Note that C₄ provides a deglitch filter on the output of U₂. Because the timer must interface with TTL counters, the timers and counters use 5-volt supplies. The entire circuit can also be built by using a single 556 dual timer and one 74390 dual-decade counter. By changing the divider network or frequency of U₁, it is possible to get an almost infinite combination of pulse outputs.

**Simple time delay.** Fig. 9 shows a simple circuit

---

*fig. 11. With the capacitor attached to the reset input, the timer cannot produce an output when power is initially applied. After Cᵣ has charged, a trigger will produce the desired output pulse.*

---

*fig. 12. In the missing pulse detector, each trigger shorts the charging capacitor, preventing the output pulse from appearing. If a pulse is missing, the capacitor will charge to the required level, producing the output pulse which indicates the missing pulse.*
fig. 13. The modulation input is applied to the control voltage input, with the width of the output pulse varying as the amplitude of the modulation.

Pin 4 is held below 0.4 volt since $C_R$ is fully discharged. When power is applied, the timer is held reset until $C_R$ charges to above 0.4 volt.

**Missing pulse detector.** Fig. 12 shows an NE555 timer hooked up as a missing pulse detector. The trigger input also drives the base of a transistor. When a trigger occurs, the transistor conducts and shorts the timing capacitor before the timer can time out. If an input trigger is missing, the timer will time out and pin 3 will go low indicating a missing pulse. The values of $R_A$ and $C$ are set to be slightly longer than the period of incoming pulses.

**Pulse width modulator.** Fig. 13 shows a circuit which uses the control voltage pin to modulate the pulse width of an incoming clock signal. As the voltage on pin 5 varies, it changes the threshold level of the internal comparator and the pulse width of the output changes.

The examples covered were used to illustrate some of the possibilities of the 555 timer. The uses for this handy little IC could well take a whole book to illustrate. With an understanding of the basic operation, you can analyze timer circuits and design a circuit for your particular need. Who knows, you may come up with an original application and add your name to an evergrowing list of 555 timer circuit designers.
receiver digital display

A digital display for your receiver featuring 100-hertz readout, single frequency input, and provisions for forward or reverse tuning VFOs.

The addition of a digital frequency display to a receiver goes a long way in enhancing operator convenience. It eliminates squinting at the fine markings found on the usual mechanical dial; and, at the same time, it provides an accurate frequency readout across an entire band without any need for recalibration between band ends. A bright digital display is not only easy to read, it also adds a touch of class to a perhaps otherwise ordinary station.

The circuit described in this article provides a stable, four-digit readout that includes the 100-kHz digit through the 100-Hz digit and will accommodate forward or reverse tuning VFOs at the flick of a switch. The resolution is greater than is usually provided by the typical receiver dial and is handy for returning to a particular frequency in a crowded band. In the interest of economy and simplicity, I did not include the MHz digits; they are easily read from the receiver band switch.

The technique used here requires only a single connection to the receiver's VFO, which, in many modern receivers, is usually available at a connector on the rear panel, making it unnecessary to tamper with the receiver in any way.

theory of operation

In the case of a backward-tuning VFO, the frequency to be displayed is equal to a fixed frequency minus the VFO frequency. All that needs to be done is store the fixed frequency, subtract the VFO frequency, and display the result. In the case of a forward-tuning VFO, store the complement of the fixed frequency (subtract the fixed frequency from zero), add the VFO frequency, and display the result. These additions and subtractions are easily accomplished by using up/down counters and sequentially gating the frequency to be added to the up input and the frequency to be subtracted to the down input, each for the same fixed interval. The fixed frequency is usually crystal controlled and changes very little with time. Thus, measuring and storing it once during an initial calibration is usually adequate.

The price paid for the single interconnection and the simpler circuit is the need to perform a single calibration step each time a new band is used. The operator, using the receiver's internal calibrator, tunes to the bottom edge of the band and presses the calibrate (RESET) button on the display. No further adjustments are necessary. Depressing the

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CALIBRATE button stores the VFO frequency and displays all zeros. The operator may now tune up the band in the normal manner and read the display as the frequency above the bottom edge of the band. On the twenty-meter band, for example, 14.0253 MHz would be displayed as 025.3, with the 14 read from the receiver bandswitch. This calibration requires only a few seconds and is something that should normally be done anyway.

The circuit will handle either forward- or reverse-tuning VFOs, but, for the purpose of this explanation, assume a reverse-tuning VFO. Pressing the CALIBRATE button will clear the 74192 up/down counters and cause them to count the VFO frequency in the up direction for 100 milliseconds. The result is then stored in the 7475 latches. All subsequent count cycles will start at the number stored in the latches and count down for 100 milliseconds. Suppose the calibrate button were pressed when the receiver was tuned to exactly 14.0000 MHz. As a result, 5000 would be stored in the latches because the VFO was at 5.5000 MHz. Tuning the receiver to 14.0253 MHz would cause the receiver VFO to be 25.3 kHz lower, or 5.4747 MHz. The display counter would now start at 5000 and count down for a period of 100 milliseconds. The resulting count would be 5000 minus 4747, or 0253. By permanently enabling the decimal point to the left of the least significant digit (LSD), the display would show 025.3. Notice that by counting for 100 milliseconds, 54747 cycles are fed to the counters, but, since it has only four displays, the figure 5 is lost.

For a forward-tuning VFO, the sequence switch is placed in the reverse position. Pressing the calibrate button will then cause the first count cycle to be a down count with all subsequent cycles up counts.

Inside view of the assembled digital display. The smaller perf board is mounted right behind the bezel.

**circuit description**

One section of U1 serves as a 100-kHz crystal-controlled oscillator (see fig. 1). A divider chain, consisting of U2, U3, U4, and U5, drives a 7490 decade counter, U6 with one clock pulse every 100 milliseconds. The output of U6 is decoded by U7, with output pins 1, 3, and 4 each sequentially going low for 100 milliseconds of each measurement cycle. A jumper between pins 2 and 8 of U6 causes it to have only four output states rather than ten.

Assume the sequence switch is in the normal position. Pressing the CALIBRATE button causes the sequence of the next cycle to be clear, count up, update. This clears the counters to zero, counts for 100 milliseconds, and stores the final count in the 7475 latches. All succeeding cycles follow the sequence load, count down, latch. This loads the number stored in the latches into the counters, counts down for 100 milliseconds, and latches the final count into the displays.

U11 synchronizes the beginning and end of the calibrate cycle with the state of U7. Pressing S1 causes pin 5 of U11 to go high. As a result pin 9 goes high at the end of the current cycle and causes the next cycle to follow the calibrate sequence, in addition to resetting the first half of U11. At the end of the calibrate sequence, pin 4 of U7 goes high, resetting the second half of U11. U10 is the equivalent of a double-pole, double-throw switch and determines the count sequence, either up/down or down/up.

The remaining sections of U11 amplify and square-up the VFO signal. Three 4049 CMOS inverters are used. The first is biased to serve as an amplifier and functions well to above 6 MHz. This is fine for the large number of receivers with 5 to 5.5 MHz VFOs; but if your VFO operates at frequencies above this, a
fig. 1. Schematic diagram of the complete digital display. C7 and C8 are silver-dipped micas; C9 is a small ceramic trimmer. Substitution of the LS-series ICs for the 7490s and 74192s will greatly reduce the power supply current permitting the use of a small regulator and transformer.
Power Supply

Schematic diagram of the power supply. Current consumption is such that a small voltage regulator could be used (LM309K).

U12 operates as a straight decade divider and provides a stable readout with no last digit flicker by acting as an undisplayed counter stage. The display may also be used as a conventional up counter or a straight down frequency counter by merely throwing the offset switch, S2, to the off position.

Construction

My display was built in a 6.4 × 18 × 15 cm (2½ × 7 × 6 inch) Ten-Tec enclosure using a commercial bezel assembly available from Digi-Key. No printed circuit board is used. Instead, I used two pieces of 2.5-mm (0.1-inch) grid perf board, one for the counters and displays, and the other for the timing and control circuits. All ICs are mounted in wirewrap sockets which are in turn cemented to the perf board. Most of the interconnections were made using standard number 30 AWG (0.25-mm) wire-wrap wire and an inexpensive hand-wrapping tool. The only exception is the power buses, which are number 18 AWG (2-mm) solid bare wire mounted on push-in solder terminals and run on either side of each row of IC sockets. Bypass capacitors (0.01 μF) were used liberally throughout the circuit. The voltage regulator is mounted on the rear of the cabinet using thermal compound, and several vent holes were drilled in the cabinet sides to remove internal heat.

The displays shown in the schematic are very compact and convenient, as they include decoders and latches, but they tend to be rather expensive if they must be purchased new. A functionally equivalent and much cheaper arrangement could be made using 7475 latches, 7447 decoder/drivers, and common-anode, seven-segment LED displays, but a slightly larger cabinet would probably be required (see fig. 2).

The 100-kHz crystal I used is of unknown lineage and different crystals may require a slight adjustment of component values in the oscillator circuit. The 8000-pF power supply filter capacitor could probably be reduced to half the value without any problems and a 1.5-amp regulator substituted for the LM 323, as the current requirement at 5 volts is about 900 mA, a bit less than I originally anticipated.

Calibration

Clip a piece of insulated hookup wire to pin 2 of U1 to act as an antenna and place it near a receiver tuned to WWV. Zero beat WWV with the trimmer and you're ready to go. If you have a good frequency counter, adjust the trimmers for exactly 100 kHz.

I hope that the ideas and circuits presented here prove to be informative and may perhaps encourage the construction of improved designs using hardware that will appear in the future. Counter latch combinations and dividers with a modulus greater than ten could serve to simplify construction considerably. For anyone attempting construction as shown, I would be pleased to supply advice or answer questions. Please include a stamped, self-addressed envelope with all inquiries. Now that I have had the use of this display for several months, I wish I'd built one years ago.

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new approach
for a
1-MHz oscillator

A reversion to basic oscillator concepts, this oscillator was built with optimized circuitry for each stage.

Most of the oscillators built by Amateurs use a single transistor to perform several functions. However, the basic principles of operation are somewhat a mystery. The original analysis of the oscillator, done in the 1920s, showed that it could be divided into separate stages (see fig. 1). By actually building separate stages, you can obtain a better understanding of oscillator operation. This article presents an oscillator built using those basic concepts.

circuit description

The RCA CA3028 differential amplifier, shown in fig. 2, is used as an amplifier and limiter. Feedback, for the actual oscillator, is the limited sinewave from the pin 6 output. Note that this is a noninverting configuration. The amount of feedback is controlled by the 5-kilohm pot connected to pin 6. Feedback is injected into the quartz-crystal resonator. Because the crystal acts as a bandpass filter, the output will be a sinewave, in this case at 1 MHz. However, the crystal cannot be connected to the limiter without a penalty. The input impedance of a limiting differential pair is nonlinear. To overcome this problem, the crystal must be isolated from the limiter. This is done by connecting a common-base amplifier between the resonator output and the amplifier input. The com-

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fig. 2. Schematic of the 1-MHz oscillator built using separate stages for each function rather than combining functions into a single device. All resistors, except $R_1$, can be $1/8$ watt; $R_1$ is a $1/4$-watt resistor. The power requirement is $36$ mA at 15 volts.

mon-base amplifier provides a constant impedance load for the crystal. Because the output of the limiter is fixed, regardless of its input, and because the resonator also sees a constant impedance, the sine-wave at the collector of the common-base stage has should oscillate. If it does not, increase the feedback until oscillation occurs at 8 volts. After the feedback is correctly adjusted, connect the output to a counter and adjust the crystal frequency with the small trimmer.

fig. 3. Diagram of a 40-meter LC-tuned VFO built using the same guidelines as the crystal oscillator.

a constant amplitude. This signal completes the loop by driving the limiter. The output is taken from pin 8, the unused inverting output. The output from the CA3028 is a limited sinewave. To make the output a clean square wave, an SD211 DMOS FET was used. The FET is driven into saturation, squaring the signal.

construction and adjustment

The oscillator was built on a piece of perforated board and housed in a 10 x 5.4 x 4.3 cm (4 x 2-1/8 x 1-5/8 inch) mini-box. Adjust the 5-kilohm variable resistor until the circuit goes into oscillation. Next, reduce the power supply voltage to zero and slowly increase the voltage. At 8 volts, the circuit

The output is a 4-volt peak signal with a rise time of 100 ns. The variation in frequency, at 1 MHz, is 0.5 Hz over a 1-second interval. This is due to the quality of the crystal and thermal effects. This variation can be reduced if a higher-quality crystal is used. Also, the environment can be temperature controlled with an oven or the use of insulation. The final frequency error was adjusted to 0.05 Hz. My present application for the oscillator is a time base in a frequency counter. However, you could modify the circuit to act as an LC-tuned VFO (see fig. 3). Since the SD211 FET amplifier is inverting, feedback is taken from pin 6, the inverting output.
novel method for matching input impedance of grounded-grid power-amplifier tubes

When I was thinking about driving a tube power amplifier with my 20-watt transistorized ssb general-coverage transceiver, I noticed that uniform matching from the 50-ohm output of the transceiver into the amplifier between 1.8 and 30 MHz might present some problems. A circuit with a grounded cathode tube wasn’t too attractive, because of the required neutralization and the various voltages needed for a tetrode or pentode. So I decided to use one of the modern grounded-grid tubes of the 8873-8875 series. Their input impedance is in the vicinity of 100 ohms, but since it’s a dynamic input, impedance varies as drive level changes.

It’s therefore highly recommended to use at least one tuned circuit between the exciter and the cathode connection of the tube to store energy and stabilize the dynamic input impedance. Since this tuned circuit requires retuning when the frequency is changed more than 10 per cent, such an arrangement would be highly impractical.

Note that this energy-storing circuit, although sometimes used in a pi configuration, does not really suppress exciter harmonics. Standard practice is to have enough harmonic suppression already in the exciter, and a 50-ohm output is provided from the exciter.

matching network

With these things in mind, and using a calculator to determine elliptical filters, I constructed a matching network that provides perfect matching and ideal energy storage between 1.8 and 30 MHz (fig. 1). This is a bandpass filter, which has Chebyshev response in both bandpass and stop-band areas. It provides a constant impedance match over the frequency range of between 1 MHz and 60 MHz. This is important, because even high-order harmonics must be properly terminated, otherwise high transient voltages can be developed in the driver stages.

If you consider the fact that the frequency range between 1.8 and 30 MHz, on the basis of individual filters, must be split into at least seven segments (and assuming three components per segment), the total number of segments would be twenty-one, while here only fourteen are required.

By Ulrich L. Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

ham radio
the key-toggle

Key-toggle, another form of fast break-in, featuring simple but effective hand-key control of the transmitter

I wonder whether some psychologist won’t someday discover that ham radio owes much of its appeal to the fact that you can say your piece in silence and the other guy can’t interrupt until you’re through. Even in most VOX systems, the other person cannot force interruption. Perhaps this is one of several reasons why “full break-in” CW operation has not yet become prevalent. It is not difficult technically, but there are still those of us who choose not to use it.

