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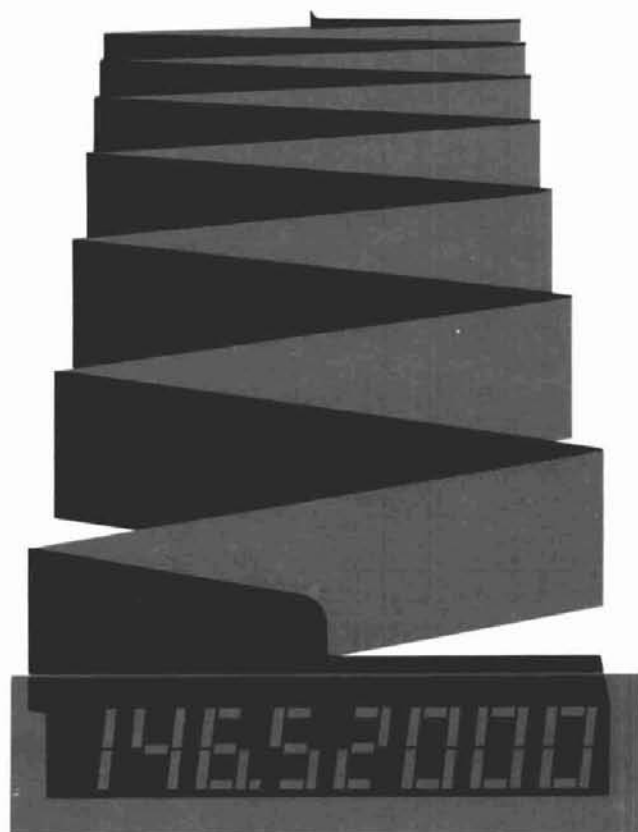
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FEBRUARY 1978

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- and much more . . .

SPECIAL ISSUE:



**frequency
counters**

TEMPO

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10W	130W	130A10	\$179
30W	130W	130A30	\$189
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10W	80W	80A10	\$149
30W	80W	80A30	\$159

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10W	40W	40D10	\$145
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FCC Type accepted models available.

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Henry Radio

Why you should buy a digital multimeter from the leader in digital multimeters.

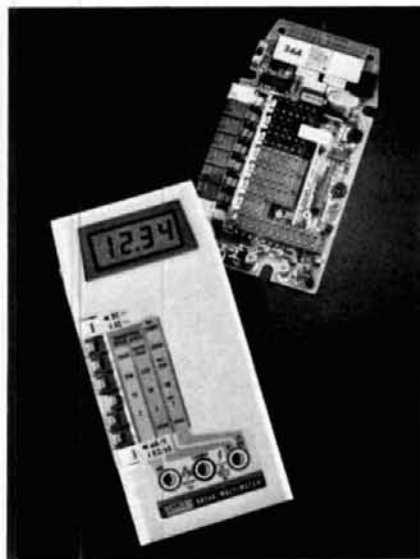
If you're shopping for your first multimeter, or moving up to digital from analog, there are a few things you should know.

First, look at more than price. You'll find, for instance, that the new Fluke 8020A DMM offers features you won't find on other DMMs at any price. And it's only \$169.*

Second, quality pays. Fluke is recognized as the leading maker of multimeters (among other things) with a 30-year heritage of quality, excellence and value that pays off for you in the 8020A.

Third, don't under-buy. You may think that a precision 3½-digit digital multimeter is too much instrument for you right now. But considering our rapidly changing technology, you're going to need digital yesterday.

If you're just beginning, go digital.



Why not analog? Because the 8020A has 0.25% dc accuracy, and that's ten

times better than most analog meters.

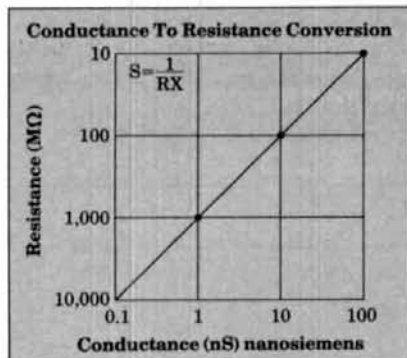
Also, the 8020A's digital performance means things like 26 ranges and seven functions. And the tougher your home projects get, the more you need the 8020A's full-range versatility and accuracy. The 8020A has it; analog meters don't.

If you're a pro.

You already know Fluke. And you probably own a benchtop-model multimeter.

Now consider the 8020A: smaller in size, but just as big in capability. Like 2000-count resolution and high-low power ohms. Autozero and autopolarity. And the 8020A has 3-way protection against overvoltage, overcurrent and transients to 6000V!

Nanosiemens?



Beginner or pro, you'll find the meter you now have can't measure nanosiemens. So what? With the 8020A conductance function, you can measure the equivalent of 10,000 megohms in nanosiemens. Like capacitor, circuit board and insulation leakage. And, you can check transistor gain with a simple, homemade adapter. Only with the 8020A, a 13-oz. heavyweight that goes where you go, with confidence.

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Of course, you can pay more. Or less. In fact, you could pay almost as much for equally compact but more simplistic meters, and get far less versatility. And, the 8020A gives you the 'plus' of custom CMOS LSI chip design, and a minimum number of parts (47 in all). All parts and service available at more than 100 Fluke service centers, worldwide. Guaranteed, for a full year.

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*U.S. price only

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FLUKE[®]

This NEW MFJ Versa Tuner II . . .

has SWR and dual range wattmeter, antenna switch, efficient airwound inductor, built in balun. Up to 300 watts RF output. Matches everything from 160 thru 10 Meters: dipoles, inverted vees, random wires, verticals, mobile whips, beams, balance lines, coax lines.



BRAND NEW

\$79⁹⁵

Antenna matching capacitor. 208 pf. 1000 volt spacing.

Sets power range, 300 and 30 watts. Pull for SWR.

Meter reads SWR and RF watts in 2 ranges.

Efficient airwound inductor gives more watts out and less losses.

Transmitter matching capacitor. 208 pf. 1000 volt spacing.

Only MFJ gives you this MFJ-941 Versa Tuner II with all these features at this price:

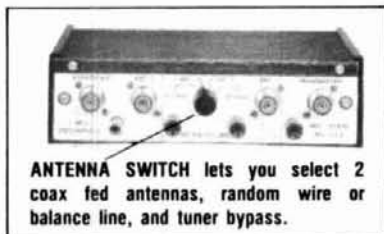
A SWR and dual range wattmeter (300 and 30 watts full scale) lets you measure RF power output for simplified tuning.

An antenna switch lets you select 2 coax fed antennas, random wire or balance line, and tuner bypass.

A new efficient airwound inductor (12 positions) gives you less losses than a tapped toroid for more watts out.

A 1:4 balun for balance lines. 1000 volt capacitor spacing. Mounting brackets for mobile installations (not shown).

With the NEW MFJ Versa Tuner II you can run your full transceiver power output — up to 300 watts RF power output — and match your



ANTENNA SWITCH lets you select 2 coax fed antennas, random wire or balance line, and tuner bypass.

transmitter to any feedline from 160 thru 10 Meters whether you have coax cable, balance line, or random wire.

You can tune out the SWR on your dipole, inverted vee, random wire, vertical, mobile whip, beam, quad, or whatever you have.

You can even operate all bands with just

one existing antenna. No need to put up separate antennas for each band.

Increase the usable bandwidth of your mobile whip by tuning out the SWR from inside your car. Works great with all solid state rigs (like the Atlas) and with all tube type rigs.

It travels well, too. Its ultra compact size 5x2x6 inches fits easily in a small corner of your suitcase.

This beautiful little tuner is housed in a deluxe eggshell white Ten-Tec enclosure with walnut grain sides.

S0-239 coax connectors are provided for transmitter input and coax fed antennas. Quality five way binding posts are used for the balance line inputs (2), random wire input (1), and ground (1).



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BRAND NEW

MFJ-901 VERSA TUNER

New efficient air wound coil for more watts out.

Only MFJ uses an efficient air wound inductor (12 positions) in this class of tuners to give you more watts out and less losses than a tapped toroid. Matches everything from 160 thru 10 Meters: dipoles, inverted vees, random wires, verticals, mobile whips, beams, balance lines, coax lines. Up to 200 watts RF output. 1:4 balun for balance lines. Tune out the SWR of your mobile whip from inside your car. Works with all rigs. Ultra compact 5x2x6 inches. S0-239 connectors; 5 way binding posts. Ten Tec enclosure.



\$49⁹⁵

BRAND NEW

MFJ-900 ECONO TUNER

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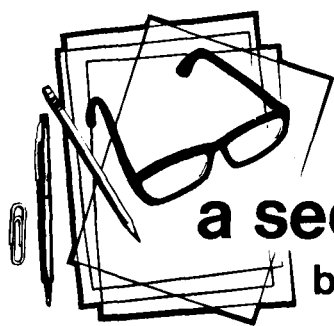
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a second look

by Jim Fisk

As this issue goes to press, it appears that the launch of the next amateur communications satellite, AMSAT-OSCAR D, is imminent (after it is in orbit this satellite will be known as OSCAR 8). Like its famous predecessors, OSCARs 5, 6, and 7, this new "bird" has two transponders: a two-to-ten meter unit similar to that used in OSCAR 7 called *Mode A*, and a two-meter, 70-centimeter transponder designated *Mode J*. The Mode J transponder was built by members of the Japan AMSAT Association in Tokyo; a similar combination of input/output frequencies was used in the short-lived OSCAR IV spacecraft back in 1966.

The new OSCAR will be launched from NASA's Western Test Range in California as a "piggyback" payload aboard the second stage of the two-stage Thor-Delta launch vehicle which will carry NASA's Landsat-C earth resources technology satellite into orbit. Because of scheduling at the Western Test Range, and the complex pre-launch checkout of Landsat-C, it's impossible to pinpoint the exact launch date, but late reports from AMSAT indicate it will be sometime in early March.