For handling commercial traffic on a clear channel, full break-in is perhaps best; but for most other activities, coupled with reasonable and unselfish operating practices, I prefer my key-toggle mode. To come on the air, you just start keying. To switch to receive, you simply hold a dash a little longer than normal (time adjustable). Key-toggle can be just about as fast as full break-in, but the other guy can’t break you. For example, when you are winding up a transmission, the last letter normally sent is dah-di-dah. With key-toggle, it’s dah-di-dahah, click, click. The first click switches your transmitter’s plate supply off and the second click, milliseconds later, switches the antenna relay and opens the receiver (optional). You are now in receive mode until you start keying again. When you hit the key, once again: click, click. But this time the antenna relay is switched and the receiver muted first; then, milliseconds later, the transmitter’s plate supply comes on and you are ON again. No combination of letters or long pauses will change the relays, switching you back to the receive mode. Only a long dash will toggle your station back to receive.

circuit description

Major components, as seen in fig. 1, include a 741 operational amplifier configured as a dual time constant integrator, a 711 comparator, a 7474 TTL flip-flop, a few transistors, and three relays. The KEY UP output of U1 is a nominal +2 volts. Q1 is conducting, which effectively limits the voltage across the 1-μF timing capacitor to near zero volts. With the +1 volt on pin 2, and 3 volts on pin 3, the output of U3 is also low. This places K2 and K3 in the de-energized (receive) state.

When the key is pressed, Q1 is cut off, which lets

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fig. 1. Schematic diagram of the key-toggle system. K1 is a Radio Shack relay (275-004); K2 and K3 are relays 275-206. Ensure that the contact ratings of the relays are not exceeded by the controlled equipment.
the op-amp circuit start ramping positive because of the +2 volts and the timing components connected to pin 2 of U1. When the level at pin 2 of U2 just exceeds the voltage at its negative input, the comparator snaps high (with the help of some positive feedback). This toggles the U3, making Q high (pin 9) and Q low. With pin 9 of U3 high, a turn-on voltage is fed to Q3 and Q4 through timing control networks. The base network for Q3 comes on first and causes it to turn on very fast and turn off slowly. Q3 operates K2 for antenna control and receiver muting. The RC network connection through K3 to the base of Q4 causes this transistor to turn on after Q3. After K3 is activated, the RC network is removed, allowing Q4 to turn off before Q3. Fig. 2 shows the timing sequence for K2 and K3. K2 is the basic keying relay and must be able to follow keying speeds. For some keying circuits, K1 may be replaced with a 1-kilohm resistor for direct keying.

By using the Q output of U3 and Q2 as a switch, the system provides a short time constant for switching from receiver to transmit and the longer, variable-time constant when switching from transmit to receive.

why key-toggle

Back in May, 1950, Hiele showed a system which, with variations, has been much used. In Hiele's system, the switch-over from receive to transmit is done quickly, just as in key-toggle, but the transmit mode switches to receive after your key is up a selectable time, sort of a "dead man's switch." This system is used in many modern transceivers through an application of the VOX system to CW operation. The relay timing system is simpler with this method because your key must be up in order to switch, but it has its own drawbacks in operation. If you adjust the time delay to allow for reasonable pauses when sending, the time seems like forever when switching over. And, if you adjust the timing for a quick turnover, your receiver blasts you when you pause. With key-toggle, you are in command. In fairness, however, it is recognized that key-toggle is not applicable to any automatic type of keyer.

power supply and construction

I used an inexpensive wall transformer supply system, although the 12-volt units usually supply 14 volts when lightly loaded, this poses no problem to the relays or op-amp. A zener diode is used to obtain +5 volts for the comparator and flip-flop. Maximum current is approximately 150 mA. Since this device is always used where strong rf fields are generated, a metal enclosure is probably mandatory.

connections

Outputs from Key-Toggle are contact closures intended to operate existing antenna and transmitter relays. Capacitors have been chosen to prevent switching transients from burning contacts and from false-triggering U3. For low voltage operation, as is the case in solid state equipment, the closures may be applied directly to the equipment.

operating with key-toggle

Assuming you are in receive status, you switch to transmit at the initial touch of your key (the switching process will shave a small piece off of your first dot or dash). Your transmitter will remain active until you hold a dash slightly longer than normal, the exact time determined by the TIME DELAY setting, which may be set between approximately 0.2 and 2.0 seconds. Adjust the time delay pot such that the time required to toggle your system is slightly longer than the time length of the dashes at the rate you intend to send.

One note, however: If you simply press your key and hold it when switching from receive to transmit, your transmitter will remain ON as long as the key is held down. After switching from receive to transmit, your key must be released at least once before a dash can toggle the relays.

This system takes a little time to get used to, but once you get the hang of it you can actually toggle for a quick listen between words, even between letters when you are not sending too fast. And switching will only occur when you choose.

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All serious communications receivers, transceivers, and most transmitters use the heterodyne principle for frequency generation. For receivers, various frequencies are converted to a single frequency called the intermediate frequency (i-f), while primary and secondary coils are resonated by a capacitor in a shielded can.

Fig. 1A shows two i-f transformers. Both of these are in the now-standard 1.9-cm (0.75-inch) shielded can. One of the transformers uses a threaded rod attached to the inductor core for tuning, while the other uses just an access hole to admit a tuning tool. In some cases, the inductor core will have a screwdriver slot to permit adjustment, while in others a hexagonal hole is cut through the center of each core. In the latter case, a special insulated hex alignment tool is used. The hex type offers the advantage of permitting adjustment of both primary and secondary tank circuits from the same side of the chassis.

Fig. 1B shows the pins at the bottom of the transformer. For an i-f transformer, at least four pins are required and some may have five or six pins. Some older models used wire leads instead of pins, and, of course, transformers intended for printed-circuit mounting will have solder tails instead of pins.

Fig. 2 shows several of the literally dozens of i-f transformers used in Amateur Radio gear. The transformer in Fig. 2A uses two tuned coils. In a few early receivers, the tuning of the tanks was accomplished by the adjusting capacitors, the adjustment screws being accessible through holes cut into the shielded can. Most modern transformers, however, use a slug-tuned coil to tune the tank. The version shown in Fig. 2A is commonly found in vacuum-tube equipment. Figs. 2B and 2C show transformers designed for use in solid-state equipment. The version in Fig.

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2B uses taps on the primary and secondary windings to match the low impedances normally found in transistor circuits. In fig. 2C the secondary is an untuned link. The transformer in fig. 2D is very similar to the one in Fig. 2A, except that a built-in bypass capacitor is provided to decouple the cold side of the secondary. These i-f transformer circuits are not unusual; they represent those most commonly found in receivers.

Fig. 2E shows a pinout configuration used by many manufacturers; note the odd sequence for numbering the pins. A color dot, usually green, identifies pin number one, with pins two through four occupying the four corners. These pins are usually connected to the tank circuit in the manner shown in fig. 2A. Pins 5 and 6 are reserved for such things at taps (fig. 2B) or bypass capacitors (fig. 2D).

Older units, using wire leads instead of pins often follow the specific color code:

- blue — plate
- red — B+
- green — grid or detector diode
- black — grid or diode return

**transformer failures**

Fortunately, i-f transformers are relatively simple devices, with failures limited to a few problems, such as open windings, shorted or open capacitors, windings shorted together, and the inability to resonate.

Open windings rarely occur anywhere except right at the transformer pins. Fig. 3 shows a close-up view of the wire from one winding that has broken at the pin. Once the signal tracing gets you to the driver to pry open the metal shield. Most of the transformers in use have a coil form mounted on a plastic base, with the entire assembly slid into the shielded can. Tabs on the can are then bent over to form retainers which keep the coil form and base in place. It is a simple matter to pry these tabs loose and then gently pull the base from the metal shell.

Use small long-nose pliers when working on the transformer. In fact, it is best if you use tweezers instead of pliers. Avoid pulling on the wire; it will break, and right at the coil form! It is then almost impossible to repair the coil without rewinding it.

The wire should be resoldered to the pin using a small, pencil-type soldering iron, not a gun. There will usually be enough pretinned wire left to allow re-soldering to the pin. But, if not, do not try stripping the wire or you will break it. Insulation may be removed by gently scraping the wire with a razor blade or sharp knife. Keep in mind that even this is at the risk of breaking the wire. The best method is to melt the insulation with the tip of the pencil iron.

Ordinarily, there will be enough slack to make the new connection without stretching the wire. If not, bend the pin toward the wire — do not try to stretch the wire. If this is not possible, then solder a short correct stage, this type of problem can be diagnosed with an ohmmeter.

The open winding is also the usual cause of the intermittent i-f. Interestingly enough, I have seen many cases over the years where the wire had never been soldered. The transformer worked nicely for many years, until either corrosion built up or a mechanical jarring knocked the wire loose.

Repairing the open-winding problem is easy in many cases — if you are gentle. Use a small screwdriver to pry open the metal shield. Most of the transformers in use have a coil form mounted on a plastic base, with the entire assembly slid into the shielded can. Tabs on the can are then bent over to form retainers which keep the coil form and base in place. It is a simple matter to pry these tabs loose and then gently pull the base from the metal shell.

Use small long-nose pliers when working on the transformer. In fact, it is best if you use tweezers instead of pliers. Avoid pulling on the wire; it will break, and right at the coil form! It is then almost impossible to repair the coil without rewinding it.

The wire should be resoldered to the pin using a small, pencil-type soldering iron, not a gun. There will usually be enough pretinned wire left to allow re-
A study by the service department of a major automobile radio manufacturer revealed that, in humid regions of the country, there are roughly two to three times as many trimmer capacitor and i-f transformer problems as in dry areas. The best remedy for this is to replace the transformer. But, if that is impossible or would take too long, try replacing the bad capacitor with a disc ceramic or silver mica capacitor.

If the original capacitor is like the one in fig. 5A, it is a simple matter to use side cutters to clip off the top spring clip contact. Then solder a capacitor of the proper value across the terminals (see fig. 6A). The proper value can be determined experimentally (check first to see if the manufacturer gives the value in the schematic) using a GDO or signal generator to locate a standard value that will allow the coil to resonate at the correct frequency.

If the bad capacitor is not easily removed or disconnected, there is still a possible cure. Disconnect the coil wire going to one end of the capacitor, and then connect it and one end of the replacement capacitor to an unused pin. Or add a pin in one of the blank pin slots (fig. 6B) by forcing a piece of hook-up wire through the slot.

The problem of primary to secondary short circuits comes about because the wires from the upper coil pass by, and may touch, the winding of the lower coil. The problem almost always results in B+ leaking through to the grid of the next stage or the plate of the detector diode.

The cure here may be as simple as moving the wires apart and then patching up the damaged spot using Q-dope, high voltage corona dope, or just glue. Replacement of the transformer, however, may be required.

You’ll sometimes find a transformer that is not causing static, passes all dc checks, but signals will...
not go through or are severely reduced in strength. Barring such rare problems as shorted windings and open capacitors, it is often the case that one of the tanks is off resonance due to a cracked or broken ferrite tuning slug. To replace the slug, break it into smaller pieces (it rarely will come out by its threads!) and then remove them by shaking. A replacement slug can then be installed. Find an i-f transformer of the same frequency range and with the same type of slug. Salvage the slug and install it in the bad transformer. Some electronic parts suppliers stock small assortments of tuning slugs for just this purpose. Note that 9-MHz ssb transceiver slugs can often be replaced with a slug from a 10.7-MHz fm broadcast receiver.

Regardless of the problem, except perhaps the open coil repair, the best solution is to replace the bad i-f transformer with a new part purchased from the equipment manufacturer. However, this is not always possible, especially in older gear. Even if the company is still in business, the set may be so old that they no longer support the product. A few years after a model is discontinued, support usually vanishes.

It is possible to buy new i-f transformers. Companies such as J.W. Miller and others still manufacture both direct replacements for many types, and so-called "universal" i-f transformers. Their catalogues should reveal at least one or two currently made models which are good candidates for replacement of the bad unit in your rig. It's unlikely that a cross-reference listing for a piece of ham gear is available, but a careful reading of the specifications and inspection of the circuit diagram will lead to a replacement.

Older 2.5-cm (1-inch) or larger i-f transformers are especially difficult to replace. However, if you are willing to make a few mechanical modifications, the standard 1.9-cm (0.75-inch) transformers can be pressed into service using the adapter plates and mounting clips (fig. 7) provided with replacement transformer.
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RF input power is 200 W PEP on SSB, 160 W DC on CW, and 100 W DC on FSK. Receiver sensitivity is better than 0.25 μV for 10 dB S/N.

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TRIO-KENWOOD COMMUNICATIONS, INC.
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updating the Heath HW-2036 to the HW-2036A

Modifications to the Heath HW-2036 to provide complete coverage of the new two-meter repeater subband

Now that the two-meter band has repeater allocations in the lower half, people with HW-2036 transceivers are feeling left out in the cold. No longer! With a small parts kit available from Heath, a few additional parts obtained (through Heath or purchased on your own), and these simple instructions, you can update your HW-2036 to cover the entire band. In fact, the transceiver will operate over a six-MHz range, from 143 MHz to beyond 149 MHz, although with reduced output at the extreme edges.

updating early HW-2036s:

Before actually beginning conversion of the HW-2036, check to see that your HW-2036 has all the updates that Heath has added since this model was introduced. There were addendum sheets included with the manual, in addition to several changes in the manual itself to reflect parts values different from those printed on the circuit boards. On the receiver board, R206 and R209 should be 10 kilohms, and C214, C215, and C216 should be 4.7-pF NPO disc ceramics. Also, Q204 was changed from a MPF-105 to an EL-131. If you have to replace these six parts, you will need to realign the receiver i-f. Wait, however, until the rest of the changes have been made.

On the synthesizer board, R419 should be 10-kilohms, C442 a 20-pF NPO disc ceramic, C403 a 100-pF NPO disc ceramic, and R445 220 ohms. These parts changes reduce drive to the synthesizer loop mixer. And finally, on the transmitter board, R141 should be 180 kilohms, ½ watt, which increases the zener bias in the 11-volt regulator. Only very early HW-2036s might need any of these changes. Your HW-2036 is now up to date. Table 1 lists all the parts needed to convert the HW-2036.

conversion to the HW-2036A

There are changes to be made on each board, and we found it easier to do one board at a time and then do the alignment board-by-board when everything was bolted back together. It is necessary to have some method of removing solder from the double-sided boards before attempting to remove the parts.

Power Amplifier Board. Remove the transceiver top and bottom covers and begin with the power amp board by removing the heatsinks from the back panel. Unsolder and disconnect the ground wires connected to the corners of the transmitter and receiver boards. Carefully pull the molex connector apart, remove the two nuts that hold the power amplifier assembly to the back of the chassis, and swing the board down on its leads. A 12-pF dipped-mica capacitor (C235 in the HW-2036A manual) is added to the output connector on the inside of the board.

Prepare this capacitor by cutting the leads 22 mm (7/8 inch) long and slipping 19 mm (3/4 inch) of spaghetti or insulation over each lead. Bend the exposed 2-mm (1/8-inch) lead at right angles away from the body of the capacitor. Lay the capacitor on the foil of the board and solder one lead to the antenna jack center pin and the other to the foil where the ground wire pokes through the board. This forms a filter across the output. The lead length is the inductance part of the filter. Reassemble the board to the rear apron, but leave the ground wires to the transmitter and receiver boards loose for now. Replace the molex connector.

VCO Board. Heath now supplies a parts kit for modifying the VCO; it consists of two mica capaci-

By Mike Miller, WB6TMH and Ed Fitzgerald, WA6ODR. Mr. Miller’s address is 173 Leveroni Road, Sonoma, California 95476. Mr. Fitzgerald can be reached at Post Office Box 75, Cotati, California 94928.
form a filter on the output of the transmitter.

Unsolder and remove L501, C513, C509, and VD502. Replace each component with the parts supplied in the Heath package. Reglue L501 to the board, C503 to the side of L501, and C513 and C509 to each other. Now, replace the board on its bolts and turn on the power. Set the frequency switches to 146.000 MHz and run the slug of the new coil up and down, checking that the synthesizer lock light on the front panel goes out at some point, indicating the VCO is operational. Install the cover, gluing the coil to the top and soldering the lugs on the side. These changes increase the tuning range of the VCO by increasing the oscillator L/C ratio and making the varicap diode a larger part of the total capacitance. A full alignment of the VCO will be completed later.

Synthesizer Board. First, remove the black wire attached to point C on the power amp board to allow room for the synthesizer board to swing up. Remove the thirteen wires that attach the thumbwheel switches to the front end of the board and the four nuts and washers. Now, swing the board up on its remaining leads and remove R427 and R428, changing the values so that R427 becomes 5.6 kilohms and R428 10 kilohms. Page 61 of the manual can be used to locate these two resistors.

Replace the board on its bolts, attach the black wire to point C on the power amp board, and refer to pictorials 4-10 and 4-11 for color code and wire placement when attaching the thirteen wires to the frequency switches. Again turn on the power and check to see that the synthesizer lock light goes out.

Receiver Board. Remove the four nuts and washers, remove the coax from points A and B, and tilt the board up on its remaining leads. Referring to pages 45-52 in the manual, change the value of R233 to 15 kilohms. This increases the drive to the receiver doubler.

Next, remove C201 and the shield alongside of L201 and L202. Remove 3 mm (1/8 inch) of the paper insulation from the vertical edge nearest L213. Cut a piece of 0.2-mm (0.01-inch) brass shim stock or copper flashing to 22 × 16 mm (7/8 × 5/8 inch), bending 3 mm (1/8 inch) of the long side at right angles to form a tab. Clean the shim stock and shield with steel wool and solder the tab to the existing shield where the insulation was removed. Alligator clips will hold the pieces together during the soldering.

Remove L202 and install the modified shield on its original pins. Solder the shield pin nearest L204 to hold it in place and run a bead of solder along the shield extension, joining it to the board foil behind C203 and L202. Also, run a bead of solder along the back of the shield next to C201. Replace L202, noting the polarity, and change C201 to a 5-pF NPO disc ceramic; solder both components in place, as well as the second shield pin.

Bolt the board down again and resolder the ground lead from the power amplifier board to the corner of

Location of C235 which, in conjunction with its lead length, forms a filter on the output of the transmitter.

Table 1. Parts list of components needed to update the HW-2036 to an HW-2036A. Unless otherwise noted, all resistors are 1/4 watt, 5 per cent tolerance.

<table>
<thead>
<tr>
<th>Designation</th>
<th>Value/Description</th>
<th>Heath Number</th>
<th>Heath Price</th>
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<tbody>
<tr>
<td>C214, C215</td>
<td>4.7-pF NPO disc ceramic</td>
<td>21-168</td>
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<tr>
<td>C216</td>
<td></td>
<td>21-51</td>
<td>$0.25</td>
</tr>
<tr>
<td>C442</td>
<td>20-pF NPO disc ceramic</td>
<td>21-51</td>
<td>$0.25</td>
</tr>
<tr>
<td>Q204</td>
<td>EL 131 (replaces MFP-105)</td>
<td>417-241</td>
<td>$2.10</td>
</tr>
<tr>
<td>R141</td>
<td>180 ohms, ½ watt</td>
<td>1-112</td>
<td>$0.22</td>
</tr>
<tr>
<td>R206, R209</td>
<td>10 kilohms 10 per cent</td>
<td>1-9 12</td>
<td>$0.24</td>
</tr>
<tr>
<td>R419</td>
<td>10 kilohms</td>
<td>6-103-12</td>
<td>$0.25</td>
</tr>
<tr>
<td>R446</td>
<td>2.2 kilohms</td>
<td>1-4 12</td>
<td>$0.24</td>
</tr>
<tr>
<td>C403*</td>
<td>100-pF disc ceramic</td>
<td>21-75</td>
<td>$0.25</td>
</tr>
<tr>
<td>R208</td>
<td>10 kilohms</td>
<td>6-103-12</td>
<td>$0.25</td>
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Conversion parts

<table>
<thead>
<tr>
<th>Designation</th>
<th>Value/Description</th>
<th>Heath Number</th>
<th>Heath Price</th>
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</thead>
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<tr>
<td>C503</td>
<td>0.1-pF monolithic ceramic</td>
<td>21-192</td>
<td>$1.35</td>
</tr>
<tr>
<td>C509*</td>
<td>22-pF 5 per cent dipped mica</td>
<td>20-99</td>
<td>$0.40</td>
</tr>
<tr>
<td>C513*</td>
<td>125-pF 5 per cent dipped mica</td>
<td>20-117</td>
<td>$0.50</td>
</tr>
<tr>
<td>C201*</td>
<td>5-pF NPO dipped ceramic</td>
<td>21-78</td>
<td>$0.25</td>
</tr>
<tr>
<td>C235*</td>
<td>12-pF 5 per cent dipped mica</td>
<td>21-130</td>
<td>$0.45</td>
</tr>
<tr>
<td>L501*</td>
<td>0.25 μH</td>
<td>40-1865</td>
<td>$0.95</td>
</tr>
<tr>
<td>VD502*</td>
<td>Motorola MV2110</td>
<td>56-640</td>
<td>$1.65</td>
</tr>
<tr>
<td>R103*</td>
<td>100 ohms</td>
<td>6-101-12</td>
<td>$0.25</td>
</tr>
<tr>
<td>R233*</td>
<td>15 kilohms</td>
<td>6-153-12</td>
<td>$0.25</td>
</tr>
<tr>
<td>R427*</td>
<td>5.6 kilohms</td>
<td>6-662-12</td>
<td>$0.25</td>
</tr>
<tr>
<td>R428*</td>
<td>10 kilohms</td>
<td>6-103-12</td>
<td>$0.25</td>
</tr>
</tbody>
</table>

*Note: These parts can be obtained as a kit with instructions (Heath part number 830-29) for $3.95

March 1979
the receiver board. Connect the power, turn on the transceiver, and tune in a local repeater to see that everything works properly. Full alignment will be done last.

Transmitter Board. Remove the phono plugs from J101 and J102, and remove the five nuts and washers holding down the board. Disconnect the molex connector to the power amplifier board and also the four large power connections labeled G, P, R, and V. Pull off the three-pin connector on regulator IC1 and unsolder the lead of the 0.001-µF capacitor from ground lug DA (refer to page 106). Carefully tilt up the board on its remaining leads.

Referring to page 74, remove R103 and replace it with a 100-ohm resistor. Remove the connectors on pins B, L, and N and solder the base of these pins to the board foil. These pins are the shield connections for interconnecting coax leads and should be rigidly grounded to both sides of the circuit-board ground plane!

Pull the red wire from pin S. Clip off and discard the 0.001-µF capacitor from the end and remove the two ferrite beads. Strip 3 mm (1/8 inch) of insulation from the end of the red wire and add approximately 7.5 mm (3 inches) of wire of a similar size, covering the soldered splice with heat shrink tubing or spaghetti. Now strip 16 mm (5/8 inch) of insulation from the extended wire and slip on the two ferrite beads just removed. There should be about 3 mm (1/8 inch) of wire sticking out. Solder this end to pin OUT, alongside the yellow wire from regulator IC1. Next, take a 2.5-cm (1-inch) piece of solid wire, bend a 3-mm (1/8-inch) tab on one end, and solder this tab to the circuit board foil next to pin N.

Install the board on its bolts again. Be sure to position the long coaxial cable along the top of the front of the board, as shown on page 106 of the manual. Replace all board connectors, phone plugs, and the connector on IC1. Don’t forget the shield connector to pin B, hidden under the wiring harness. Now, solder the other end of the 2.5-cm (1-inch) wire from the board foil near pin N to solder lug DA on the rear apron. Make this lead as short as possible and trim off the excess. Reconnect the ground wires from the receiver and transmitter boards to the power amplifier board.

Install two ground lugs at stud X on the power amplifier board. Solder one lug directly to the chassis. Remove the coax shield from point B on the power amplifier board and solder it to the second lug on stud X.

alignment

VCO Board. Set C511 on the VCO board to mid capacity. Attach the dc probe of a VTVM to TP401 on the synthesizer board; set the frequency switches to 146.000 MHz; and set the mode switch on the front panel to SIMPLEX. Turn the slug of coil L501 up until it shows at the top of the coil. Key the transmitter with the dummy load attached and rotate the slug down until the VTVM reads 2.2 volts. The SYNTHESIZER LOCK light should be out at this point.

With the transceiver not keyed, adjust C501 until the VTVM again reads 2.2 volts. This completes the VCO adjustments.

Transmitter Board. Since the transmitter strip is already aligned to a center frequency of 147.000 MHz, an abbreviated procedure is used to center the transmitter at 146.000 MHz, providing full output over the entire frequency range. First, install the dummy load. Set the thumbwheel switches to 146.000 MHz and connect the rf probe to TP101. The kit-supplied probe is fine and can be used with the dc scale of a standard VTVM for better sensitivity. Adjust L101 and L102 for maximum output with the transmitter keyed. Several adjustments may be necessary.

Move the rf probe to TP 102, key the transmitter,
and adjust L103 and L104 for maximum. Remove the probe, and, with the transmitter keyed, adjust L105, L106, and L107 for maximum output as indicated on the front panel's relative power meter. The peak on L106 will be very broad.

Now, set the frequency switches to 144.500 MHz and adjust L101 for maximum output at this lower band edge. Then, with the frequency set to 147.500 MHz, adjust L102 for maximum. Repeat these two steps, which are designed to stagger the tuning over the full band. Finally, set the frequency to 144.000 MHz, key the transmitter, and check the relative power meter to be sure the transmitter has full output at the extreme band edge. Repeat this step at 147.990 MHz. There may be a slight difference at each end, but not much.

**Power Amplifier Board.** Set the frequency switches to 145.000 MHz, and, with the transmitter keyed into the dummy load, adjust the power amplifier trimmers in this order: A, B, C, D, E, D, C, A, B. Do not adjust trimmer capacitor A any further clockwise than is necessary for maximum output. This single adjustment series should provide nearly equal output across the entire band. Remember to replace the heatsinks!

**Receiver Board.** Set the frequency to 146.000 MHz, connect the rf probe to gate 2 of Q202 on the receiver board, and adjust L402 and L403 on the synthesizer board and L212 and L213 on the receiver board for maximum level. Perform these adjustments in the order listed and repeat at least once. This provides maximum local-oscillator injection to the first mixer.

Next, remove the antenna coax from points A and B on the receiver board and connect the kit-supplied jumper from points A and B to points TP106 and TP107 on the transmitter board. Now, adjust L201, L202, L203, and L204 for maximum S-meter reading with the frequency set to 146.000 MHz. It may be necessary to adjust the frequency up or down slightly to pick up the harmonics of the reference oscillator being used as the alignment signal source. Also, since there is a high signal level, C147 on the transmitter board may be adjusted to maintain the meter reading at half scale or less. If this does not provide enough signal reduction, slightly detune i-f coil L207 on the receiver board. Once L201 through L204 are adjusted properly, L207 can be repeaked by removing the jumper from point A and holding it a small distance from the pin. This will provide enough signal to peak L207 without overloading the i-f strip.

At this point, if you replaced the parts in the i-f strip, refer now to the manual instructions for aligning the i-f stages. Few will need this however, and, if not disturbed, the rest of the i-f will not need attention. This completes the conversion.

**remarks and notes**

We found that the birdies on 146.000 MHz and 146.520 MHz were slightly decreased in signal strength but not eliminated. The reference oscillator and offset oscillator should not need attention if care was taken not to disturb their trimmers during the modification.

**Reducing power consumption.** Several ICs can successfully be replaced with low-power Schottky versions (LS series). IC104, IC105, and IC106 (7490s) may be replaced with 74LS90s. IC107 (7492) may be changed to a 74LS92. IC405 (7400) may be changed to a 74LS00, and IC404 (7473) may be changed to 74LS73. IC103 (7400) may not be replaced! These changes will reduce current drain about 20 per cent. It may be possible to change these devices to the CMOS series for further power reduction, but we haven't yet tried that substitution.

![Close-up of the receiver shield with the addition soldered to the right edge.](image)
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10 meters. Complete with trunk lid mount assembly.
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Complete...
BM-5800T - 144-174 MHz 1/4 wave gain antenna 200 watt
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Note: Deduct $4.00 from "M" price
for "P" assemblies.

More Details? CHECK—OFF Page 118
OX OSCILLATOR
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Specify when ordering.

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0.2% Calibration Tolerance

EXPERIMENTER CRYSTALS
(HC 6/U Holder)

<table>
<thead>
<tr>
<th>Cat. No.</th>
<th>Specifications</th>
<th>Price</th>
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<tr>
<td>031080</td>
<td>3 to 20 MHz — for use in OX OSC Lo</td>
<td>$6.25 ea.</td>
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<tr>
<td>031081</td>
<td>20 to 60 MHz — For use in OX OSC Hi</td>
<td>$6.25 ea.</td>
</tr>
<tr>
<td>031300</td>
<td>3 to 20 MHz — For use in OF-1L OSC</td>
<td>$5.22 ea.</td>
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the dasher

Update your Vibroplex bug by using the Dasher — a simple circuit that provides a continuous series of dashes.

Back about 1946, during my days as a commercial CW operator, I used to spend an eight-hour shift sending traffic in strings of 40 or more messages. Many of them were weather messages which had groups of five digits and many, many groups per message. I was getting a glass arm doing this at 40 words per minute or better every working day.

I saw a circuit in an Amateur magazine which used two tubes in a multi-vibrator circuit that could be mounted on top of a regular Vibroplex bug and would automatically make the dashes while you still get the dots from the mechanical side of your bug. I built and used one for years, saving my arm a thousand times over. It had the advantage that you could make the dashes much longer than the traditional 3 to 1 ratio. Therefore, you could retain your original sending style, yet send perfect code without fatigue.

Since then I haven’t been too active — until recently when I got back into both MARS and CW work. That’s when I remembered that silly thing I once built into a cake pan and mounted on top of my old Vibroplex.

Today, however, with all solid-state devices and small parts, one can be built in a small box, using a 9-volt battery for power. I used the 555 IC as a timer and a small 6-volt, dc-sensitive relay.