The new spacecraft is a 38 cm rectangular solid 33 cm high, weighs 27 kg, and is solar powered. The solar cells, combined with the 12-cell rechargeable nickel-cadmium battery, should be adequate to power the satellite in Mode A for several years. The receiving antenna for both modes is a turnstile comprised of four 48 cm lengths of 12 mm carpenter's rule. Four permanent magnets located inside the spacecraft provide stabilization; this is the same technique used in OSCARs 6 and 7. The polarity of the magnets is such that the top of the satellite always points toward the earth's magnetic north pole. Permalloy damping rods mounted behind the solar panels are designed to reduce the spin of the spacecraft; their operation is similar to a shorted transformer turn as it cuts the lines of flux of the earth's magnetic field. OSCAR 7 used the same system with good success.

The spacecraft will be automatically powered up upon ejection from the Thor-Delta launch vehicle over northern Greenland. It is designed to come on in Mode J; the Mode A transponder will not be turned on until the satellite is almost completely stabilized in orbit, which may take as long as a week. This is because the 10-meter dipole antenna cannot be deployed until the spacecraft's spin rate is less than 1 revolution per minute; otherwise the antenna may be severely damaged. The deployment process takes about 15 seconds and cannot be reversed — the elements can't be retracted once they are extended — so correct deployment is crucial.

OSCAR 8's orbit is planned to be sun-synchronous, with passes repeating at approximately the same time each day on a one-day cycle (as opposed to the two-day cycle of OSCARs 6 and 7). Since the altitude of OSCAR 8's orbit, at 900 km, is just over half the altitude of OSCARs 6 and 7, the maximum communications range will be slightly shorter. The usable time on an overhead pass will be about 18 minutes instead of the 22 minutes provided by OSCAR 7, and the horizon range will be 3220 km (down slightly from the 3940 km horizon range of OSCAR 7). In practical terms this means that transatlantic communications will still be possible with OSCAR 8, but not as often as with OSCAR 7.

One of the big advantages of the 900 km sun-synchronous orbit is that keeping track of OSCAR 8 is going to be much simpler than it was for earlier amateur satellites; it will come into range at nearly the same time every day — the overhead descending node pass is planned for 9:30 AM local time. The satellite's anticipated useful operating lifetime is three years.

Since the prime mission of the OSCAR 8 spacecraft is to use the Mode A transponder for the ARRL OSCAR educational program in schools, the spacecraft may be left in Mode A during weekdays and put into Mode J on weekends. Because of the relatively high current drain of the Mode J transponder, however, the power budget may not support the Mode J transponder for continuous full-time operation over an entire weekend. The spacecraft may also be switched to Mode J during the evening hours in the Western Hemisphere, depending on the burden to the command stations and the condition of the on-board batteries.

The Mode A transponder on the new spacecraft has the same frequency passband as OSCAR 7 (input between 145.85 and 145.95 MHz, output between 29.40 and 29.50 MHz). Approximately -95 dBm is required at the transponder input terminals for an output of one watt; this corresponds to an effective radiated power from the ground of about 80 watts. The 250 mW telemetry beacon operates at 29.402 MHz.

The Mode J transponder operates with an input frequency passband between 145.90 and 146.00 MHz — the output is between 435.10 and 435.20 MHz. Power output is 1 to 2 watts PEP, and the output is inverted (upper-sideband uplink signals become lower-sideband downlink signals). Uplink sensitivity for 1 watt output is -105 dBm which corresponds to an effective radiated power from the ground of about 8 watts (note the greatly improved sensitivity of this mode, and keep your power down). A 100 mW beacon at 435.095 will carry telemetry information.

Jim Fisk, W1HR
editor-in-chief



IC-245/SSB, the Mobile Maximizer

MULTI-MODE MOBILE TRANSCEIVER

ICOM's new, fully synthesized **IC-245/SSB** maximizes mobile FM, SSB and CW operations with a very compact dash mounted transceiver like nothing else in the mobile world. This Maximizer's single knob dial makes the most of the mobile modes while totally minimizing manipulations. One fast moving detent knob gives the **IC-245/SSB** accurate tuning in all modes with positively no time lag or backlash in display stability, even when flying through steps of 100 Hz at 5 KHz per second. **And just as easily, you can work the new 144.5-145.5 repeaters regardless of splits or spacing.**

- **Single knob frequency selection:** The **IC-245/SSB** is synthesized with convenient single knob frequency selection over the entire band. No more fussing with two or more knobs just to check what is going on around the band. One easy spin of the 50-position detent knob does it all.
- **Two VFO's built-in:** The second VFO, which is a more money tack-on with most other transceivers, is a standard item with every **IC-245/SSB**.
- **Variable offset:** Any offset from 10 KHz through 4 MHz, in multiples of 10 KHz, can be programmed with the LSI synthesizer.
- **Remote programming:** The **IC-245/SSB** LSI chip provides for the input of programming digits from a remote key pad, which can be combined with Touch Tone circuitry to provide simultaneous remote program and tone. Computer control from a PIA interface is also possible.
- **FM stability on SSB and CW:** The **IC-245/SSB** synthesis of 100 Hz steps makes mobile SSB as stable as FM. This extended range of operation is attracting many FM'ers who have been operating on the direct channels and have discovered SSB.

The **IC-245/SSB** is the very best and most versatile mobile transceiver made: that's all. For more information and your own hands-on demonstration, see your ICOM dealer. When you mount your **IC-245/SSB**, you'll have the very maximum in multi-mode mobile.

Maximize the new repeater band: both the **IC-245/SSB** and the **IC-211** operate the new FCC repeater spectrum with no modification. They always have.

All ICOM radios significantly exceed FCC specifications limiting spurious emissions.

Specifications: Frequency Coverage: 144.00 to 148.00 MHz Modes: FM (F3), SSB (A3J), CW (A1) Supply Voltage: DC 13.8V ± 15% Size: 90mm (h) x 155mm (w) x 235mm (d) Weight: 6.8 Kg TX Output: F3, 10W; A3J, 10W (PEP); A1, 10W Spurious Radiation: -60 dB below Carrier Microphone Impedance: 600 Ohms Sensitivity: A3J & A1, 0.5 microvolt input gives 10 dB S+N/N; F3, 0.6 microvolt or less for 20 dB quieting S+N+D/N, at 1 microvolt input, 30 dB Spurious Response: -60 dB or better Synthesizer Frequency Range: 144.00 MHz to 148.00 MHz Synthesizer Step Size: 5 KHz for FM, 100 Hz or 5KHz for SSB

VHF/UHF AMATEUR AND MARINE COMMUNICATION EQUIPMENT

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presstop

ANOTHER WARC ADVISORY Committee Meeting on Amateur Radio is beginning to look likely for February, and if so it'll be an important one. The Sixth (or possibly Seventh) Notice of Inquiry on the 1979 World Administrative Radio Conference is scheduled for release in January, and it's expected that a further revised frequency table will be included. Since our previous efforts at generating strong U.S. support for both expansion of present HF bands and the creation of new ones in the LF and HF spectrums were only marginally successful (Presstop, July, 1977) — and that success, expansion of 40, 20, and 15 meters, seems threatened by the Department of Defense's expressed frequency needs — a new and determined offensive is badly needed.

Some Strong Ammunition in support of our position has been developed overseas and should help greatly. The continued increase in the U.S. Amateur population, along with the current slackening of CB enthusiasm, should also help out.

K5NY RECEIVED AN 18-MONTH JAIL sentence and was fined \$500 by U.S. District Court Judge Edward J. Boyle, Jr. last November as a result of his pleading guilty to three counts of transmitting obscene language and interfering with a New Orleans repeater this past summer. K5NY must serve 90 days of his sentence after which the remainder will be suspended; his co-defendant — WB5AWN — received a suspended sentence but must pay costs for the public defender. K5NY received a severe reprimand from the Judge, himself a CBER.

POINT-OF-SALE CONTROL for linear amplifiers has been instituted by Canada's Department of Communications. In a Canada Gazette announcement, the DOC stated that all linear buyers must sign a special form including their names, and addresses at the time of purchase. The form is then forwarded to the DOC and the buyer's name compared with lists of General Radio Service (CB) licensees to determine whether the purchaser is in violation of DOC rules barring linear possession by an operator in the General Radio Service.

No-Code Canadian Amateur License is expected by next fall as a result of the November 26 Department of Communications/Canadian Amateur Radio Federation National Amateur Radio Symposium in Ottawa. More than 100 Amateurs from across Canada, representing 27 Canadian Amateur organizations, attended the jointly sponsored session, during which the suggestion for a Canadian Novice-class license was firmly turned down while a no-code "experimenter's" license for 200 MHz and up was strongly supported. Expectations are that the new license will have a tough technical exam; other details are still to be worked out.

A New Permanent Canadian Prefix may be forthcoming as Ontario Amateurs have just about used all the VE3x3 call signs. The DOC, with the assistance of the Radio Society of Ontario, will be looking into alternatives.

1978 Could See VC7 in use by Vancouver Island Amateurs. The Victoria Short Wave Club has asked the DOC to authorize the special prefix to mark the bicentennial of Captain Cook's exploration of Vancouver Island.

A Canadian Assault on 420-430 MHz is in the wind. A recent Department of Communications study of 406-960 MHz is expected to propose 420-430 MHz for mobile services, with 430-450 MHz shared between radiolocation and Amateurs as at present. On the positive side, the same study also will propose a new 902-928 MHz Amateur band to be shared with fixed services and radiolocation.

OSCAR 7'S THIRD BIRTHDAY was Tuesday, November 15, when the satellite reached its three year design lifetime. If OSCAR 6's performance can be extrapolated, however, expect OSCAR 7 to be serving us for a long time to come. Mode jumping problems are continuing, with the combination of low battery voltage (from seasonal sun angle on the solar panels) and excessive user signals (especially from Europe) the probable culprits.

Oscar 7's Mode Jumping problem isn't entirely from user abuse, W9VI (ex-W9OII) found when he observed a jump on an early morning pass with no user signals audible. A high noise output, often observed before jumps, has been linked with the problem. One theory is that it's an internal malfunction, while another says it's related to an ionization buildup as it does seem seasonal.