Circuit description

Fig. 1 shows the circuit I used for the Dasher. In some circuits, pin 7 on the 555 was keyed. I tried this configuration but it caused a slight “hangover” on the last dash of a train, making it very difficult to handle. I decided to try keying the 9-volt supply to the entire circuit, but this caused the first dash of a series to be stretched. Then, I tried leaving the 9 Vdc connected to all parts of the circuit and keying the lead to pin 8. This got rid of the problems, resulting in perfect dashes.

The two pots permit adjustment of both speed and weight. The only change needed on a Vibroplex bug is to disconnect the lead under the base from the dash contact post. Then, use a two-wire, shielded mike cable to provide three connections from the bug to the input of the Dasher. The shield forms one lead from the frame post of the bug, and the other two leads connect to the dot and dash contacts. Inside the Dasher, the lead from the dot contact is connected to the output lead, with the dash contact used to control the 555.

Two things I might add for information. I used one of the rechargeable, 9-volt, ni-cad batteries and brought leads from the battery to screw terminals on the side of the box. This way I can connect a charger without opening the box. Also, a note of caution. Do not ground any oil the 9-volt points to a ground that is common to the relay output or the bug common lead. When I tried the Dasher with my Collins KWM2A barefoot, no problems were encountered. However, when I ran the linear with a full kilowatt, rf was induced into the leads, causing erratic dash length. Putting ferrite beads on the input and output leads cured the problem. So have fun and don’t end up with a glass arm; use the Dasher.

By Joseph H. Fenn, KH6JF, 3612 Puuku Makai Drive, Honolulu, Hawaii 96818

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formation of passive lumped constant 90-degree phase-difference networks

Construction details for designing equal amplitude, 90-degree, phase-difference networks

A variety of networks, both active and passive, have been used over the years for splitting a signal into two equal amplitude signals while maintaining a quadrature phase relationship between the two components. Roger Harrison's article presented a number of such circuits and rekindled my interest in these devices. I first became interested in these circuits while designing image outphasing systems for sweeping microwave receivers about 20 years ago. At UHF and microwave frequencies, 90-degree phasing networks are readily achieved with coaxial or stripline hybrids which can be designed for bandwidths of several octaves. These distributed element hybrids become large and difficult to build for the lower frequencies. Sometimes, it is also desirable to have circuits capable of greater bandwidth.

I have prepared a table (see table 1) of prototype values from which passive, lumped-constant, first-order lattice networks (fig. 1) may be designed. This table is an adaptation of early work by Darlington and later work by Bedrosian. Darlington's work is basic and requires background in network theory plus a comprehensive table of elliptic functions to arrive at a practical design. Bedrosian's paper included a number of tables of prototype values which can be used to calculate inductances, but you are stuck with the arbitrary bandwidths he chose unless you have access to a table of elliptic functions. Even then, there are several calculations to be done before arriving at a design. I have chosen a number of bandwidths which I believe would be most useful for Amateur Radio users, 2 to 1 through 20 to 1.

network description

The phase-splitting networks consist of two separate lattice networks, a P-network and a Q-network. For a given bandwidth, the amount of phase ripple (deviation from 90-degree phase difference) depends upon the number of lattice sections, the more networks the smaller the deviations. \( N \) is the total number of lattice sections in both branches.

Although the first-order networks are simple, they may not be the most practical. This sometimes becomes apparent when you try to produce a pair of networks operating over a relatively broad range.

You may discover that the networks, which are operating over the low-frequency end of the band as "all pass" networks, are also behaving at the higher frequencies like lowpass filters with a cut-off frequency within the design bandwidth. The reason for this is the self-capacitance of the inductors, which makes it difficult to place the self-resonance of the coils in the low-frequency lattices outside the design band.

Orchard presented a transformation whereby two first-order networks may be combined in a single second-order network as shown in fig. 2.

figure 1. Diagram of a first-order lattice network.

description and use of the table

Once you have determined the bandwidth required (ratio of upper to lower frequency) and the phase ripple that can be tolerated, you are ready to select the appropriate prototype design from the table. Assume a 3 to 30 MHz 90-degree network is needed, requiring

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You are now ready to complete the design using the $P$ and $Q$ values from Table 1. For the $P$ network assume $Z_o = 50$.

$$L_1 = \frac{P_1 \cdot Z_o}{2\pi f_0} = \frac{0.151 \times 50}{2\pi \times 9.487 \times 10^6} = 0.1267 \, \mu H$$

$$L_2 = \frac{1.675 \times 50}{59.6 \times 10^6} = 1.405 \, \mu H$$

$$C_1 = \frac{P_1}{Z_o \cdot 2\pi f_0} = \frac{0.151}{50 \times 59.6 \times 10^6} = 50.7 \, pF$$

$$C_2 = \frac{1.675}{50 \times 59.6 \times 10^6} = 562 \, pF$$

For the $Q$ network:

$$L_1 = \frac{Q_1 \cdot Z_o}{2\pi f_0} = \frac{0.597 \times 50}{2\pi \times 9.487 \times 10^6} = 0.5 \, \mu H$$

$$L_2 = \frac{6.61 \times 50}{59.6 \times 10^6} = 5.54 \, \mu H$$

$$C_1 = \frac{Q_1}{Z_o \cdot 2\pi f_0} = \frac{0.597}{50 \times 59.6 \times 10^6} = 200 \, pF$$

$$C_2 = \frac{6.61}{50 \times 59.6 \times 10^6} = 2218 \, pF$$

Now the complete circuit can be sketched (see fig. 3). An example of an $N = 5$ network for the same bandwidth is shown in fig. 4.

The first-order lattice is a balanced circuit whose unbalanced equivalent requires the use of a transformer. Several types of balanced-to-unbalanced transformers, which are suitable for adapting the lattice to an unbalanced circuit, are described in reference handbooks. You have the option of choosing the impedance level of the lattices to effect the desired impedance match at the common junction of the networks.

**Theoretical and experimental results**

The values for Table 1 were obtained from a BASIC computer program (PHASESO) which is capable of designing a wide range of first-order networks with bandwidths up to 100 to 1. Several designs from this program were analyzed using a BASIC (GNET) network analysis program and were found to perform as predicted.

Finally, to prove the usefulness of the theoretical design, a four-section ($N = 4$) network consisting of first-order lattices was built for the 20 to 150 MHz range. The phase deviations from 90 degrees were about 5 degrees, which is greater than the theoretical tolerance of 0.7 degree, but adequate for most requirements. This experimental data shows that the first-order lattices can be used for an 8 to 1 band-
width. For greater bandwidths you may have to resort to the second-order lattice configuration.

acknowledgments

I am indebted to R. E. Booth, WB6SXY, who wrote the GNET network analysis program, and whose assistance in writing the PHASE90 program is gratefully acknowledged.

references


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fig. 3. Component values for the 3 to 30 MHz, four-section network.

fig. 4. A five-section equivalent of the network shown in fig. 3.
The SE.01 is a kit that contains all the parts needed to build a programmable sound effects generator. Designed around the new Texas Instruments 76447 Sound Chip, the board provides banks of Mini DIP switches and pots to program the various combinations of the SLF Oscillator, VCO, Noise, One Shot and Envelope Controls. Another IC is used to implement an Adjustable Pulse Generator, Level Comparator and Multiplex Oscillator for even more versatility. The 3½" by 5" plated PC Board features a prototype area to allow for user added circuitry. Easily programmed to duplicate Explosions, Phaser Guns, Steam Trains, or an almost infinite number of other sounds, the unit has a multitude of applications. The $18.95 price includes Assembly Manual, Programming Charts, and 76477 Chip specifications (speaker not included). Available from stock.

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cabinet construction techniques

A new construction technique which allows the builder to build a cabinet to fit his equipment — rather than fitting his equipment into the cabinet.

All builders of electronic devices eventually come up against a continuing problem — what type of cabinet to put their creation in. The magnitude of the problem increases in direct proportion to the dimensions desired. Commercially manufactured answers to the problem are few. If a cabinet is inexpensive, it is usually too small or flimsy; if it is rugged enough it will usually be too expensive. Trying to find a cabinet with the desired size and configuration often seems next to impossible. It has been my experience that few local dealers ever carry a large percentage of a cabinet manufacturer’s total line. Therefore, mail order must be resorted to, which takes time.

Many times, the builder must choose “the best of the worst” from stock on hand and then tailor construction of the device to fit the cabinet. This practice usually results in unnecessary work and design trade-offs. Another problem that occasionally occurs is the need to increase or decrease the depth of a cabinet, something not easily accomplished. All of this adds up to what I call the “cabinet frustration syndrome.”

What follows is an idea that evolved from attempts to solve these problems. The result is a rugged, inexpensive cabinet which is easy to construct in any needed size. Fabrication materials are universal in that they can be used for most any size cabinet. Most of the pieces from scrapped cabinets can be used again. With this method, the depth of the cabinet can be easily extended or reduced.

fabrication

The basic cabinet is made of only two items, the end plates and the end-plate support brackets. The end plates can be the front and rear panels of your unit or they can be subpanels, using a method to be described later. The two end plates are simply cut from aluminum sheet stock to the desired front-panel dimensions. A mounting hole for each bracket is drilled in each corner. The end-plate support brackets are the key to this whole method of construction. Bolted between the end plates, they provide cabinet rigidity, while at the same time allowing construction flexibility.

The brackets are made of aluminum U-channel, cut to the desired depth. Aluminum U-channel was chosen because it is light, relatively inexpensive, and readily available at most hardware stores. My first attempt, for which I used 12.5-mm (1/2 inch) U-channel, was acceptable but resulted in a more rugged (and heavier) cabinet than was necessary for that application. Cabinets built of 9.5-mm (3/8-inch) channel proved to be best for most of my uses.

Once the U-channel size is decided upon, the next step is to locate or fabricate something to lay in the channel that will accept the end-plate mounting screws. My junk box yielded some threaded metal standoffs about 5 cm (2 inches) in length. These fit snugly in the channel. The standoffs had a 10-32

By Bob Thompson, W7KDM, 12213 South Oneida, Phoenix, Arizona 85044
As seen in this photograph, the standoff was drilled and tapped for two additional holes, providing a means of securing the standoff in the channel stock.

(M5) threaded hole through their length. Two other mounting holes were drilled perpendicular to the length and tapped for a 10-32 (M5) thread. These holes are for bolting the standoffs to the U-channel.

If you don’t have any standoffs, square aluminum stock is usually sold along with the U-channel. This can be cut to length, drilled, and tapped as necessary. Of course, the square stock can be cut to a length equal to the cabinet depth and used as the entire mounting bracket. In that case, only one hole must be drilled and tapped in each end. However, a solid bracket will have a few disadvantages when compared with the U-channel approach; it will be heavier, and it will offer less flexibility in the mounting of parts to the brackets.

Once each bracket has a standoff mounted at each end, the four brackets are bolted to the end plates, and the basic cabinet is finished! Component mounting plates can be above or below the brackets. If a center mounting plate is needed, this can easily be accomplished with L-brackets. Incidentally, a good source of L-brackets for any project is L-shaped aluminum stock, usually sold in the same place as the other materials. Merely cut the stock to the width of L-bracket desired. The L-bracket will be much stronger than one made by bending a piece of flat stock.

The basic cabinet is essentially a frame that can be covered by cutting and bending lightweight sheet aluminum. The design is such that the cover is not needed to provide the basic structural integrity and strength. This is a big problem with all but the most expensive commercial cabinets. Here again, the lightweight aluminum cover material is usually sold where all the other materials are.

Rather than making a cover by bending, cutting individual pieces of stock and attaching them to their respective sides with small screws is also a possibility. Where a heavier gauge metal will be used for the cover, it is possible to entirely eliminate the brackets.

In this case, a central chassis plate is fastened to the end plates with L-brackets. Unless the cabinet is very small, the aluminum sheeting used for the end plates and chassis plate should be heavy enough to provide sufficient rigidity. Most of the aluminum that I use for chassis and end plates is 3 mm (1/8 inch) thick, although I’ve also used material 1.5 mm (1/16 inch) thick with success.

Aluminum sheeting is used because of its widespread availability and low cost. In the past, I have used some magnesium alloy sheet which provided excellent rigidity and light weight. The problem is, of course, that it is both expensive and difficult to come by.

As mentioned earlier, cabinets made with the U-channel technique can be easily lengthened or shortened. To shorten a cabinet, merely remove the standoffs from one end of each bracket, cut the brackets to the desired length, and remount the standoffs in the ends.
The process of lengthening a cabinet is just a little more involved. If materials are available, just make longer brackets. However, if you don’t have a piece of U-channel long enough to make four new brackets you can use four small pieces and add to the length of the existing bracket. This approach will require more work, but could be the most expedient and most economical under certain circumstances.

First, cut and fashion four extension brackets for the desired added depth. Remove one of the end

plates. Insert adequate lengths of threaded screw stock (I favor 10-32 [M5]), into the ends of the brackets to be lengthened. Leave about 12.5 mm (1/2 inch) protruding. Screw the extension brackets onto the exposed screw stock. Now comes the only tricky part: when the two brackets are snugged up tight against each other, they should be aligned. If they do not align, insert a piece of metal shim stock or a thin washer between the two pieces and try again until alignment is achieved. Mount the other end plate on the extended brackets. If desired, one end plate can be used as a subpanel for false front panels or speaker grilles. This is accomplished with metal standoffs.

The only item which might prove difficult to procure is the aluminum sheet used for the chassis and end plates. Sheet metal dealers will usually sell remnants at scrap value. Scrap metal dealers sometimes have acceptable pieces. Your best bet is to get out and browse around to see what is available.

It’s my hope that this article has provided you with a basic method (plus some variations) for relieving "cabinet frustration syndrome."

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More Details? CHECK OFF Page 118
So far, the basic gate has been examined under dc or static conditions. This part will discuss dynamic conditions of switching, propagation delay, and the most simple flip-flop, the latch.

All digital devices have delay from internal charge storage. Input and output may switch rapidly, but the delay between input transition and output transition is usually longer than the transition itself. Fig. 1 is an example of dynamic conditions in a simple inverter.

The term $t_p$ or $t_d$ is associated with propagation delay and will have a number of different subscript letters to distinguish different types of delay. A table in the front of the Texas Instruments TTL Data Book lists most of them. The term propagation is used, since the input-to-output path within a device may go through several stages.

The inverter shows only two delays. $t_{pLH}$ is the input transition-to-low delay to output transition-to-high state; $t_{PHL}$ is the opposite. They may occur at different times. Measurement is made at half amplitude between maximum low-level voltage and minimum high-level voltage.

Output-state transition time is also listed, almost always at a specified load of maximum capacity. Load capacity affects output transition delay strongly but has only a slight effect on propagation delay.

Multi-function devices have a number of different delays. Data sheets must be carefully studied to understand the effect of delays on the overall circuit. Lack of understanding of delays can waste a lot of debugging time (most times are too short to measure on inexpensive scopes) and also can cause a circuit to fail entirely. Delay time can’t be underestimated!

By Leonard H. Anderson, 10048 Lanark Street, Sun Valley, California 91352

the RS latch flip-flop

Fig. 2 shows the block symbol, waveforms, and an equivalent circuit constructed of NAND gates. The RS term comes from reset/set. A flip-flop is a bistable circuit (stable in two conditions) and is either set or reset by an external input. Because it’s bistable, it’s said to be latched in one state or the other.

The circuit of fig. 2 has active low set and reset inputs, indicated by inversion bubbles on the symbols. Outputs are not shown with bubbles, since the flip-flop is a bistable device.

Flip-flop terminology sometimes refers to the Q output as true, $\overline{Q}$ as false, or NOT. Output-to-input cross-connection would appear to create an oscillator. This would be true with linear devices, but it’s the key to holding a bistable condition with digital devices.

Assume initial conditions of both inputs and Q high, $\overline{Q}$ low. NAND gate G1 is held high by the low from G2. G2 is held low since both inputs are high (the NAND RULE). The initial state is stable. Either output could have been high at power turn-on, depending on which gate was the fastest.

The first S low input does nothing; G1 is already high. The first R low will force G2 high. G1 then

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fig. 1. Propagation delay of an inverter.
goes low, since all inputs are high and the latch has flipped to a reset condition (Q low). This is the other stable condition.

Note the exaggerated time delays and sequence of changes. Each input must hold at a low state long enough for both gates to change state. Flip-flop spec sheets will give this parameter of time as hold.

A following R low will do nothing: the latch is already reset. A second S low will force G1 high. G2 will go low since all inputs are high; the latch has flopped into a set condition.

What happens if both inputs are low? Both outputs will be forced high as long as that condition persists. The unknown condition occurs when both inputs return high. Either stable condition could occur and, as in power-on, it depends on which gate is the fastest. This is the indeterminate state and should be avoided.

**a switch debouncer**

All switches and relays, except for mercury wetted types, have contact bounce. Contact bounce is the opening and closing of contacts because of mechanical vibration. It lasts between 0.2 and 50 milliseconds, depending on construction, and can raise havoc with certain digital circuit inputs; a cure is found in fig. 3.

Contact bounce still exists, but the controlled circuit does not see it. The first contact closure will flip the latch. Subsequent bounce will not affect the latch as long as the switch is break-before-make. Compare this circuit with that of fig. 2 for full understanding.

---

**fig. 3. Switch debouncer.**

**fig. 2. Simple RS flip-flop latch symbol, circuit, and waveform.**

The RS latch is an interface between the switch and controlled circuit. The grounded switch arm provides the active low latch input signal. Pull-up resistors provide a high state for open contacts. The 4.7k pull-up is for TTL; CMOS values can range from 56k to 270k.

---

**clocked flip-flops**

The term clock refers to a timing signal and came from early computer work, when the clock synchronized all functions. Applied to flip-flops, the clock input is really a trigger to initiate a change of state.

Fig. 4 shows the most common clocked flip-flops, the JK and D, in symbolic form with truth tables. Two things should be noted here: These truth tables are time-dependent, and the clock input may or may not have an inversion bubble. The clock edge is the trigger. A positive-going (low-to-high) edge is shown direct as in the D flip-flop. A negative-going (high-to-low) edge is shown with the bubble.

J, K, or D inputs are control or data inputs. Each truth table shows the effect of these inputs on the Q
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output after receipt of a clock edge. If both J and K are held low, Q of the JK does not change. This is indicated by Qn. Holding one low and one high will force the 0 and 1 condition. With both high, Q will flip to the state opposite what it was before the clock edge arrived. This is what Qn means.

The D flip-flop is easier to understand. Its Q state takes the D input state after receiving a clock edge.

Both flip-flops will change Q based on the control input state before a clock edge occurs. Control inputs do not directly affect output.*

Clocked flip-flops require an additional time parameter: setup. J, K, and D states must exist a specific minimum number of nanoseconds before a desired clock edge occurs. Too short a setup time will result in skipping a clock period for a Q state change.

An output state change is sometimes called toggling. A JK with both inputs high will toggle on each clock. Output period will be twice clock period, and the flip-flop becomes a divide-by-two device. Connecting Q to D in a D flip-flop will accomplish the same function.

Both flip-flops may have direct set and reset inputs. These override any clock input and may occur at any time. As such they’re sometimes called asynchronous. Terms vary and set is sometimes preset; reset is sometimes called clear.

The next part of this series will look inside clocked flip-flops, present NAND gate circuits, and discuss timing.

*Some old JK types did flip when both J and K changed together with the same state.

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The MK5085/86 as shown in the February, 1977, issue of *Ham Radio*, can be used to make a very simple, inexpensive, and reliable tone encoder. In fact, with the exception of the addition of the transmitter keying relay and driver transistor, the circuits shown have been taken directly from the Mostek data sheet.

A couple of additional comments are in order, however. First, it is usually inadvisable to feed the encoder's output into the microphone line as indicated in the article. The pre-emphasis in the audio amplification circuits will distort the level differential between the high and low group tones, which should be on the order of 3 dB. Thus, the levels actually transmitted, i.e., presented to the decoders at the receiver, will not be correct and decoding difficulty may result. In order to use the microphone input, some kind of rolloff circuitry is necessary, and this adds to the complexity of the overall encoder.

A second problem with using the microphone input is that of background noise; the tones receive as much amplification as the ambient noise. Since the output of the MK5085/86 is on the order of 700 millivolts for the low group and 1100 millivolts for the high group, these high-level signals do not need high amplification.

The chip was primarily designed for telephone application, and to feed a load of approximately 600 ohms. In the telephone application, a convenient output load is the telephone line itself. However, the line is also the power source. With this background, it is easy to see that the chip really prefers to deliver its output right into the power line. Using an independent power source, this characteristic can be ignored. Powered from the radio, however, the effect will be to modulate the entire primary power. In short, the tones will wind up in all kinds of places where you don't want them.

A simple solution is shown in fig. 1. A large capacitor will swamp the tones, provided the power source impedance is low enough; take note that the MK5085/86 is (mostly) CMOS and can generate appreciable spikes.

I prefer the AMI S2559 over the Mostek 5085/86 primarily because of its ability to stand power of up to 15 volts, eliminating the need for the zener and dropping resistor shown in fig. 1. The S2559 is pin-compatible with the Mostek chips and can direct-
ly replace them. It is also driven from the standard 3.58-MHz color-burst crystal. Aside from the wider power tolerance (down to 3 volts or so), there is yet another advantage: The S2559 can be used with either a Class A or 2-of-8 (or 2-of-7) DTMF keyboard. The choice of which Mostek chip to use is dictated by the keyboard key configuration.

A final observation concerns the transmitter keying arrangement shown in the article. It will work, and is simple and inexpensive, but it does have a potential drawback. The two-second time delay is too long (easily changed by choice of capacitance), but it is not very constant or reliable. This could cause problems when the circuit is used with an autodialer or repertory dialer.

In my opinion, a better (and only slightly more expensive) arrangement can be made using the workhorse 555 timer. The circuit, as shown in fig. 2, gives an easily adjustable hang time of from 0.1 second, to about 1.1 seconds, by adjusting the 1 megohm pot in series with the 100k fixed resistor. Once set, the hang time is extremely consistent.

The circuit works as follows. With no key pressed, the MUTE output at pin 10 of the MK5085/86/AMIS2559 capacitor discharges to 1/3 Vcc (approximately one time constant) the 555 triggers, returning the 555 to its previously stable state. The capacitor recharges virtually instantaneously through CR2. If another key is pressed before the capacitor discharges to 1/3 of Vcc, it immediately recharges to full voltage through CR1 and remains there until the key is released.

CR3 is transient suppression for K1. I found it useful to use a relatively low-voltage relay with series resistance because (a) I got a real buy on 5-volt reed relays and (b) by adjusting the resistor the timing circuit will operate down to the lower functional limit of the 555 and the S2559. The LED and its associated series resistor are optional.

If this timing circuit is used with the MK5085/86, an 82k resistor should be used in series with CR1. The MUTE output current is capable of providing 10 μA. The S2559 will provide a couple of mils.) However, since the capacitor never discharges below 1.3 Vcc and because the majority of charging current comes from the 555, the 82k resistor suggested isn’t really required. However, it is cheap insurance to protect an $8.20 IC.

It is preferable to use a pot in series with the encoder output (see fig. 3). This provides both level adjustment and isolation concurrently. Taking the output from a pot in the output emitter line (as shown in the article) is not recommended, as it more greatly affects the impedance the chip is working into. I have found a 25K pot to be about the right value, using it in place of the 20k isolation resistor previously mentioned.

One final thought: A useful addition would be to short out the microphone (or open the microphone audio line) when the encoder is active, eliminating background noise. This can be done in a variety of ways, the most obvious being to use another pole on the keying relay connected to the 555 timer.

Concerning the relay, the 10 mA operating current does not include the relay current. Care must be taken if a 9-volt battery is used; they’re not made to drive current-hungry devices like relays. Also, the 10 mA figure is with Vdd equal to 6 volts, but the circuit shows a 9-volt source. Under these conditions the current will be around 20-25 mA.

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- 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 receiver and Power Out on Transmitter.
- RECEIVE: Better than 25uV sensitivity, 15 POLE FILTER as well as monolithic crystal filter and AUTOMATIC TUNED LC circuits provide superior selectivity - COMPARE!
- HIGH/LOW POWER OUTPUT: 16 watts and 1 watt, switch selected. Low power may be adjusted anywhere between 1 and 16 watts. Fully protected - short or open SWR.
- RF ATT: Live right next to King Kong Repeater and can't operate? With the 2016A You Can - Just flick the RF ATT switch. Only the 2016A has this feature.
- OTHER FEATURES: Dynamic Microphone, built in speaker, mobile/aircraft type dual control system, external 5 pin a cessorary jack, speaker jack, and much, much more. Size 2½ x 7 x ¾. All cords, plugs, fuses, microphone hanger, etc. included. Weight 5 lbs.

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DEALER INQUIRIES INVITED

March 1979
**NEW products**

For literature on any of the new products, use our Check-Off service on page 118.

**NDI 800-channel 2-meter transceiver**

NDI, Incorporated, of Torrance, California, has announced the HC-1400, a new, high-performance 2-meter fm mobile transceiver of advanced design. This microprocessor-controlled, digitally synthesized unit has 800-channel capability within the Amateur 144-148 MHz band, and offers 5- or 10-kHz channel spacing. A fast-acting, single-knob selector shifts LED digital frequency readout in 10-kHz steps; the HC-1400 also has a 100-kHz "speedup" button.

Transmit frequency offsets are pre-programmed, and are switchable to plus or minus 600 kHz. Simplex operation is also available. The HC-1400 can be programmed to hold three transmit-receive frequency pairs in memory, with the capability of instant recall. The transmitter delivers 25 watts output, and requires 13.8 Vdc for operation.

The receiver is sensitive, selective, and uses an FET front-end with three coaxial resonators. The dual i-f has a crystal lattice filter and multiple tuned circuits.

For more information write to NDI, Inc., 22125 ½ South Vermont, Torrance, California 90502.

**350-watt solid-state amplifier**

R.F. Power Labs, Inc., of Kirkland, Washington, is proud to announce their newest amplifier, Model V350. This solid-state vhf amplifier is designed for the Amateur Radio 2-meter band, but can be used (on special orders) over a 5-MHz bandwidth from 135 to 170 MHz. It is capable of producing more than 350 watts of rf output power into 50 ohms when driven with 10 to 15 watts. The amplifier weighs only 23.5 kg (52 pounds), including the built-in ac power supply.

The V350 is priced at $895.00; delivery is one to three weeks. For continuous-duty operation the F135 (115 Vac) or F235 (230 Vac) fan kit should be used, priced at $59.00 each.

For more information, contact R.F. Power Labs, Inc., 11013 - 118th Place N.E., Kirkland, Washington 98033.

**Cushcraft vertical antennas**

Many hams are convinced that to work a lot of DX they need a couple of thousand watts and a monster antenna array. While that undoubtedly helps, where is it written that the ham suffering from a money or space cram can't compete for his day in the DXCC sun?

The three new Cushcraft verticals, the ATV-3, ATV-4, and ATV-5, provide a common-sense solution to a commonplace problem. Specifically designed for the DXer, these antennas provide the low angle of radiation necessary for long-haul DX communications, along with the performance and quality long associated with the Cushcraft name. The ATV-3, ATV-4, and ATV-5 operate over the 10, 15, and 20 meter Amateur bands. The ATV-4 has built-in 40-meter coverage, and the ATV-5 is all set for complete five-band operation.

All antennas feature a built-in PL-259 coax connector and stainless-steel hardware for all electrical connections; all are matched to 50 ohms and rated for a full 2000 watts PEP. Factory-marked tubing and plain English instructions make assembly a snap.

Built to withstand the severest weather, the ATVs feature specially designed, high-Q traps employing large diameter enameled copper wire and solid aluminum, air-dielectric capacitors. The trap forms are manufactured from fiberglass for minimum dielectric loss and high strength.

Available from dealers worldwide,
the ATV-3, ATV-4, and ATV-5 retail for $49.95, $89.95, and $109.95 respectively. For more information and a full-color catalog highlighting the entire Cushcraft antenna line, write to Cushcraft, P.O. Box 4680, Manchester, New Hampshire 03108.

**B&K model 820 digital capacitance meter**

Dynascan Corporation has just announced the introduction of the B&K-Precision Model 820 portable digital capacitance meter. The 820 is a compact instrument capable of measurement over the wide capacitance range of 0.1 pF to 1 farad. Accuracy will greatly exceed the tolerance of most capacitors. The unit features a bright, four-digit LED display for easy reading in laboratories, product lines, or field applications.

The capacitance of virtually any capacitor can be measured quickly and accurately with the 820. Because the accuracy of this unit greatly exceeds the tolerance requirements of most users, required values can be “hand selected.” Matched capacitors can also be singled out for use in bridge circuits and other critical applications. Ten ranges cover from 0.1 pF to 1 farad with resolution up to 0.1 pF.

The 820 allows quick measurement of unmarked capacitors or verification that a capacitor is within tolerance. Virtually any type of capacitor can be measured, from miniature...
ERC PROMISES UP TO THE MINUTE STATE-OF-THE-ART DESIGN AND PERFORMANCE WE’VE DONE IT FOR 1979

FOUR SIMULTANEOUS FILTERS IN ONE FOR UNPARALLELED QRM FREE RECEPTION (SSB & CW) *PLUS A SPECIAL PATENTED CW PROCESSOR*

The brand new SL-56 audio active filter surpasses our SL-55 in both concept and performance. The quality of many components has allowed us to make 16 operational amplifiers. (compared to 6 in the SL-55) into a filter with a quarter to operate any other at a cost only slightly higher than the SL-55. The features of this SL-56 are so advanced from its predecessor that calling it the SL-56A is not justified. Unlike other filters that simply offer a choice of one or two filter types at a time (narrow, bandpass, etc.) SL-56A proves what is really needed—simultaneous action of a 6 pole 200 Hz fixed high-pass filter and a 6 pole 1600 Hz fixed lowpass filter with a 600 Hz fixed notch. 

A 6 pole 1600 Hz fixed lowpass filter with a 600 Hz fixed notch tuned over the 200-1600 Hz range. This 3 filter combination is unbeatable for the ultimate in QRM free SSb reception. Adjacent channel QRM is eliminated on the high and low sides of the same time and does not introduce any fiddling to the desired signal. On CW the SL-56 is a dream. The lowpass, highpass, and notch filters are engaged along with the tunable bandpass filter (400-1600 Hz). Providing the needed action of 4 simultaneous filter types. The bandpass may be made as narrow as 15 Hz (3dB). Additionally, a special patented circuit follows the filter sections which allows only the peaked signal to “gate itself” through to the speaker or headphones. 