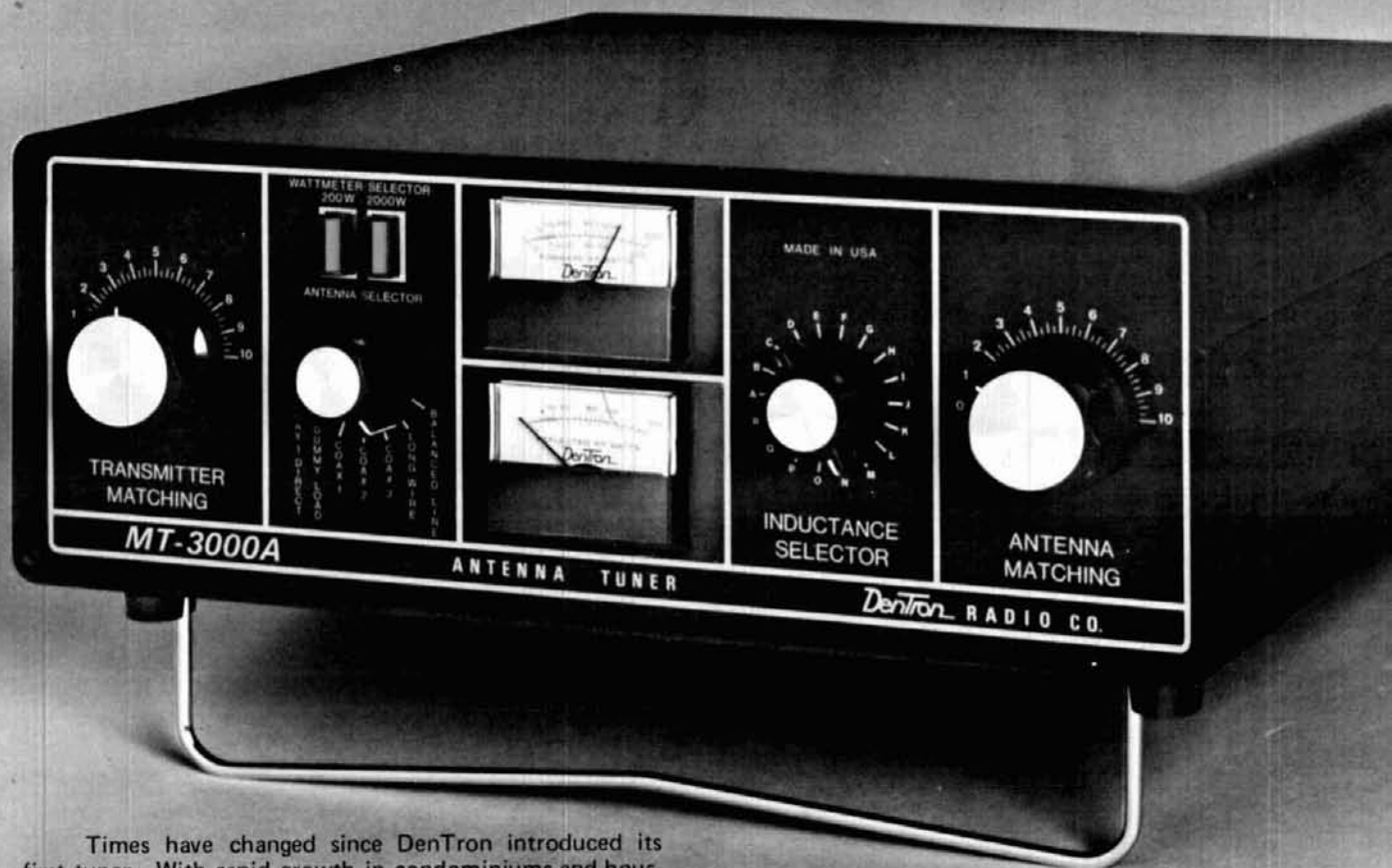
FCC EXTENDED its working day to 5:30 PM effective January 3. The new hours (for Washington, D.C. offices only) are from 8:00 AM to 5:30 PM, with employee schedules rearranged to insure adequate staffing of Commission offices throughout the day.

REPEATER LICENSE APPLICATIONS are still arriving at the FCC despite the "deregulation" Report and Order and its subsequent stay. Until a final decision on repeater licensing is made, however, repeater applications are being returned without action.

License Fees are starting to turn up again in some Amateur applications, even though fees were suspended almost a year ago. The Personal Radio Division reports about 20% of applications received now include fees, with most of them from CBERs.

MICROWAVE EXPERIMENTERS should contact ZS1HS of the SARL, who is compiling information on Amateur microwave work for use in preparations for the World Administrative Radio Conference in 1979.

Look closely at the new MT-3000A. You've never seen anything like it.



Times have changed since DenTron introduced its first tuner. With rapid growth in condominiums and housing developments, we have new problems that require new solutions.

DenTron decided to rethink the tuner and what its total capabilities should be.

The MT-3000A is a capsulized solution to many problems. It incorporates 4 unique features to give you the most versatile antenna tuner ever built.

First, as a rugged antenna tuner the MT-3000A easily handles a full 3KW pep. It is continuous tuning 1.8-30mc. It matches everything between 160 and 10 meters.

Second, the MT-3000A has built-in dual watt meters.

Third, it has a built-in 50 ohm dummy load for proper exciter adjustment.

Fourth, the antenna selector switch; (a) enables you to by-pass the tuner direct; (b) select the dummy load or 5 other antenna systems, including random wire or balanced feed.

The compact size alone of the MT-3000A (5½" a 14" x 14") makes it revolutionary. Combine that with its four built-in accessories and we're sure you'll agree that the MT-3000A is one of the most innovative and exciting instruments offered for amateur use.

At **\$349.50** the MT-3000A is not inexpensive. But it is less than you'd expect to pay for each of these accessories separately.

As unique as this tuner is, there are many things it shares with all DenTron products. It is built with the same meticulous attention to detail and American craftsmanship that is synonymous with DenTron.

After seeing the outstanding MT-3000A, wouldn't you rather have your problems solved by DenTron?

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The Kenwood family is growing!
The TL-922, a brand new linear amplifier, is now a reality. Give yourself the "big signal" that commands attention on today's crowded bands. The TL-922 runs the full legal limit on all the ham bands from 160-10 meters and is compatible with most amateur exciters. The TL-922 is a must in any Kenwood station. Make yourself heard like you've never been heard before, with the Kenwood TL-922 linear amplifier.

TL-922

What makes one linear amplifier different than all the rest? Check out these features:

- Full amateur band coverage** — Includes 160 meters.
- Instant heating filaments** — The 3-500Z tubes require no warm up period. Just turn it on and go!
- Time delay fan circuit** — Even after you turn the TL-922 off, the super quiet fan continues to work for approximately 2 minutes to greatly extend tube life.
- Adjustable ALC output voltage** — Lets you tailor the ALC voltage to your exciter.
- Standby position** — Provides amplifier bypassing without having to turn the AC power off.
- Two independent safety interlocks** — One disconnects AC line voltage and the second shorts B+ to ground when tripped.
- Vernier plate control** — For smooth, easy tune-up.
- Diecast side panels** — Includes functional carrying handles for easy transportation.
- Thermal protection of power transformer** — Amplifier automatically switches to standby if power transformer temperature exceeds 145°F.
- Tuned Input Circuit** — Means improved spurious characteristics.
- Line voltage selector** — Easily switched between 120 and 240 VAC.
- Plate Current Meter** — Separate meter allows continuous monitoring of plate current.



Shown with top panels removed

Frequency Range: Amateur bands, 160-10 meters
Drive Power Required: 80 W nom, 120 W max
Mode and Duty Cycle: SSB, cont for 30 min CW and RTTY, key-down cont for 10 min
RF Input Power: SSB: 2,000 watts PEP, CW, RTTY: 1,000 watts DC
Plate Voltage: (at idle) 3.1 KV SSB, 2.2 KV CW, RTTY
Circuit Type: Class AB₂ grounded grid linear amplifier
Input Impedance: 50 Ω, unbalanced at better than 1.5 SWR
Output Impedance: 50 to 75 Ω, unbalanced
Harmonic Suppression: min 40 db, depending on exciter used
Fan Motor Delay Time: 140±30 seconds, (at room temperature)

ALC: Neg going, adjustable threshold, -8V DC max output (typ).
Tubes: 2 x Eimac 3-500Z
Semiconductors: 18 Diodes, 1 Zener diode
Power Requirements: 120V, 28A; 220/240V, 14A; 50/60Hz, for maximum SSB input
Dimensions: 390 mm (15½") x 190 mm (7½") x 407 mm (16")
Weight: Net 31 kg (68 lbs) Shipping 38kg (83 lbs)

The above specifications are subject to change without notice due to developments in technology.

...pacesetter in amateur radio

Kenwood offers this handsome pair for the amateur who appreciates the advantages of operating a receiver/transmitter combination. Discover the difference in performance, features and price of the 599D "Twins".

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

The R-599D is the most complete receiver ever offered. It is entirely solid state, superbly reliable and compact. It covers the full amateur band, 10 through 160 meters, CW, LSB, USB, AM and FM. The T-599D is solid-state with the exception of only three tubes, has built-in power supply and full metering. It operates CW, LSB, USB and AM and, of course, is a perfect match to the R-599D receiver. If you have never considered the advantages of operating a receiver/transmitter combination, maybe you should.

Because of the larger number of controls and dual VFOs the combination offers flexibility impossible to duplicate with a transceiver.

Compare the specs and prices of the R/T-599D combination with any other brand of separates. Remember, the R-599D is all solid state (and includes four filters).

Your choice will obviously be the Kenwood.

R-599D/T-599D



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The Kenwood family is growing! The TL-922, a brand new linear amplifier, is now a reality.