Receiver noise and direct signals are rejected. This is not a regenerative, but a modern new concept in CW reception. The SL-55 connects four simultaneous filter types, the SL-56 connects in series with the amplifier receiver output and drives any speaker or headphones with 1 watt of audio power. Requires 115 VAC, easily converted to 12 VDC operation. Colling gray cabinet and wrinkle gray panel.

WARRANTY ONE YEAR FULLY RFI PROOF FULLY WIRED AND TESTED AVAILABLE NOW $75.00 POSTPAID IN THE USA AND CANADA. VIRGINIA RESIDENTS ADD 4% SALES TAX.

ATTN SL-55 OWNERS: THE CIRCUIT BOARD OF THE SL-56 IS COMPLETELY COMPATIBLE WITH THE SL-55 CHASSIS. OUR RETROFIT KIT IS AVAILABLE AT $35.00 POSTPAID.

ERC INTRODUCES A NEW CONCEPT IN THE MEASUREMENT OF VSWR AND POWER ACCEPTED BY THE LOAD

REQUIRES 115 VAC AT LESS THAN 1/16 AMP.

COLLING GRAY CABINET.
WRINKLE CABINET - BRIGHT RED LED DIGITS (.33”).
DECIMAL POINT IS THE PILOT LIGHT.

THE MODEL SL-56A (20-2000 WATTS) AND THE QRP MODEL SL-56A* (0.2-20 WATTS) DIGITALLY INDICATE ANTENNA VSWR UNDER ANY TRANSMISSION MODE - SSB, CW, RTTY, AM, etc. THERE IS NO CALIBRATION REQUIRED. NO CROSS-CONNECTED METER NEEDLES TO INTERPRET. SIMPLY LOOK AT THE READOUT AND THAT IS THE VSWR.

THE PEAK OF THE PEP FOR ANY MODULATION. 50% PEAK TO PEAK, 150% PEAK TO PEAK. DISPLAY UPDATE IS CONSTANT THE Flicker Free As You May Change The Power According To Your Voice. THERE IS NOTHING TO ADJUST THE TRANSMITTER.

POWER ACCEPTED BY THE ANTENNA.

THE POWER ACCEPTED BY THE TRANSMITTER IS TALKING UP TO DISPLAY DECAY TIME IS ABOUT ONE SECOND.

THE PEAK DISPLAYED IS THAT WHICH IS ACCEPTED BY THE ANTENNA. (FORWARD LESS REFLECTED)

THE POWER DISPLAYED IS THE PEAK OF THE PEP FOR ANY MODULATION. THIS IS THE POWER VOLTAGE AT THE TRANSMITTER WHICH YOU MAY CHANGE THE POWER ACCEPTED BY THE ANTENNA. 

SL-65

SL-56

VSWR INDICATOR

WARRANTY ONE YEAR

SL-65 NET POWER INDICATOR

* TWO DIGIT DISPLAY SHOWS VSWR TO AN ACCURACY OF .1 FOR VALUES FROM 1.0 AND 2.0. ACCURACY IS +1 PERCENT FOR VALUES FROM 2.0 TO 3.0 AND +3 PERCENT FROM 3.0 TO 4.0. FROM 4.1 TO 6.2 THE INDICATION MEANS THAT VSWR IS VERY HIGH.

* FOR VSWR VALUES NEAR 1.0, THE POWER RANGE FOR A VALID READING IS 20 - 2000 WATTS OUTPUT. FOR HIGHER VALUES THE UPPER POWER LIMIT FOR A Flicker Free Valid Reading Is SOMETHING LESS (35-1000 WATTS FOR VSWR AT 2.0). 

* SIDE BY SIDE POWER LEVELS BY 100 TO DETERMINE THE PERFORMANCE OF THE SL-56A QRP MODEL.

PRICE: $189.50 POSTPAID IN USA & CANADA. CANADA RESIDENTS ADD 6% SALES TAX.

BOOKLET AVAILABLE AT $2.00 REDEMPTIBLE TO
WANT TO PURCHASE.
* PATENT PENDING.

Yaesu two-meter transceiver

A new state-of-the-art, two-meter, all-mode transceiver, the FT-225RD, has been added to Yaesu's quality line of Amateur Radio equipment.

The new transceiver covers the entire 4 MHz and provides for USB, LSB, CW, and FSK. Power output is variable, 1-25 watts. Squelch, VOX, PTT, semi break-in CW with side tone, and tone burst are standard features of the FT-225RD. A superb noise blanker permits mobile ssb operation, and a discriminator zero-center meter allows precise tuning of discs to pole-mounted power types. Even the small amounts of capacity encountered in cable or switches can be measured.

For production-line applications, the 220 is an excellent means of pre-testing critical capacitors or accurately adjusting trimmer capacitors. The simplicity of operation allows even untrained workers to be quickly instructed in proper operation. To facilitate fast incoming component sorting and selection, slot-type front panel lead-insertion jacks are used. The slot jacks eliminate the time-wasting step of guiding a capacitor lead into the type of small lead insertion holes commonly found on other instruments.

In classroom applications, the 220 can be used to verify capacitor network calculations by measuring the actual value of a network. The operation of a variable capacitor can also be demonstrated, as can the temperature effects on a capacitor.

The B&K-Precision Model 820 comes with a 26-page detailed manual. Optional accessories include the BP-28 rechargeable battery pack, BC-28 charger, and LC-28 carrying case. The 820 is now available for immediate delivery at local B&K-Precision distributors at a price of $130.00. For additional information, contact B&K-Precision, Sales Department, 6460 West Cortland Street, Chicago, Illinois 60635.
fm signals. Repeater splits are the
standard 600 kHz; however, any
split up to 1 MHz is possible with op-
tional crystal. Provision has been
made for up to eleven fixed channels
using optional crystals.

The transceiver uses high-quality,
plug-in circuit boards throughout,
and an optional memory unit enables
the storage and recall of any fre-
cuency within the range of the unit.
This allows instant programmable
QSY to a favorite repeater or calling
frequency with just a flick of the
switch. The digital frequency is ac-
curate to 0.1 kHz or to 1 kHz with the
FT-225R, which offers the analog dial
readout only and at slightly less cost.

A built-in power supply provides
taps for operation on 100/110/117/
200/220 and 234 volts 50/60 Hz; dc
operation covers 11.5 to 16 volts,
negative ground at 6.5 amps on
transmit, 1.2 amps on receive.

An attractive, four-color brochure
is available at your nearby authorized
Yaesu dealer or from Yaesu Elec-
tronics Corporation, 15954 Downey
Avenue, P.O. Box 498, Paramount,
California 90723.

second generation
VMOS

Siliconix has introduced the first of
a new family of high-current and
high-voltage VMOS power fets, serv-
ing both analog and digital applica-
tions. The VN84GA has a rated out-
put of 12.5 amperes, 80 volts, and 0.4
ohms — about sixfold current in-
crease over previously available units.

The VN84GA, with input power in
the microwatt range, produces up to
80 watts output at lower frequencies,
or as much as 50 watts at 30 MHz.
Heavy Duty Power Supplies from ASTRON
High Quality • Rugged • Reliable

SPECIAL FEATURES
- SOLID STATE ELECTRONICALLY REGULATED
- FOLD BACK CURRENT LIMITING Protects Power Supply from excessive current & continuous shorted output
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PERFORMANCE SPECIFICATIONS ASTRON
12 AMP REGULATED POWER SUPPLY Model RS-12A
9 Amps continuous
9 Amps continuous

This input power is several orders of magnitude less than that required by Darlington bipolar output, yet the new VMOS retains the benefits of fast switching, high gain, and high-input impedance that are characteristic of lower power VMOS units. The VN84GA also provides inherent fail-safe operation with no failure from secondary breakdown and no thermal runaway; it has the ability to current limit by controlling the gate voltage.

The VN84GA is meant to satisfy a designer's need for a universal high-power device for switching regulators, motor controllers, audio amplifiers, and as an interface for microprocessor logic outputs. It interfaces directly with CMOS, TTL, DTL, and MOS logic families.

The VN84GA is available in a TO-3 package from franchised Siliconix distributors. It is priced at $19.76 in single quantities. For additional information, contact Siliconix Incorporated, 2201 Laurelwood Road, Santa Clara, California 95054.

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Ham Radio's Bookstore
Greenville, NH 03048

Newnes Radio and Electronics Engineer's Pocket Book
In this, the latest edition of this handy pocket book first published in 1940, author H. W. Moorshead has succeeded in keeping this standard reference book up to date. Obsolete material has been eliminated to make room for the most current information. The section on ICs, in particular, has been expanded to cover the many recent advances in the field. Among the other topics contained in the book are transistor and diode data and busing diagrams, transistor circuits, transformer data, and many tables of information useful to the working technician and engineer. Pocket-notebook size (5 inches long and 3 inches wide), hard cover, 191 pages. $5.95, plus $1.00 per order for shipping, from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048; order NR-PC.
soldering-tool holder and mini-soldering station

EDSYN, Incorporated, of Van Nuys, California, announces a low-cost soldering tool holder which also serves as a compact soldering station. Designated the Idle-Rest TL194 General Purpose Holder, the unit holds most available soldering tools due to its specially designed tool cradle. The exterior of the holder remains within safe skin-touch temperature limits because of the low heat transfer characteristics of the molded Bakelite housing. Also, an efficient heat shield and ventilating-grill system helps keep heat concentrated to the soldering tip area.

The holder sits securely on the bench top without tipping. If desired, the integral fastener nut may be used for permanent installation to the bench top.

In addition, the holder contains a supply of desoldering wick and solder (included with purchase), parts or tip-storage tray, and a large tip-cleaning sponge which self-wets at the touch of a finger. The sponge may be removed if desired to obtain an additional storage tray for components or soldering supplies.

For more information, contact an authorized Edsyn distributor; for the name of the one nearest you, write Edsyn, Incorporated, 15958 Arminta St., Van Nuys, California 91406.

Mirage MP1, peak-reading wattmeter

Mirage Communications is now offering the MP1 high-frequency (1.8-30 MHz) peak-reading wattmeter. It is designed to provide the
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Amateur Radio operator with a ver-
satile fixed or portable wattmeter,
without having to use cumbersome
plug-in or add-on accessories.

Three power ranges are available
(25, 200, and 2000 watts), providing
reading convenience whether you
use it for low-power operation or with
the biggest DX setup. The wattmeter
will read both forward and reverse
power at the flip of a switch. The
MP1, like its vhf predecessor, the
MP2, has the ability to display either
average or peak power reading. The
peak-reading feature is a must for
ssb transmitters. The MP1 also will
display swr, which is measured di-
rectly, without having to use extra
charts or graphs. For ease of installa-
tion, the MP1 has a removable r-f
coupling unit that may be placed up
to 1.2 meters (4 feet) from the in-
dicator.

The MP1 is portable, containing
the latest in low-power ICs and pow-
ered by a 9-volt battery. For long-term,
fixed-station operation, an optional
ac adaptor is available. A low-battery-
voltage indicator has been built into
the wattmeter to indicate when the
battery needs changing.

For more information please con-
tact your local dealer or Mirage Com-
munications, P.O. Box 1393, Gilroy,
California 95020.

J. W. Miller catalog
A new 100-page catalog with
specifications for more than 5,000
coils, filters, and communications
essentials is now available from the J. W. Miller Division of Bell Industries, Compton, California.

Newest additions to their catalog include direct reading SWR/power meters, an rf speech-processor, and coaxial switches. Included also is the broad line of high pass, lowpass, audio, and ac power-line filters.

Catalog 79 gives detailed specifications for rf coils, chokes, filters, and related communication components. To assist in selection, coils are categorized by frequency from 0 through 500 MHz in the table of contents. Schematic diagrams for all shielded and unshielded coils, showing adjustment accessibility, are given.

For additional information, contact Jerry Hall, Operations Manager, Bell Industries, J. W. Miller Division, 19070 Reyes Avenue, P.O. Box 5825, Compton, California 90224.

Astatic amplifying microphone

A microphone designed for repeater-control and phone-patch operations has been introduced by the Astatic Corporation, Conneaut, Ohio. It's the new Astatic T²M amplified microphone with touch-pad encoder.

For fingertip convenience, the touch pad encoder is an integral part of the microphone itself. It has a tactile twelve-key keyboard and provides visual feedback through a front-mounted LED.

Priced comparably to a conventional microphone, the T²M is of an
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Coaxial RF Probe for Frequency Counters and
Oscilloscopes That Lets You Monitor Your
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your transmission line and plug the output
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HAM SHACK, the CoaxProbe* eliminates
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A SPECIAL METHOD OF SAMPLING
keeps output relatively constant with a wide
variation of power. Power output of 8 watts
gives .31v out, while 800 watts will give
1.8v out. (rms 3-30 mhz.) 2000 watts PEP
rating too!

*Trademark of CoaxProbe Co. for rf sampling device.

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WATTS output, unsurpassed
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The attractive Cycolac housing re-
sists abusive treatment. It also fea-
tures shielding to reduce external in-
terference and rf feedback. The T2M
operates from filtered 6 to 16 volt dc
sources. Included is an eight-foot,
three-conductor (one shield) cable
and coiled cord.

For additional information on the new
Astatic T2M amplified microphone
with touch pad encoder, write
The Astatic Corporation, Conneaut,
Ohio 44030.

communications
equipment catalog
Harrison Radio is pleased to an-
ounce the availability of the all-new
1979 Harrison Radio Communications
catalog. The catalog has 120 pages of
illustrations, descriptions, specifi-
cations, and prices covering several
hundred transceivers, antennas, and
accessories for Amateur Radio,
marine radio, business-band radio,
public-service-frequency equipment,
and CB.

The catalog is free and available by
writing to Harrison Catalog, 22 Smith
Street, Farmingdale, New York
11735.

receiver preamplifier
Telco Products Corporation an-
nounces two new mobile in-line re-
ceiver preamplifiers. Known as Mod-
eIs VHF 144 and UHF 450, they are
specifically designed for amateur,
police, emergency, business-band,
and Class A CB transceivers.

These new computer-designed
preamplifiers boost receiver sensi-
tivity. A unique rf-sniffer circuit al-
ows the preamplifier to sense when
the transceiver is keyed, and switches
the preamplifier off during the trans-
mit cycle. This feature allows the pre-amplifier to be connected directly into the coax line with no modification to the existing mobile system.

The VHF 144 has 20 dB gain over a 5-MHz band in the frequency range of 140-180 MHz. Model UHF 450 features 10-12 dB gain over any 1-MHz band in the frequency range of 400-512 MHz. The VHF 144 will safely handle transmitter output of 40 watts, and UHF 450 will work with 100-watt transmitters.

Suggested retail price for the VHF 144 is $49.95; and for the UHF 450 is $59.95. For additional information, contact Telco Products Corporation, 44 Sea Cliff Avenue, Glen Cove, New York 11542.

solderless breadboards

Continental Specialities Corporation is probably best known for their popular line of versatile and easy-to-use solderless breadboards. One of the best is their Proto-Board model 203A, which is actually more than a solderless breadboard. In addition to offering a very large socket-type solderless breadboard array to work on, Proto-Board 203A has three separate, regulated, power supplies built into the box. They offer 5 Vdc at 1 percent regulation and up to 1 amp of output current, and ±15 Vdc (nominal) supplies, each good for 0.5 amp and each independently internally adjustable over a 9-18 Vdc range.

The statistics on the Proto-Board 203A are impressive. The whole package measures only $25 \times 17 \times 8$ cm ($9.8 \times 6.6 \times 3.3$ inches). It features 2250 tie points for a 24-DIP (14-pin package) capacity. The ±5, ±15 Vdc and ground connections are available at four five-way binding posts. The power switch and a power pilot are adjacent to the power supply terminals.

The PB-203A is designed for 117 Vac, 60-Hz operation and carries a suggested amateur net price of $129.95. For further information, contact Continental Specialties Corporation, 70 Fulton Terrace, New Haven, Connecticut 06509; or phone (203) 624-3103.
Repeater Jammers Running You Ragged?

Here's a portable direction finder that REALLY works--on AM, FM, pulsed signals and random noise! Unique left-right DF allows you to take accurate (up to 2°) and fast bearings, even on short bursts. Its 3dB antenna gain and 0.6μV typical DF sensitivity allow this crystal-controlled unit to hear and positively track a weak signal at very long ranges--well beyond the built-in RF gain control with 120 dB range permits positive DF to within a few feet of the transmitter. It has no 180° ambiguity and the antenna can be rotated for horizontal polarization.

The DF is battery-powered, can be used with accessory antennas, and is 12/24V for use in vehicles or aircraft. It is available in the 140-150 MHz VHF band and/or 220-230 MHz UHF band. This DF has been successful in locating malicious interference sources, as well as hidden transmitters in "T-Hunts", ELTs, and noise sources in RFI situations.

Price for the single band unit is $195, for the VHF/UHF dual band unit is $235, plus crystals. Write or call for information and free brochure.

GILFER SHORTWAVE
Dept. HR3, Box 239, Park Ridge, NJ 07656

16-POLE R-4C SSB!

Improve the early stage selectivity of your Drake R-4C while adding 9 additional points (total 16) with an internally mounted, triaxial set of first IF crystal filters. Reduce RFI, leakage, overload, Ideal for DX and contest work. Overall shape factor better than 1.8. Maximum skirt selectivity with maximum instability. Total bandwidth with CF-2.1kHz: 2100 Hz at dB, 2900 Hz at 60 dB. Minimum variable band-width. Other bandwidths available. Filter set can be mounted in receiver and relay switched with our kits which start at $5.00. 3 and 4 filter switching options can include our CF-600 kHz and/or existing 8kHz first IF filter, all internally mounted, controlled from rear or front panel. USB and LSB CF-2.1kHz pair is $120.00 per set. Money back if not satisfied. Auto $3 shipping per order, $6 overseas air. Dealer inquiries welcome.

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GILFER SHORTWAVE

160-90-40-20-15-10 bands 2 trap -- 197 Ht. with 90 ft. RG213 -- coax -- connector - Model 19977BU -- $49.95
- 19950-20-15-10 bands 2 trap -- 217 Ht. with 90 ft. RG213 -- coax -- connector - Model 19978BU -- $54.95

BAND FULL PRICE FOR POST PAID INSURED DEL IN USA (Canada is $5.00 extra, for postage, handling, insurance, mailing, etc.) or order using VISA Bank American - MASTER CHARGE - AMEX - EXPRESS. Give number and ass. date. Ph 1-209-203-5233 9AM - 6PM week days. We ship in 2-3 days. PRICES MAY INCREASE SO--ORDER NOW AND SAVE! All guarantee for 1 year. Money back trial! Made in USA. FREE INFO AVAILABLE FROM:

WESTERN ELECTRONICS
Dept. AR-3 Kearney, Nebraska, 68847
Antenna Sale!

<table>
<thead>
<tr>
<th>HY-GAIN</th>
<th>Regular</th>
<th>Special</th>
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<tbody>
<tr>
<td>TH6-DXX Super Thunderbird</td>
<td>$296.95</td>
<td>$209.95</td>
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<tr>
<td>TH3-MK3 3 ele. 10, 15, 20 Mtr. beam</td>
<td>219.95</td>
<td>179.95</td>
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<tr>
<td>Hy-Quad 2 ele. Quad 10, 15, 20 Mtr.</td>
<td>229.95</td>
<td>199.95</td>
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<td>TH3-Jr. 3 ele. 10, 15, 20 Mtr. beam</td>
<td>144.50</td>
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<tr>
<td>18 HT Hy-Tower 10-80 Mtr. Vertical</td>
<td>299.95</td>
<td>259.95</td>
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<td>14AVQ/WB 10-40 Mtr. Trap Vertical</td>
<td>67.00</td>
<td>57.00</td>
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<td>18AVT/WB 10-80 Mtr. Trap Vertical</td>
<td>97.00</td>
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<tr>
<td>205 BA 5 ele. &quot;Long John&quot; 20 mtr beam</td>
<td>289.95</td>
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<td>203 3 ele. 2 Mtr. beam</td>
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<td>205 5 ele. 2 Mtr. beam</td>
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<td>208 8 ele. 2 Mtr. beam</td>
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<td>214 14 ele. 2 Mtr. beam</td>
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<td>Classic 33 3 ele. 10, 15, 20 Mtr. beam</td>
<td>304.75</td>
<td>219.95</td>
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<td>Classic 36 6 ele. 10, 15, 20 Mtr. beam</td>
<td>392.75</td>
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<tr>
<td>TA-33 3 ele. 10, 15, 20 Mtr. beam</td>
<td>264.00</td>
<td>199.95</td>
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<td>TA-36 6 ele. 10, 15, 20 Mtr. beam</td>
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<tr>
<td>TA-33 Jr. 3 ele. 10, 15, 20 Mtr. beam</td>
<td>197.00</td>
<td>149.95</td>
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<tr>
<td>TA-40KR 40 Mtr. add on</td>
<td>119.50</td>
<td>89.95</td>
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<tr>
<td>CUSHCRAFT ATB-34 4 ele. 10, 15, 20 Mtr. beam</td>
<td>259.95</td>
<td>209.95</td>
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<tr>
<td>ARX-2 2 Mtr. Ringo Ranger</td>
<td>36.95</td>
<td>32.95</td>
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<tr>
<td>A147-20T 2 Mtr. Twist</td>
<td>59.95</td>
<td>52.95</td>
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<tr>
<td>A144-10T 10 ele. Twist 2 Mtr.</td>
<td>39.95</td>
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<td>59.95</td>
<td>52.95</td>
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<tr>
<td>A147-11 11 ele. 146-148 MH2</td>
<td>34.95</td>
<td>30.95</td>
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<tr>
<td>A147-22 Power Pack 22 ele.146-148 MH2</td>
<td>99.95</td>
<td>88.95</td>
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<tr>
<td>A432-20T -430-436 MH2 20 ele. TWIST</td>
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<td>49.95</td>
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<td>ATV-4 10,15,20,40 MTR Vertical</td>
<td>89.95</td>
<td>79.95</td>
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<tr>
<td>ATV-5 10,15,20,40 MTR Vertical</td>
<td>109.95</td>
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<td>34.95</td>
<td>30.95</td>
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<tr>
<td>WILSON System One 5 ele. 10, 15, 20 Mtr. beam</td>
<td>274.95</td>
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<td>System Two 4 ele. 10, 15, 20 Mtr. beam</td>
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<td>189.95</td>
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<tr>
<td>System Three 3 ele. 10,15,20 MTR beam</td>
<td>179.95</td>
<td>159.95</td>
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<tr>
<td>WLV-1 Vertical 10-40 MTR Vertical</td>
<td>79.95</td>
<td>69.95</td>
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**CUSHCRAFT ATB-34**

**HY-GAIN**

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**ARX-2 2 Mtr. Ringo Ranger**

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**System Three 3 ele. 10,15,20 MTR beam**

**WLV-1 Vertical 10-40 MTR Vertical**

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More Details? CHECK — OFF Page 118

March 1979
10.7 MHz FILTERS, 8 POLE
- X107-A 14 kHz NBFM $51.55
- X107-B 16 kHz NBFM $51.55
- X107-C 32 kHz WBFM $51.55
- X107-D 38 kHz WBFM $51.55
- X107-E 42 kHz WBFM $51.55

10.7 MHz FILTERS, 4 POLE
- X107-GI 30 kHz NBFM $23.95

OSCILLATOR CRYSTALS, 1 kHz to 200 MHz available. Write for Data Sheet and prices.

SOCKET (for Pre-Selector Filter Amplifiers)
- MMc432
- MMc432.144 $340.00
- MMc432.50 $330.00
- MMc432.28 $230.00

RECEIVE CONVERTERS
- MMt43R-ATV Prime Receiver
- Output Power 10 W PEP

Specifications:
- Drive, 10 Meters
- Receiver N.F. 3.0 dB
- Receiver gain 10 dB typ.
- Prime Power 12 V D.C.

Shipping: $3.50 per filter

432 MHz SSB TRANSVERTERS

Use your HF Transceiver on the 432 MHz band with the addition of the MMc432 Linear Transverter. The MMc432 operates on all models: SSB, CW, AM, FM. It contains the linear transmit up-converter and the receive down-converter. An external PIN diode TR connects to your Transceiver T/R line. The MMc432 is for HF and is compatible. Add the 70/MBCP MULTIBEAM and operate direct into OSCARS J & B. Write for application note.

Specifications:
- Output Power 10 W PEP
- Drive, 10 Meters
- Receiver N.F. 3.0 dB typ.
- Receiver gain 10 dB typ.
- Prime Power 12 V D.C.

Shipping: $3.50 per filter

OSCAR-J FILTERS

Suppress 2m Tx Third Harmonics. Low 2m loss (0.5 dB typ.). High loss at 435 MHz (30/40dB).

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RECEIVE CONVERTERS

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POWER 12V D.C. Shipping $2.50

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MMc144 N. F. 2.8 db typ. $59.95
MMc232 N. F. 3.0 db typ. $69.95
MMc432 N. F. 3.3 db typ. $69.95
MMc432-S N. F. 3.0 db typ. $84.95
MMc438/ATV Ch 2 or Ch 3 IF $94.95
MMc1296 N. F. 3.5 db typ. $84.95

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- • Selectable Gate Times, 1 & 1 sec.
- • Crystal Time Base (1 ppm after cal.)

OPTIONS: Portable w/ Ni-Cad Battery (Built-in Charger) $39.95
Crystal Oven (1 ppm to 50'C) $39.95 Handle $5.00

Order direct from factory. Add $3.00 for shipping. $10.00 extra for C.O.D., and 7% sales tax in N.Y. State. Payment by certified check, money order, Master Charge. VISA. Credit rated company. 30 days returnable only unassembled.

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TELETEYPE EQUIPMENT for beginners and experienced operators. RTTY machines, parts, supplies. Beginner's special: Model 15 Printer and demodulator $139.00. Dozen black ribbons $6.50; case 40 rolls 11/16 perf. tape $17.50 FOI. Atlantic Surplus Sales, 3730 Nautilus Ave., Brooklyn, N.Y. 11224. Tel: (212) 372-0349.

RTTY - Bandpass active filter 5625/2295 Hz. Kit $10.95, with $16.95 pade. SASE for info. Nat Sinettemic Electronics, Tavares, FL 32778.

440-MHz — Regency HR-440 like new $220.00. SBE-450 very clean $140.00. Maxwell test item 70 items test equipment, FM, Ham mic, write K6KZT, 2255 Alexander, Los Osos, CA 93402.

MOTOROLA low band Mocom 70 29.6 MHz, no accessories $275. Motorola low band PT300 SW portable 52.5 MHz with charger $300. Motorola SSR 2 solid state 30W repeater 146.22/146.88 MHz $600. Homebrew "ACCU-MEMORY" keyer with pad $125. Basket Case H700 VHF $50. Eric Mhet, VE2AS, 6716 Sherbrooke W., Apt. 201, Montreal, Que. H4B 1P2, (514) 482-4668.

TEKTRONIX 310A oscilloscope, mint condition, calibrated, $175.00 or trade for test equipment, commercial FM VHF, Ham gear, etc. H. Beene, HLSL, P.O. Box 5762, Santa Fe, NM 87502, (505) 471-5458.

THE MEASUREMENT SHOP has used/reconditioned test equipment at sensible prices; catalog 2 West 22nd St., Baltimore, MD 21218.

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AUTHORIZED DISTRIBUTOR F9T Antennas, Microwave Modules, RSW Products' new tandem reflector, 19 element, 432 MHz Yagi — Radio Clinic — N2MB (former-


TUNGSTEN CARBIDE CIRCUIT BOARD DRILLS. 12 assorted sizes for $3.95 p/l. R. R. Design, Box 41, McMINNIVILLE, OR 97128.

SWAN 250, TV-2, 14-117 power supply, bandswitching, manuals, mike, 250 Watts input on both 6 and 2 meters. $375.00. Fred Chenger, 640 Grove Avenue, Galion, Ohio 44833, W9UJ.

PORTA PAK — Make your FM mobile a self-contained portable. Models in stock for most popular makes. $4.50 amp hr model $67.50, 9 amp hr $88.00. Charged included, shipping extra. P.O. Box 67, Somers, W13171.

TEST EQUIPMENT CATALOG listing used Tektronix, HP and GR equipment at bargain prices. P.O. Box 8699, White Bear Lake, MN 55110. Price $1.00 refundable with first order.

WANTED: 150 MC HT RCA, Comco, Repco type with plug-in modules, 73.75 TV camera tube, 2 meter cavities, K6KZT, 2255 Alexander, Los Osos, CA 93402.

WANTED — Radio transcription discs. Any size or speed. Larry, W7FZQ, Box 724, Redmond, Washington 98052.

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Here's R.U.N.'s partial reward listing: Accessory: keyer ( ), tuner ( ), amp ( ), other ( ). Antenna mobile ( ), fixed vertical ( ), tower ( ), other ( ). A second rig: mobile ( ), GRP ( ), home special ( ). A new band: 2 meters higher ( ), to 144 MHz. Solid state HF for your ARS ( ). We can help you select from top amateur equipment brands. And will even mail along a "Thank You" card from U to U. (Or anyone else.) Don't just leave to North Pole Nick (or the Bunny) again. R.U.N. in or jingle the telephone bells now to Radios Unlimited, 1780 Easton Avenue, P.O. Box 347, Somerset, New Jersey 08873.

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MICROWAVE low-noise TWT preamplifiers for satellite/hamateur bands: 2-6 GHz, 1150 MHz gain; 110 volt 50 Hz power; TNC connectors; used units tested okay but sold as-is. $200 plus $4 shipping. M. Murseck, 5511 West Jeromy, Milwaukee WI 53219.

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Coming Events

RADIO EXPO '79 September 15th and 16th, 1979, Lake County Fair Grounds, Routes 120 and 45, Grays Lake, Illinois. Manufacturers' displays, flea market, seminars, ladies programs. Advance tickets: $2.00. Write EXPO, P.O. Box 305, Maywood, IL 60153.