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TRIO-KENWOOD COMMUNICATIONS INC.
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understanding and using electronic counters

Only a few short years ago, it was extremely rare to see an electronic counter outside of a laboratory or a specialized service installation. Today counters can be found in ham shacks all over the world. Of course the reason for this proliferation is obvious — the integrated circuit, and in particular, medium- and large-scale integration. In the early sixties, a typical 10-MHz counter weighed nearly 120 pounds (55kg), occupied about 5.75 cubic feet (165,000 cubic centimeters), and dissipated approximately 600 watts as heat. By way of contrast, a 520-MHz counter currently produced by the same manufacturer weighs 4.75 pounds (2.16kg), has a volume of approximately 213 cubic inches (3890 cubic centimeters), and dissipates less than 20 watts.

As size has decreased, so has cost. That antediluvian counter cost \$2600 in "1966 dollars;" today you can buy a counter for under \$100 if you want a bare-bones instrument, and can get a 250-MHz multifunction counter for less than \$400. Because of today's relatively low costs, counters have become a versatile tool in the ham station and on the work bench. If you have one, this article may help you make better use of it. If you are planning to buy one, it may help you to decide what to look for.

You may have noticed that the title of the article uses the term *electronic*, rather than *frequency*, counter. This was not a pedantic choice; electronic describes the type of counter and is inclusive of all functions that the counter may perform, only one of which may be the measurement of frequency. We shall discuss these various functions, although emphasis will be placed on frequency measurement, which is of primary interest to the average ham.

Before discussing the applications and limitations of the frequency counter, it is important to cover the method by which frequency is measured by the counter. Regardless of the type and complexity of the instrument, all counters measure frequency by comparing the frequency of the input signal with a known frequency or time period. Fig. 1 shows the basic functional blocks of a typical counter. The main function of the signal conditioner is to con-

vert the input signal to one whose amplitude and waveshape are compatible with the internal circuitry or logic of the counter. It generally includes an amplifier to increase the amplitude of the incoming signal, and may also contain an attenuator for input signals of high amplitude, trigger level and slope selection circuits, and so on. No matter how the signal is processed, the output of the signal conditioner is a pulse train in which each pulse corresponds to one cycle or event of the input signal.

The conditioned signal is applied to a gating circuit, which is shown symbolically as a single logic gate, but which is actually a more complex circuit. The gate is opened for a predetermined, accurate time interval, during which the signal passes through to the decade counters. These counters count the number of pulses which are gated through, and transfer the count to the display. The number of decade counters determines the number of digits which are displayed, one counter being required for each digit. The display can utilize any type of visual readout device, such as gas-discharge numeric tubes, light-emitting diode arrays, or liquid-crystal displays.

Since the decade counters count the number of pulses which pass through the gate, it follows that the accuracy of the instrument is a function of the time that the gate is open. This interval is, in turn, a function of the time-base accuracy. The time-base oscillator in the modern counter is invariably a crystal-controlled oscillator operating at a frequency between 1 and 10 MHz, although there have been counters made in the past which used crystal frequencies as low as 100 kHz, or even used the ac line frequency as a time base. Even though the oscillator frequency must be divided, crystals in the 1- to 10-MHz range are used because they are inherently more stable than those which work at lower frequencies; the optimum range for stability is between 4 and 10 MHz for most types of crystals.

The divided time-base frequency drives the gate-

By Robert S. Stein, W6NBI, 1849 Middleton Avenue, Los Altos, California 94022.

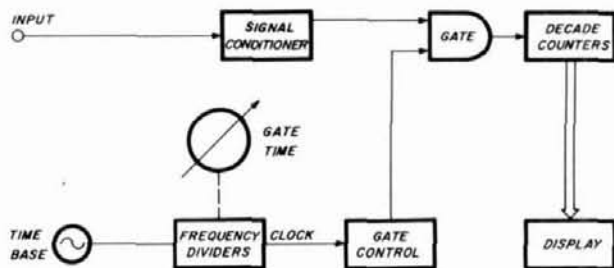


fig. 1. Functional block diagram of the basic frequency counter.

control circuit, which controls the gate-time interval in accordance with the divided time-base frequency. To explain the need for dividing the time-base frequency, we must at this time discuss resolution, or the smallest frequency increment which the counter displays.

Let us assume that the frequency of the time-base oscillator is 10 MHz, and that the gate control holds the gate open for exactly 10 clock pulses (clock being the term used to designate the time-base signal or the signal derived, through the dividers, from the time base). Since each cycle from the 10-MHz oscillator has a period of 0.1 microsecond, the gate will be open for 1 microsecond. If a 1-MHz input signal were being measured, only one pulse would be gated through, and the counter would display a 1. Frequencies below 1 MHz might or might not produce a reading at all, and those above 1 MHz could be read only to within one digit of the nearest megahertz (more about this later). Thus, the resolution would be 1 MHz at best, an obviously unsatisfactory arrangement.

Suppose, instead, that the time-base frequency were divided down to 10 Hz, or a period of 0.1 second. The gate will now be open for 1 second, and 1 million pulses from a 1-MHz input will be counted. Now the counter will display 1000000, which provides us with a resolution of 1 Hz. Thus, the resolution is the reciprocal of the gate time, and in fact, some counters with selectable gate times have the switch positions designated by the gate time.

The limiting factors governing resolution are the number of digits in the display and the tolerable gate time. Usually 0.1 Hz is the smallest resolution practical, in that it involves a 10-second gate time and a 9-digit display up to 99.999999 MHz. The gate time can be reduced in a computing counter, but that is outside the scope of this discussion.

Of course, it is not always necessary to read frequency to a tenth of a hertz, nor is it particularly convenient to have to wait for a 10-second count. By selecting the appropriate output from the frequency-divider chain, you can reduce the gate

time and resolution to values which may be more appropriate to the measurement. The normal range of gate times is typically between 1 millisecond and 10 seconds, corresponding to resolutions of 1 kHz to 0.1 Hz.

The number of digits in the display can be reduced by switching both the displays and gate times, a technique which is used in many low-priced counters having a 5-digit display. A 2-position switch is used to select gate times of 1 second (1-Hz resolution) and 1 millisecond (1-kHz resolution). When the gate time is 1 second, the five decade counters can produce a display up to 99,999 Hz; when the gate time is 1 millisecond, up to 99,999 kHz or the frequency limit of the counter can be displayed. Thus by switching the clock, the equivalent of eight digits is obtained, with overlap between the two readings. This is an economical, but oftentimes inconvenient, way of obtaining improved resolution.

It should be reiterated at this point that a counter displays a pulse count. Whether the display reads out in Hz, kHz, or MHz is simply one of convenience and the location of the display decimal point. The decimal point is either fixed, or is switch selected with the gate time, and its position is independent of the actual count.

time-base accuracy

It should be apparent from the preceding discussion that the time-base oscillator is the most critical part of the counter, in that it determines the overall accuracy of the instrument. Let us examine its effect, in terms of the specifications usually given for a counter.



The Yaesu YC-500 Frequency Counter has a frequency range of 10 Hz to over 500 MHz. Its six-digit display provides the equivalent of eight digits when the gate time is switched. A room-temperature crystal, TCXO, or ovenized crystal time base may be ordered (photo courtesy Yaesu Electronic Corporation).

First of all, the accuracy of the time base, either in per cent or parts per total, is directly translatable to the measurement of frequency, period, interval, or any other function which the counter may be capable of measuring and which utilizes the time base. This holds true, regardless of the magnitude of measurement. For example, if a 1-MHz time-base oscillator is off frequency by 2 Hz, that represents an error of 0.0002 percent. The gate interval, therefore, will also be in error by the same percentage, and the displayed count will have the same error. If the frequency being displayed is 50.000000 MHz, the error will be 100 Hz. If the time-base frequency is high, the displayed count will be low, since the higher the frequency, the shorter the gate time. If the time-base frequency is low, the opposite will hold true.

Time-base accuracy specifications should include the parameters listed in **table 1**, although most lower-priced instruments may omit one or more. Typical values for the various types of oscillators are included as examples. It can be seen that temperature change has the greatest effect on frequency. In the examples listed, the specification for temperature stability can be improved by one order of magnitude by using a TCXO (temperature-compensated crystal oscillator) instead of a room-temperature crystal. In the real world, however, a good room-temperature crystal may be better than a poor TCXO; you must compare the specifications.

The oscillator aging rate is not as important, since this will manifest itself as a gradual change in frequency, and is predicated on the oscillator running continuously. If the counter is designed so that the oscillator circuit is powered as long as the instrument is connected to the primary power source, the specified aging rate is valid. If the



The Fluke 1910-A Multi-Counter is one of a series which provides frequency, period, period-average, ratio, and totalize functions. The 1910A is rated to 125 MHz; the 1911A and 1912A are similar in appearance and will measure frequencies to 250 and 520 MHz, respectively (photo courtesy John Fluke Manufacturing Company).

table 1. typical specifications for time-base oscillators.

	room temperature crystal	TCXO*	oven oscillator
Aging rate (long term stability)	5×10^{-7} /mo	3×10^{-7} /mo	5×10^{-10} †
Temperature, 0-50°C	5×10^{-6}	5×10^{-7}	7×10^{-9}
Line voltage, $\pm 10\%$	1×10^{-7}	5×10^{-8}	5×10^{-9}
Short-term stability per day			1×10^{-10}

* Temperature-compensated crystal oscillator

† After 24-hour warm-up

‡ rms/sec

oscillator is deenergized along with the rest of the counter when the instrument is turned off, however, the aging specification means little or nothing.



fig. 2. Gate time and signal pulse train, showing ± 1 count ambiguity.

Short-term stability is generally specified only for very stable, ovenized oscillators and is pertinent only to laboratory-type measurements.

Time-base errors can be corrected by recalibrating the oscillator against a known standard or against WWV. Virtually all counters incorporate an adjustment control for this purpose. The techniques used in recalibrating the oscillator will be covered later in this article.

frequency-measurement accuracy

Although the preceding discussion of time-base

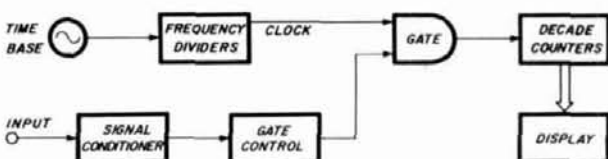


fig. 3. Functional block diagram of an electronic counter configured for period measurement.

accuracy would appear to account for any inaccuracies in the measurement of frequency, such is not the case. The specification for the frequency-measurement accuracy of all electronic counters is invariably stated as \pm time-base accuracy ± 1 digit. The last term of that statement is known as the 1-count ambiguity — but what does this mean?

Fig. 2 shows the signal under measurement and its relationship to the gate. Although the successive gate times, t_1 and t_2 , are equal in duration, the gate is not synchronized with the signal.

Therefore it is possible, during gate time t_1 , for five signal pulses (numbered 2 through 6) to be gated, while during time t_2 , six signal pulses (numbered 1' through 6') may be gated. Thus there is always an irreducible ± 1 -count ambiguity in the least significant digit of the display.

The per cent of error due to the 1-count ambiguity is reduced as the measured frequency increases, since it becomes increasingly less significant compared to the total count.

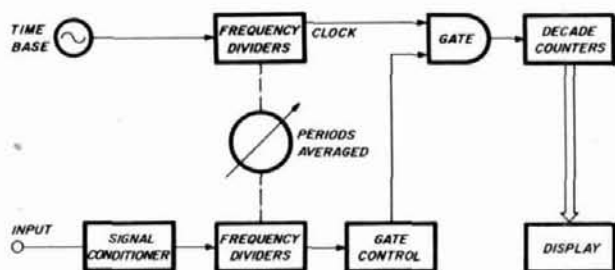


fig. 4. Functional block diagram of a period-averaging counter.

The maximum error is an inverse function of the frequency being measured and the number of pulses being counted, *i.e.* the gate time, and is expressed as

$$\% \text{ Error} = \pm \frac{100}{f \cdot t}$$

where f is the frequency in Hz, and t is the gate time in seconds.

From the above equation, it can be seen that measuring a 10-MHz signal with a 1-second gate time will be subject to a ± 0.001 per cent error. However, measuring a 20-Hz signal with the same gate time may result in a counter display between 19 and 21 Hz, a ± 5 per cent error. This would not be very satisfactory if you were attempting to calibrate the low-frequency end of an audio oscillator, and must be taken into account.

period and period-averaging measurements

One of the ways in which accurate low-frequency measurements may be made is to measure the period of the signal, rather than the frequency. Since the period of a signal is the reciprocal of its frequency, the frequency can be calculated accordingly. It might also be expected that a simple reciprocal arrangement of the functional blocks of an electronic counter would provide a measurement of period, which turns out to be true.

In fig. 3, the time-base and signal inputs have been interchanged. If the gate-control circuit is configured so that the gate is open for one period

of the input signal, and the 1-MHz clock is applied to the gate input, a series of pulses having a period of 1 microsecond will be gated through to the decade counters. Therefore the counter will indicate the period of the input signal in microseconds. If the clock frequency were reduced to 1 kHz, the counter would display the signal period in milliseconds.

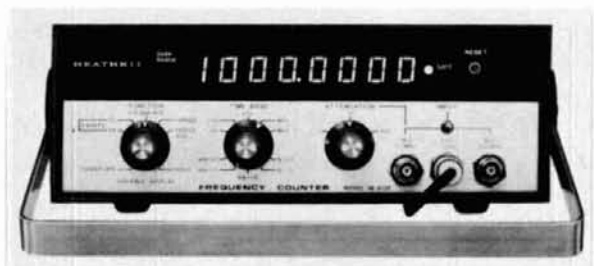
Let us now reconsider the frequency measurement of the aforementioned 20-Hz signal to see how we can improve the possible ± 5 per cent error which can occur with a 1-second gate time. The 20-Hz signal period is 0.05 second, so that if the counter were configured to measure period, it would display 50,000 microseconds. Thus, the number of significant digits in the display has been increased from two to five, and the gate time reduced from 1.0 to 0.05 second. Now if you were calibrating the audio oscillator and wanted a 1 per cent dial accuracy, you could accept any reading between 49,500 and 50,500 microseconds, subject to the correction for period-measurement accuracy.

Period measurements are inherently less accurate than frequency measurements because it is the signal, rather than the time base, which controls the gate time. Noise on the input signal, regardless of the measurement mode, causes an uncertainty in the point at which the trigger circuit in the signal conditioner switches. (It is the trigger circuit which converts the input signal to a waveform which is compatible with the counter's circuitry.) If the noise is not great enough to cause false triggering which would result in more or less output pulses than correspond to the input, no significant error is introduced in a frequency measurement.

For period measurements, however, this uncertainty results in an error in the gate time, since the in-



Ballantine's model 5720A Frequency Counter covers the range from 10 Hz to more than 80 MHz and provides frequency and ratio measurements. This counter also includes an audio multiplier circuit for input frequencies from 50 Hz to 1 kHz which provides resolution of 0.01 Hz with only 1-second measurement time (photo courtesy Ballantine Laboratories).



Heath's model IM-4130 is capable of measuring period, period average, events (totalizing), and frequency over a 5-Hz to 1-GHz range. Since it has provisions for connecting an external time base, ratio measurements can also be made, as explained in the text (photo courtesy Heath Company).

put signal controls the gate time. This error is known as trigger error, and is part of the instrument specification for period measurement, usually expressed as \pm time-base error \pm trigger error \pm 1 count. Notice that the trigger error has been added to the previously discussed expression for frequency-measurement error. For low-frequency noise on a sine-wave input, the approximate worst-case errors are \pm 3 per cent for a 20-dB signal-to-noise ratio,

\pm 0.3 per cent for a 40 dB signal-to-noise ratio, and \pm 0.03 per cent for a 60-dB signal-to-noise ratio.

In addition to the trigger error caused by noise, the stability of the input signal may be such that successive gate times are of differing durations. Even though the differences may be minute, they will manifest themselves as a continuously changing display on the counter, especially at high resolutions. This is not to be considered a counter error, since it does not occur with a stable input signal.

Period errors may be minimized by averaging the readings over several periods of the input signal. If the input-signal frequency is divided to a lower frequency, the gate will remain open for a multiple of the input-signal period, so that the counter will display the number of clock pulses for 10, 100, 1000, or more periods. A typical counter configuration for the period-averaging mode appears in fig. 4. The frequency-divider chain is split so that both the time-base oscillator and/or signal frequencies are divided to obtain the desired resolution and number of periods which are to be averaged. The counter will display the period measurement, regardless of the number of periods averaged, simply by having the

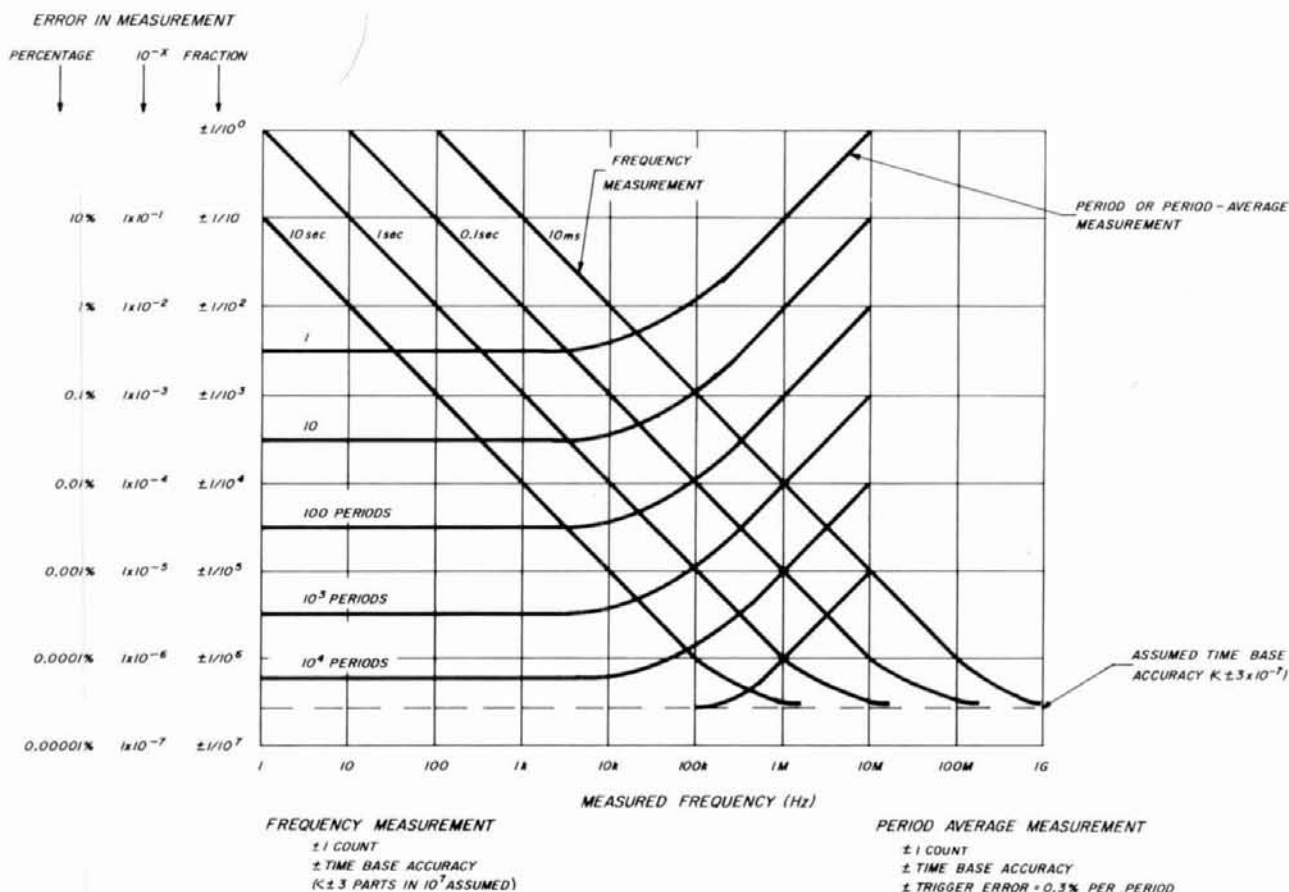


fig. 5. Measurement accuracy of a counter having a 10-MHz time base with an assumed accuracy of better than 3×10^{-7} .

display decimal point moved as the periods-averaged switch is changed.