MASSACHUSETTS: University of Lowell Wireless Society 5th annual ham radio auction. Friday, March 20, 6 to 8 pm. Doors open 6 pm, auction begins 7 pm. Light refreshments. Directions: From Route 495 take exit 17 west (Route 16) to East Boston Street, right on East Boston Street, left on Cambridge Street, left on West Street, left on Cambridge Street. For more information: Robert, 825-0933.

ROCHESTER Hamfest & NY State ARRL Convention, May 25-27. Add your name to mailing list: Send QSL to Rochester Hamfest, Box 1388, Rochester, NY 14603. Phone: (716) 424-1100.

MINNESOTA: Rochester ARC and the Rochester Repeater Society's Hamfest. Saturday, April 7, St. John's School Gymnasium, 490 W. Center St., Rochester, MN 55901.

MICHIGAN: ARRL Great Lakes Division Convention and Hamfest, Muskegon Community College, Muskegon, March 30-31. Exhibits, technical forums, large swapshop, dining facilities. Friday evening — "Ham Hospitality" at Muskegon Ramada Inn courtesy of MAARC, and a Wolf Hington. For information: MAARC, P.O. Box 369, Muskegon, MI 49443 or W4GK (818) 722-13789.

OHIO: First Annual Lake County ARA Hamfest, April 1, 8:00 AM to 4:00 PM, Lake County Armony, N. E. corner Painesville Fairgrounds, Rt. 20, Painesville, (35 miles east of Cleveland). All info for exhibitors and flea market, 1:00 PM auction. Reserved tables — full $200, half $100. Admission: $2.00. Under 12 free. Talk in on 146.2282. For info: RARC, c/o KITS, 2514 N. 4th Ave., Rochester, NY 55901.

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100W Input, CW Filter

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OMNI-J & heavy duty magnet mount complete $49.95
TRIEX W-51 FT Self Support Tower (Reg. $850.00)
Your Cost (F.O.B. California) $275.00
Tonna 99FT Antennas 144/161 $79.95
RIW 432/192 $59.95
KLITZING VHF-UHF Amplifiers
2M 10W In — 100W Out $198.00
432 10W In — 50W Out $198.00
BIRD 43 & Slugs, UPS Paid Stock
Microwave Modules 432-285, UPS Paid $329.00
Telrex TBSEM, In Stock $415.00
New Palomar Engr. Transceiver Preamp $89.50
Bencher Paddles... $39.95; Chrome $49.95
ETO 76 Amplifiers Stock
Lunar 6M 2M-220 In Line Preamps $49.95
Lunar 2M Amp 10-80 with Preamp, UPS Paid $198.00
Janel QSA 5 $41.95
CDE HAM-3... $129.00; HAM-X $249.00
VHF Engrs blue line amps Stock
Cetron 572B $29.50
Ampex 6146B... $7.95
Motorola HEP 170 $0.29 ea.
Malloyy 2.5A/1000 PIV Epoxy Diode... $0.19 ea.
Aerovox 1000PF/500V Feed thru $1.95
GE 6146B or 8950 $7.95

Technical books: AMECO, ARRL, Sams, TAB, Reader, Radio Pub., 50 many others Call

NEW Belden 9405 (2#16) 6 wire Rotary Cable $0.26/ft.
8448. Std 8' wire Rotor $0.16/ft.
9888. double shield RG8 Foam $0.39/ft.
8214. RG8 Foam... $0.25/ft.
8237. RG-8 $0.21/ft.
8267. RG-213 $0.25/ft.
Amphonol Silver Plate PL259 $0.59
TIMES 1/. Foam Hardline $0.65/ft Connectors $15.00 ea.
BERKEK RG 6X 52 ohm. kw $0.16/ft.
Consolidated HD-16 Galv. Tower, 10' Sec $29.95
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Alliance HD73 Rotor $109.95
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TEXAS: Midland Amateur Radio Club's annual Swapfest, March 18, Midland County Exhibit Building. Door prizes. Talk in on 146.16/146.76. Preregistration: $4.50 to Midland ARC, Box 4401, Midland, TX 79701 or $5.00 at door.

WISCONSIN: Tri-County ARC Hamfest, March 18, Jefferson County Fairgrounds, Jefferson. (Formerly at Whitewater.) Advance tickets $1.50. Reserve 1-18 tables $2.00 advance; 6 tables $1.00. Send SASE to G. M. Eisenbrant, WA8VYL, 711 East St., Fort Atkinson, WI 53538.

WISCONSIN: Madison Area Repeater Association's 7th annual Swapfest, March 18, Dane County Expo Center Forum Building, Madison. Open 7:AM for exhibitors and sellers, 8:00 AM open. Admission: $1.50 advance, $2.00 door. Twelve and under free. Tables $2.00 advance, $3.50 door. Flea market, plenty of parking, overnight camping and hotels within walking distance. Door prices. All-you-can-eat pancake breakfast and Bar B Q lunch. Free movies throughout the day. Talk in on WRIAT - 146.16/76. For reservations and info: MARA, P.O. Box 3403, Madison, WI 53704.

FLORIDA: Ninth Annual Florida Swapfest, Ft. Walton Beach Fairgrounds, Monday and Tuesday, 8:00 AM to 4:00 PM. QCWA, MARS, ARES meetings. Talk in 52 and 1979. For info: CARC, Box 873, Ft. Walton Beach, FL 32546.

GSP PARTY - NEW ZEALAND: ZL3 Canterbury Chapter of the International 1010 net. From 0000 GMT Saturday, March 31, 1979 to 1200 GMT Sunday, April 1, 1979 (36 hours). Contact each station only once on the ten meter band only: exchange call sign, name, QTH, 10X number, if any, Canterbury Chapter number (if held) and your own local Chapter name and number, if any. Awards include: trophy to highest score; trophy to highest score; and a pennant to the highest scorer in each U.S., Canadian, Australian, Japanese, and New Zealand call area, to each Central and South America, to each European, to each African, and to each Pacific Ocean zones. Logs must be received no later than May 15, 1979. Please write clearly, show name, call area, and address and 10X number. Send logs to Z3ME - C. M. Smythe, 198 Greers Road, Christchurch 5, New Zealand.

TEXAS: "Luckenbach DXpedition", 0800 CDT May 12, 1979 to 1200 CDT May 13, 1979, continuous Call W5LXC. CW: 7.110 MHz and 21.110 MHz; FM: 52.525 MHz, 29.6 MHz, and 145.525 MHz; SB: 3.9, 7.235, 14.285, 21.280, 28.625, 50.100, 50.195, and 144.200 MHz. Luckenbach sits nestled between two small rivers in the heart of the Texas hill country. Virtually untouched by modern civilization, this 1870's cowtown got its first telephone only last year. Luckenbach's general store, saloon and post office remain as they were a century ago. Visit and try their world famous limberlost and rustic treats. For reservations contact: Luckenbach, P.O. Box 62, Luckenbach, TX 78642.

Stolen Equipment

ICOM IC-701, serial number 8000-2786: ICOM matching power supply serial number E7802647. Yaesu FT-101, serial number 3K7291710, plus matching accessories. Equipment packed in two bags and checked in at British International, Inc., at Miami International Airport on January 6, 1979. It is believed that the equipment never reached the airplane and will probably be sold locally. Any information, please advise Miami Police Department and/or Thomas Roessler, PY2DFR, Caixa Postal 6742, 01000 Sao Paulo, SP, Brazil; telephone 268-7232.

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Ham Radio's new Log Book is a tool you just can't do without. With room for over twice as many entries as other popular log books, this log covers more than 2,000 contacts! There's spare ruling too for all FCC-required information, plus extra space for the name and address of each station you contact. For contesters, there is a consistent 30 entries per page for easy scoring. In addition there is a handy frequency spectrum chart showing the exact privileges for each station license band. Privileges from novice to extra, plus a listing of all worldwide amateur prefixes currently in use. And, it's all spiralbound to lie flat on your operating table. This is unquestionably the best log book value anywhere! 8 1/4 x 11 80 pages 1978.

Spiralbound $1.75
Sure you can buy a cheaper linear ... 
But is that really what you want?

You can buy a so-called "maximum legal power" linear for quite a bit less than the price of an ALPHA. What makes the ALPHA worth more ... or the other model less?

TALK TO AN ALPHA OWNER — Notice how scarce used ALPHA's are? Owners are rarely willing to part with them, and will be delighted to tell you why.

CHECK ETO's TWO-YEAR (limited) WARRANTY — Others give you 90 days. But EIGHT TIMES as much protection is only part of the ALPHA warranty story: there's also a clear message about durability.

NO ALPHA 76/374/77D OWNER HAS EVER BURNED UP A POWER TRANSFORMER, despite our No Time Limit (NTL), full power key down rating. Maybe it's because our 76A transformer is nearly TWICE the size of those in competitive desk-top amplifiers and is cooled by ETO's full-cabinet, ducted-air system to boot.

LOOK INSIDE AN ALPHA — the difference in quality and ruggedness is conspicuous. Big coils ... axial-flow ceramic triodes thoroughly cooled by a centrifugal blower ... yet ETO's new acoustic-isolation blower system makes the ALPHA 76A series now even quieter than ever before.

EFFICIENCY, versatility, ease of operation, resale value — the story of ALPHA superiority goes on. Before you decide on a new linear, get all the details from your dealer or ETO direct. And ask for our free guide, "Everything You Always Wanted to Know About (Comparing) Linears ... But Didn't Know Whom to Ask." Finally, do talk to an ALPHA owner ... look for him on top of the nearest pile-up.

ALPHA/VOMAX can boost the "talk power" of any rig up to ten times or more. The new 9BP4 split band speech processor uses the only system more effective than rf clipping — AND distortion is extremely low so your voice sounds natural. Under tough conditions VOMAX can help as much as most linears. Combine VOMAX with a good linear and WOW! It's simple to install and operate with any rig.

Ehrhorn Technological Operations, Inc.
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**KLM Tribander KT-34**
Covers 20m., 15m., 10m. gain: 7dB over a dipole. VSWR: better than 1.5:1. 4 elements on each band. 4 KW PEP. wind area: 6 sq. ft. boom: 16 ft. x 3", turning radius: 15 ft. max. element length: 24 ft. 2" o.d. mast or larger recommended. F/B: 20 dB or better.

349.95 Call for quote.

**HY-GAIN 5-element monoband**
20m beam antenna
5 elements up to 36\(^\circ\)" long spaced on a 34" by 2" diameter boom. bandwidth: 400 KHz. average gain: 11.6 boom length: 34 ft. longest element: 36.6 turning radius: 25 ft. wind load: 230, surface area: 9 ft. F/B: 20 dB input impedance: 50 ohms.

289.95 Call for quote.

**MOSELEY TA-33 Jr.**
Covers 10, 15, 20 meters. 3 element beam features. Gain: 10.1 dB (over isotropic source). F/B: 20 dB, 1KW PEP SSB input. VSWR at resonance: 1.5/1 or better. longest element: 26 ft. 8", wind surface: 4.0 sq. ft.

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**CUSHCRAFT A147-11**
2m antenna

36.95 Call for yours today.

**KLM 144-148 13LB**
2 meter antenna
Has 13 elements. covers 144-148 MHz. gain: 15.5 dBi VSWR less than 1.2:1. feed impedance: 200 ohm balanced. boom length: 21.5 ft. center mounting, max mast size: 2 ins. 4.1 balun for 50 ohm coax feed supplied with the antenna.

69.95 Call for yours today.

**HY-GAIN TH6DXX**
The Ultimate tribander
Gain: 8.7 dB • Front-to-back ratio 25 dB • SWR (at resonance): less than 1.5:1 • Number of elements: 6 • Frequency: 10, 15, and 20 meters • Longest element: 31.1" • Boom length: 24" • Wind load at 160 m.p.h.: 207 lbs. • Surface area: 8.09 sq. ft. Balun BN-86 recommended: 15.95

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**CUSHCRAFT ARX-2**
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Three ½ waves in phase and a ¼ wave matching stub. Extremely low angle of radiation for better signal coverage. Tunable over a broad freq range. Matched to 52 ohm coax.

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HF 4-element beam
Catch DX instead of chasing DX with the ATB-34 • Covers 10, 15, and 20 meters • High-Q coax traps rated for 2 Kw power • Direct 52 ohm feed thru 1-1 balun • Forward gain: 7.6 dB, all bands • Front-to-back ratio: 30 dB • Turn radius: 18.5' • Wind survival: 90 MPH

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**FT-101ZD**

**HIGH-PERFORMANCE HF TRANSCEIVER**

Today's technology, backed by a proud tradition, is yours to enjoy in the all-new FT-101ZD transceiver from YAESU. A host of new features are teamed with the FT-101 heritage to bring you a top-dollar value. See your dealer today for a "hands on" demonstration of the performance-packed FT-101ZD.

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**SPECIFICATIONS**

**TRANSMITTER**
- **PA Input Power:** 180 watts DC
- **Carrier Suppression:** Better than 40 dB
- **Unwanted Sideband Suppression:** Better than 40 dB @ 1000 Hz, 14 MHz
- **Spurious Radiation:** Better than 40 dB below rated output
- **Third Order Distortion Products:** Better than ~31 dB
- **Transmitter Frequency Response:** 300-2700 Hz (~-6 dB)
- **Stability:** Less than 300 Hz in first 30 minutes after 10 min. warmup; less than 100 Hz after 30 minutes over any 30 min. period
- **Negative Feedback:** 6 dB @ 14 MHz
- **Antenna Output Impedance:** 50-75 ohms, unbalanced

**RECEIVER**
- **Sensitivity:** 0.25 μV for S/N 10 dB
- **Selectivity:** 2.4 KHz at 6 dB down, 4.0 KHz at 60 dB down (1.66 shape factor); Continuously variable between 300 and 2400 Hz (~-6 dB); CW (with optional CW filter installed): 1.2 KHz at 60 dB down (2:1 shape factor)
- **Image Rejection:** Better than 60 dB (160-15 meters); Better than 50 dB (10 meters)
- **IF Rejection:** Better than 70 dB (160, 80, 20-10 m); Better than 60 dB (40 m)
- **Audio Output Impedance:** 4-16 ohms
- **Audio Output Power:** 3 watts @10% THD (into 4 ohms)

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**Price And Specifications Subject To Change Without Notice Or Obligation**

**YAESU ELECTRONICS CORP., 15954 Downey Ave., Paramount, CA 90723 • (213) 633-4007**

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Two EIMAC 3-500Zs provide the punch in Kenwood's new amplifier.

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The new heavy-duty Kenwood TL-922A linear amplifier provides 2 kW PEP input for SSB service and 1 kW input for CW, RTTY, and SSTV operation.

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Send for the EIMAC Quick Reference catalog covering the complete line of EIMAC products and for the 3-500Z Data Sheet. Learn why the important manufacturers of communication equipment choose EIMAC. Varian, EIMAC Division, 301 Industrial Way, San Carlos, California 94070. Telephone (415) 592-1221. Or contact any of the more than 30 Varian Electron Device Group Sales Offices throughout the world.

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