Period averaging reduces the possible trigger error by a factor equal to N , the number of periods averaged, so that the error for this mode is \pm time-base error \pm (trigger error)/ $N \pm 1$ count. If enough periods are averaged, the trigger error can be reduced to a value which may be of little significance. It must be remembered, however, that the gate time increases by the same factor, which may make the measurement time quite long. For example, a 20-Hz signal has a period of 0.05 second; averaging 100 cycles results in a gate time of 5 seconds; averaging 1000 cycles entails a 50-second gate time, which is normally too long for convenient measurements.

From the preceding discussion, we can deduce that there is a point below which period or period-averaging measurements provide a more accurate reading than a corresponding frequency measurement. This can be calculated, taking into account the 1-count ambiguity, time-base error, trigger error, and gate time. More conveniently, it can be plotted, as shown in **fig. 5**. These curves apply to a counter having a 10-MHz time base of the accuracy specified, and indicate which measurement mode should be used for the desired measurement accuracy.

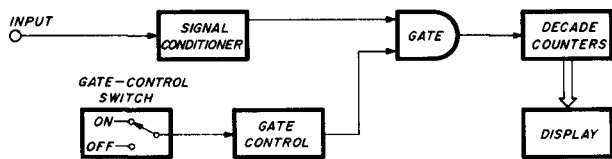


fig. 6. Functional block diagram of a totalizing counter. The gate-control switch can be either a manual switch or an internal switching circuit.

In many instances, measuring the ratio of two frequencies is a time-saving procedure. A typical case might involve designing or troubleshooting a phase-locked loop, where the output frequency is a discrete multiple of a reference oscillator. Since the output frequency may be divided by a factor of up to several thousand within the loop, an error or glitch causing a one-count error in this division may not be readily apparent unless a ratio measurement is made.

In conjunction with **fig. 1**, we discussed the method by which frequency is measured. Another way of defining this measurement is to state that the counter displays the ratio of the input frequency to the clock frequency. By using an internal clock whose frequency is known, the ratio can be displayed in megahertz, kilohertz, or hertz. If an external signal were used in place of the time-base oscillator,

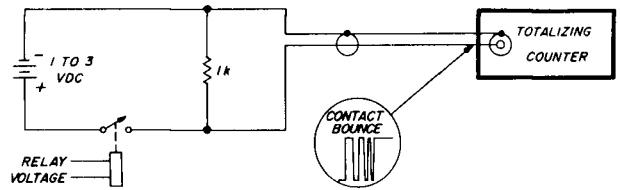


fig. 7. Using a totalizing counter to measure contact bounce. When relay voltage is applied, the counter will display the number of contact bounces.

the counter would still display the ratio of the two frequencies, except that it would no longer be in hertz or a multiple thereof (unless the external frequency were the same as the time-base frequency). Counters which provide specifically for ratio measurements incorporate provisions for changing the display to a dimensionless number, and position the decimal point accordingly.

Many counters do not have an apparent capability of measuring ratio, but can actually be used in this mode. If the counter has provisions for using an external time-base oscillator, the reference signal against which ratio is to be measured can be introduced into the external time-base connector. It is necessary that the amplitude of this external reference signal be as specified for the counter being used, that its frequency be within the range that the counter will accept as an external time base, and that the internal time-base oscillator frequency be known.

The ratio of the input signal frequency to the external reference frequency is determined from the expression

$$\frac{f_{sig}}{f_{ref}} = \frac{f_{ctr}}{f_{int}}$$

where

- f_{sig} is the input frequency
- f_{ref} is the external reference frequency
- f_{ctr} is the frequency displayed on the counter
- f_{int} is the internal time-base oscillator frequency.

totalizing

Perhaps the simplest function of which an electronic counter is capable is that of totalizing, or accumulating, a count of input events. Because this mode does not require a time base, as indicated in **fig. 6**, it probably should have been covered previously as the most basic counter circuit. However, totalizing is not usually a function of low-priced counters, nor does it have major applications in amateur work; therefore I have delayed discussing it until the modes of greater interest were described.

The gate-control switch shown in **fig. 6** can be

either a manual switch or an internal switching circuit actuated by the input signal. Switching the gate-control switch to *on* resets the decade counters to zero and allows the processed input signal to pass through the gate for the length of time that the switch is held on. When the switch is turned *off*, the count stops and the number of input events which has occurred is displayed on the counter.

An application which is of interest in these days of digital logic circuits is that of measuring contact bounce. **Fig. 7** shows a simple circuit which permits such a measurement for either a relay or a manually actuated switch. When voltage is applied to a relay coil (or a manual switch is operated), the contacts will usually open one or more times after the initial

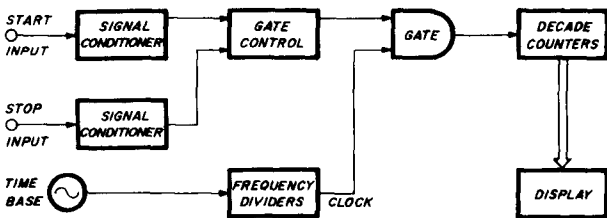


fig. 8. Functional block diagram of a time-interval counter, which counts the number of clock pulses between the time the gate is opened by the START input and closed by the STOP input.

closure because of the elasticity of the switch materials. This results in the waveform shown in the illustration, which is applied to the counter. The counter will then totalize and display the number of bounces.

time interval measurements

A counter may be used to measure the time interval between two input events, but this mode of operation requires two input-signal-conditioning circuits and a more complicated gate-control circuit; it is therefore found only in the more expensive professional instruments. As shown in **fig. 8**, the gate control has two inputs, one from each of the signal conditioners. The gate is opened by the processed *start* input, allowing the accurate clock pulses to pass through to the decade counters until the *stop* input closes the gate. Thus the counter will display the time interval between the two input signals.

The start and stop points are determined by the triggering levels and slopes selected by circuits in the signal conditioners. The time-interval resolution is limited by the clock frequency, and is subject to the same ± 1 -count ambiguity as all other measurements. As with period averaging, this ambiguity can be reduced for repetitive signals by averaging the time-interval measurements. When averaging, the

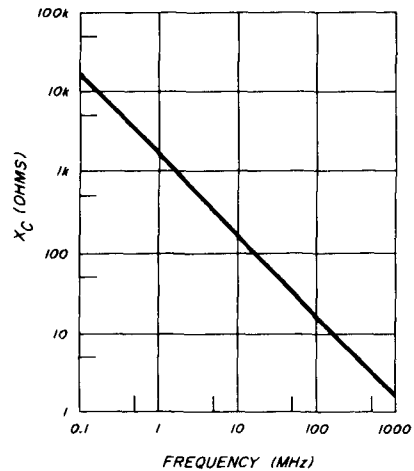


fig. 9. Reactance of 120 pF (typical for a high-impedance counter input with a 3-foot or 1-meter cable) plotted against frequency.

ambiguity becomes $\pm 1 \text{ count} \div \sqrt{N}$, where N is the number of time intervals averaged.

An important application of time-interval measurement is the accurate determination of pulse width. The signal under measurement is applied to both the *start* and *stop* inputs. If the *start* channel is set to trigger on the positive slope, and the *stop* channel on the negative slope (or vice versa), the counter will indicate the time interval between the leading and trailing edges of the input signal. Adjustment of the triggering levels will permit the measurement to be made between the desired points on the edges.

The upper frequency limit of the modern basic counter is dependent on the type of digital logic devices used in the signal conditioner and the first decade counter. This frequency limit may be as high as 50 MHz for conventional TTL, 120 MHz for Schottky TTL, and 250 MHz for ECL. Above those frequencies, prescaling is generally used to increase the frequency range, up to about 1300 MHz.

Prescaling simply means that the input frequency is scaled, or divided, down to one which is within the basic range of, and is measured by, the basic counter. The divisor may be any integral number. If the prescaler is external to the counter, it will usually divide by 10 or 100, so that the frequency can be read directly from the counter after you have mentally multiplied the counter reading by 10 or 100, as applicable. If the prescaler is built into the counter, it may scale by any integral factor.

The advantage of using an external prescaler is obvious — it permits extending the frequency range of an existing counter at relatively low cost. Its disadvantages become equally obvious after it has been used. First, there is the necessity of mentally moving the decimal point, since the counter is actually dis-

playing the divided input frequency. Second, one digit of resolution is lost for every decade of scaling. For example, a 145,600.0-kHz signal measured with a scale-by-ten prescaler will read 14560.0 kHz on a

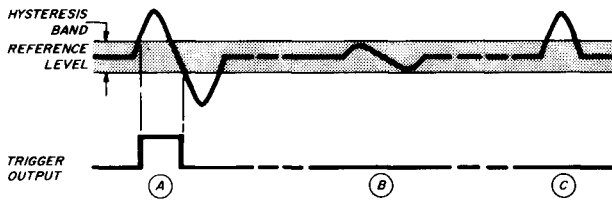


fig. 10. The effect of triggering hysteresis. The waveform at (A) will result in output from the trigger circuit, while those at (B) and (C) will not because neither crosses both limits of the hysteresis band.

counter having a 0.01-second gate time (0.1-kHz resolution). Multiplying by ten yields a frequency of 145,600 kHz; the 0.1-kHz resolution is lost by scaling. It can be re-established only by increasing the gate time by a factor of ten, provided the counter has that capability.

If the prescaler is an integral part of the counter, mentally scaling and moving the decimal point is eliminated, since this will be accomplished in the counter when the mode is changed from direct count to prescaled count. Nevertheless, the loss of resolution remains. It can be minimized, however, by scaling by a factor less than ten, and simultaneously increasing the gate time by the same factor.

Suppose that the internal prescaler divides the input frequency by four. If the gate time is increased by the same factor, there will be no change in the number of signal pulses gated through to the decade counters, and the display will read out the correct frequency. Consequently, prescaling is accomplished with only a fourfold increase in gate time, which is generally acceptable.

Switching from direct to scaled operation may be carried out in one of three ways. If a single input connector is used, the counter mode is generally switched manually. If two separate input connectors are employed, one for low-frequency signals and the other for high-frequency inputs, the counter mode may be switched manually or automatically when the input signal is present at the high-frequency input.

input impedance

Counters which measure frequencies below 250 MHz or so usually present a high input impedance — typically 1 megohm shunted by 30 to 40 picofarads. Above that frequency, the input impedance is generally a nominal 50 ohms, although the vswr may be as high as 2.5:1. At audio and low radio frequencies, a high input impedance is normally desirable,

since it minimizes the load on the circuit under test. But just how high in frequency is this true?

Consider a counter with an input impedance of 1 megohm shunted by 32 pF, which is used with a three-foot (91cm) cable made from RG-58C/U coax. The capacitance of RG-58C/U is 29.5 pF per foot (96.8 pF per meter), so that the total shunt capacitance presented to the circuit under test is approximately 120 pF. The reactance of this shunt capacitance, plotted against frequency, is shown in fig. 9. It can be seen that the reactance drops to approximately 1300 ohms at 1 MHz, and is only about 130 ohms at 10 MHz. So the input impedance can no longer be considered high. On the other hand, if the counter had a nominal 50-ohm input, you would

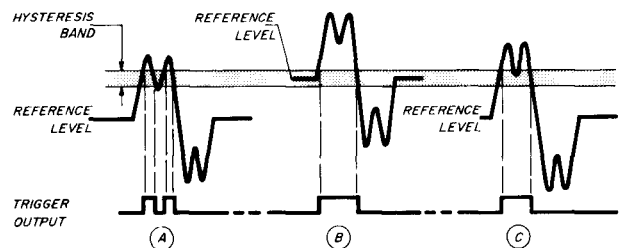


fig. 11. Erroneous counting caused by harmonic distortion is shown at (A). The false count can be eliminated by adjusting the level control, as indicated at (B), or by increasing the signal amplitude, as shown at (C).

know the loading effect, within the limits defined by the specified vswr.

Suppose that you had to check the frequency of a 70-MHz crystal oscillator which was designed to feed a 50-ohm load. If your counter has a 50-ohm input, all is well. However, if it has only a high impedance input, the shunt capacitance of the counter plus a cable will more than likely load down the oscillator and change the frequency, if it continues to oscillate at all. Fortunately, a relatively inexpensive accessory will solve the problem. By using a 50-ohm feed-through termination* at the counter connector, a 50-ohm interconnecting cable will be reasonably well terminated, and will present a load close to 50 ohms at the oscillator.

The same thing may be accomplished by using a 20-dB loss pad at the input connector of the counter, provided that there is enough signal to trigger the counter after having been attenuated by the pad.

Even at low frequencies, the shunt capacitance may be too high for certain applications, such as checking filters. Capacitive loading can be reduced by using a 10X oscilloscope probe. Such probes typi-

*Such as the Heath SU-511-50, Hewlett-Packard 10100C, Tektronix 011-0049-01, Systron-Donner 454, and other similar types.

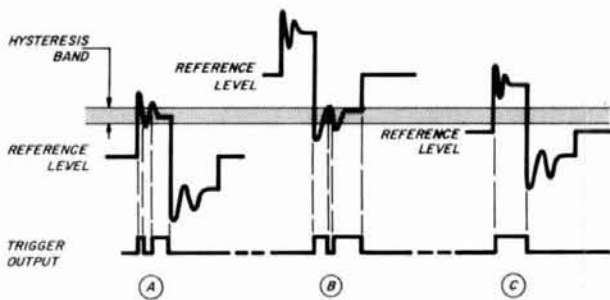


fig. 12. False counting caused by ringing is shown at (A) and (B). Proper adjustment of the reference level and/or amplitude, as shown at (C), corrects the fault.

cally present a 10-megohm resistive load shunted by 5 to 15 pF, but of course attenuate the signal by a factor of 10.

input signal levels

One of the parameters invariably specified for a counter is sensitivity, generally in millivolts, but often in dBm for 50-ohm inputs. This indicates the minimum signal needed at the *counter input* to ensure reliable triggering. Of equal, and possibly more importance, however, is the maximum signal which may be applied to the input without damaging the instrument.

For high-impedance inputs, the maximum signal voltage is usually specified as the sum of a dc value plus a peak ac value. The peak ac value may vary with frequency, going down as the frequency increases. The sum of ac plus dc is limited by the input blocking capacitor; the limiting ac value alone is a function of the input device in the signal conditioner. To be safe, when measuring at any point in a circuit where dc is present, always use an external blocking



B&K Precision 1820 Universal Frequency Counter will measure frequency from 5 to 520 MHz, and permits high-resolution period measurements from 5 Hz to 1 MHz. Decimal point position and unit-of-measure display is selected automatically for best resolution (photo courtesy B&K Precision).

capacitor of the smallest value which will permit reliable triggering. And if there is any possibility of the ac signal exceeding the specified maximum for the counter, use an external attenuator or dividing probe.

For low (50-ohm) impedance inputs, the maximum signal level is limited by the input circuit of the signal conditioner. This level is generally much lower than

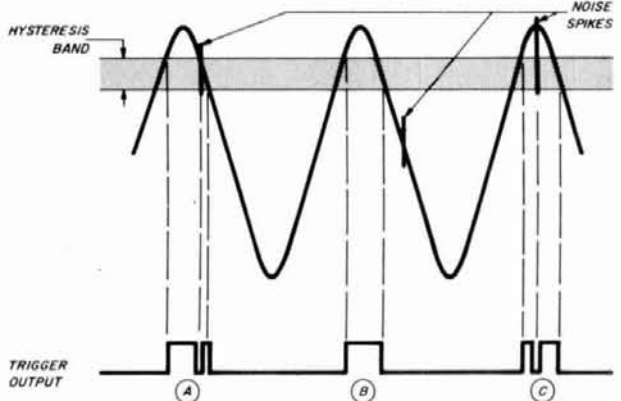


fig. 13. Spurious counts can result from noise on a signal when the noise is of sufficient amplitude to cross the hysteresis band.

for high-impedance inputs, and is typically between +19 and +27 dBm (2 to 5 volts rms across 50 ohms). Because of the relatively high cost of the high-frequency input device and the possibility of applying excessive power from a transmitter, the 50-ohm inputs are fuse-protected in many counters which can function at 500 MHz and higher.

Although it should be obvious, the following warning must be included: *Never connect a counter directly to a transmitter or any other high-power signal source!* Use a short length of unshielded wire as an antenna at the counter input connector, an inductive coupling loop at the end of a shielded cable, and/or an attenuator of sufficient power rating. The counter you save may be your own!

If the counter is battery-powered, and there is no direct connection between it and the circuit or generator under test, the counter should be grounded. This will reduce noise pick-up, especially when using a counter with a high input impedance.

triggering

The signal-conditioning circuits in all counters include a trigger circuit which, as previously stated, provides output pulses whose amplitude and wave-shape are compatible with the counter circuitry which follows. The sensitivity of the counter depends on the threshold level of the trigger input and the amplification between it and the input of the

counter. If the amplified input signal has insufficient amplitude to reach the threshold level, the instrument will not count or will perform erratically.

All trigger circuits have a hysteresis band, through the limits of which the input signal must pass in order

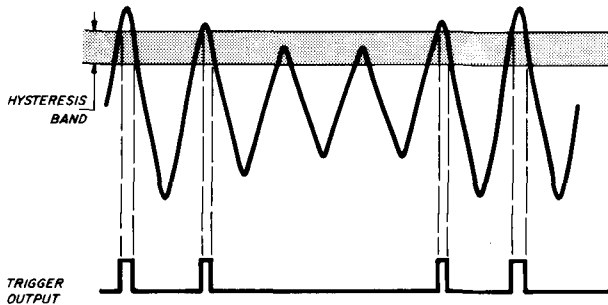


fig. 14. Amplitude modulation of the input signal can cause missing counts when the signal amplitude is too low.

to result in an output pulse. Fig. 10 shows three input signals in relationship to the hysteresis band. Sine wave A crosses both the upper and lower limits of the band, and will actuate the trigger circuit; the amplitude of sine wave B is too small, so triggering will not occur; and waveform C crosses the upper threshold, but not the lower, so again no output will be produced by the trigger circuit.

It is the action of this hysteresis effect which can result in erroneous counting which is so confusing to a relatively inexperienced operator. Suppose that the input to the counter were a sine wave with considerable second-harmonic distortion, a not uncommon situation. In fig. 11A, the amplitude of the signal is such that the positive half-cycle crosses the hysteresis-band limits twice, instead of once. The trigger circuit will generate two output pulses for each input cycle, and the counter will display twice the fundamental frequency of the signal. If the counter has a level control, which adjusts the reference level at the input of the trigger circuit, it can be adjusted to eliminate the false count, as shown in fig. 11B. If there is no level control, as is the case with most low-priced counters, the problem can be eliminated by increasing the amplitude of the input signal, as depicted in fig. 11C.

A similar problem may arise when measuring the frequency or period of a signal comprised of fast pulses. If the interconnecting cable is not terminated in its characteristic impedance, or if other impedance discontinuities exist, ringing will occur on the pulses. If the ringing traverses the hysteresis band, as shown in fig. 12A and 12B for two different reference levels, a false count will result. Proper adjustment of the signal amplitude and reference level, indicated in fig. 12C, will provide the correct count.

Another way of solving the ringing problem, which is useful when the reference level and/or amplitude cannot be changed, is to use a low-value resistor (100 to 1000 ohms) between the circuit point under test and the counter cable. This resistor, in conjunction with the cable and counter input capacitance, integrates the pulse and minimizes the pulse aberrations which reach the counter.

Figs. 13 and 14 illustrate two other conditions which can result in false counts. The noise transients on the signal shown in fig. 13 will cause additional counts, while amplitude modulation may result in missing counts, as shown in fig. 14, if the amplitude of the input signal is too small. In either case, the solution is the same as previously prescribed — change the reference level and/or the signal amplitude.

In our earlier discussion of period and period-averaging measurements, it was stated that the trigger error resulting from noise on the input signal contributed to the measurement error. This is shown in fig. 15, in which a sawtooth wave is used to demonstrate the effect of slope, or slew rate, on the trigger error. It can be seen that the noise voltage on the relatively slow rise-time can cause a much greater trigger error than that which occurs on the fast fall-time. Thus we can see that the trigger error can be minimized by triggering on the steepest portion of the input signal to the counter. For a sine wave, this will be that part of the waveform at the zero axis, leading to the conclusion that a signal of the maximum possible amplitude should be used.

It should be apparent from the preceding discussion that an input attenuator on the counter can be of considerable help in establishing the correct input level to the trigger circuit. In many of the lower-priced counters an attenuator has been omitted because of cost and because it was felt that limiting diodes at the input of the signal conditioner would protect the input device. The latter reason is valid

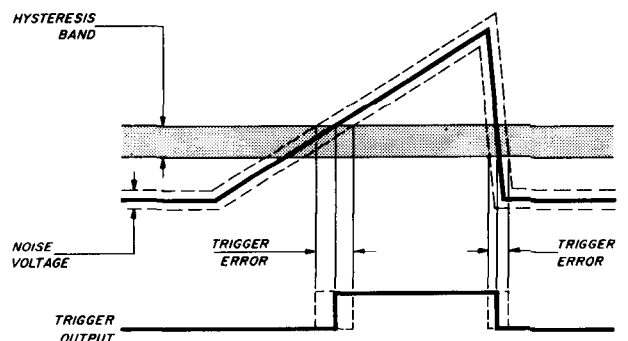


fig. 15. Trigger error in period and period-averaging measurements, caused by noise in the input signal. The error is minimized by a fast slew rate through the hysteresis band.

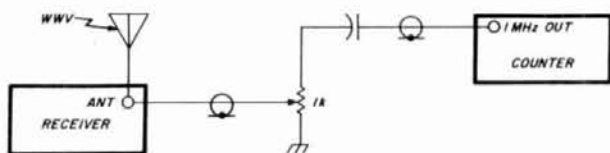


fig. 16. Test setup for calibrating the time-base frequency against WWV.

only where overload is considered, for even a two-position attenuator can be extremely valuable in eliminating false counting.

time-base calibration

Unless an oven oscillator or TCXO is used as the time base in a counter whose oscillator circuit is energized continuously, the oscillator frequency should be checked, and adjusted if necessary, whenever accurate measurements are to be made. Be sure, however, that the counter is fully warmed up before checking or recalibrating the oscillator.

In order to calibrate the time-base frequency, either a standard of known accuracy or a communications receiver capable of receiving WWV is required. If the standard is at least five times more accurate than the best resolution of the counter at the time-base frequency, it can be applied directly to the input of the counter. Then adjust the time-base oscillator frequency control for the correct frequency read-out on the counter.

A more accurate adjustment may be made if the counter has an output connector from which a 1-MHz time-base signal can be obtained. Connect this output to the vertical input of an oscilloscope, and connect the output of the frequency standard to the horizontal input.* The scope will display a Lissajous pattern, which will probably be moving. Adjust the time-base frequency control until a stationary pattern is obtained.

If the counter is to be calibrated against WWV, a time must be chosen during which transmissions are received with an absolute minimum of fading. Select the highest receiving frequency possible (e.g. 15 MHz) to achieve the greatest calibration accuracy. The calibration technique involves obtaining a visual beat indication on the receiver S-meter, and adjusting the time-base oscillator frequency for as close to a zero-beat as possible.

In order to obtain a good beat null, the time-base

*These connections are based on the assumption that both the horizontal and vertical amplifiers in the oscilloscope will pass a 1-MHz signal. If the counter has a time-base output of higher frequency, it can also be used, provided that it is within the frequency range of the scope. The lower of the two frequencies (the standard and the counter time base) should be connected to the horizontal input, since the horizontal frequency limit is usually lower than the vertical.

signal which is applied to the receiver must be of the correct amplitude relative to the signal level from WWV. Since we cannot control the latter, we have to be able to vary the signal level from the counter. If the counter has a 1-MHz output from the time-base circuit, make the connections shown in fig. 16. If a time-base signal is not brought out to a connector on the counter, substitute an insulated wire for the coax shown connected to the receiver antenna terminal and place it near the counter time-base oscillator or frequency-divider chain. In either case, the harmonic of the 1-MHz signal should result in a low-frequency



Leader LDC-822 Frequency Counter measures frequency to 80 MHz and features selectable gate time and input attenuation (photo courtesy Leader Instruments).

beat with WWV. (This will not be an audible beat unless the time-base oscillator is very far off frequency; more likely it will be observed as a rhythmic variation in the S-meter reading.) Adjust the potentiometer shown in fig. 16, or change the position of the insulated wire, to obtain the deepest beat null on the receiver S-meter.

It should be possible to adjust the time-base frequency so that the beat-frequency period is several seconds, which corresponds to a remarkably accurate short-term frequency setting. To demonstrate this, assume that eight beats are observed on the S-meter in a 60-second period. The beat frequency is therefore equal to $8/60$, or 0.133 Hz. If the beat is measured at 15 MHz, the error is $0.133/15 \times 10^6$, or 8.9×10^{-9} . Of course, this degree of accuracy may hold only for a short period of time, because the stability of the counter time-base oscillator, unless it is an oven type, is nowhere near that good. Nevertheless, highly accurate measurements may be made until the counter is turned off or a temperature change affects the time-base oscillator.

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simplifying the digital frequency counter

Some innovative ideas
for high-resolution
counters using CMOS —
TTL devices

Radio amateurs have had a long history of pioneering in electronics. My experiments with the new IC technology resulted in the following article. I

and pulse conditioning. Power required is 5 volts at 1.5-2 amps or more! Only one example was found using CMOS. It still used 21 chips and was limited to 4 MHz.

A number of the newer CMOS combinations are available from which to choose for simplifying the counter and decreasing power requirements; the Intersil 7208/7207A combination seemed to be the most promising, so it was chosen for this project.

The counter is shown in **figs. 1** and **2**. Both circuits comprise a complete frequency counter with 1-Hz resolution from below 20 Hz to above 50 MHz, and with 10-Hz resolution to above 300 MHz. Nine ICs and four transistors are used including the power supply and prescaler. Current drain is 200 mA for frequencies below 50 MHz; an additional 130 mA is required for higher frequencies.

Device description. The heart of the counter is the 7208 CMOS chip. This device contains a 7-decade

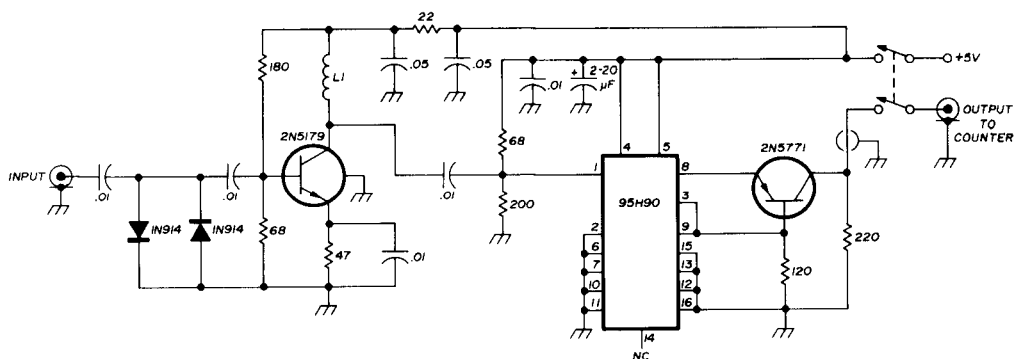


fig. 1. 300-MHz prescaler for the high-resolution counter.

decided to find out just how much the digital frequency counter could be simplified by using some of the newer CMOS combination chips.

Today's literature shows that much can be done. Almost everyone seems to be using TTL technology; for example, the 7490/7447 combination dating from the 1960s: an 8-digit counter using 7490 devices requires 24 chips for the main counter alone, plus 12 or more (typically) for a crystal-controlled time base

counter, multiplexer for the display, 7-segment decoder, digit and segment drivers, and the logic required for blanking, reset, input inhibit, and display on-off — a regular one-man band!

The 7207A is another CMOS chip that teams up

By Holton E. Harris, W1WP, 5 Newtown Turnpike, Westport, Connecticut 06880

with the 7208. It contains a high-stability oscillator and a frequency divider for dividing the 5.24288-MHz crystal frequency to obtain the 1-second gate required for counting. It also provides outputs to synchronize the multiplexer for the displays as well as short pulses for latching and resetting the counters.

CMOS and TTL combination. The one real deficiency of CMOS is that it is slow — i.e., low-frequency response. The 7208/7207A combination alone, with power supply, LED displays, and crystal can be used to make a complete counter (as shown later) but it won't count above about 6-7 MHz. Almost all of the remaining circuitry in this counter is the old work-horse TTL, purely and simply to extend the frequency range.

The 74196 counter is used to get to 50 MHz. This device is similar to the 7490 but has a higher frequency range. A prescaler could have been used instead, of course, but a synchronized counter was preferred because it retains accuracy in the least-significant digit. The 9368 performs the decoder/driver functions for the least-significant digit of the LED display.

When Intersil first announced the 7208, the companion driver was the 7207. This combination gave a gate signal of 0.1 second, so the resolution was 10 Hz instead of 1 Hz. But the worst of it was that, if all seven decades of the 7208 were used, the most-significant digit was 10 MHz, which was above the frequency range of the counters in the 7208. In effect, therefore, all seven decades could never be used.

The 7207A, announced in late 1976, corrects the problem. It has a gating pulse of 1 second duration, permitting 1-Hz resolution. Unfortunately, to make room for the added counter stage, the output buffers on the reset and enable lines had to be eliminated. These signals therefore can't be used to drive the

7432 TTL gate directly, and an extra buffer was required. The 4049 IC is a hex buffer/converter designed for this CMOS to TTL interface. The signal is passed through the buffer twice with a 4.7k pullup

FND-503 COMMON CATHODE DISPLAYS

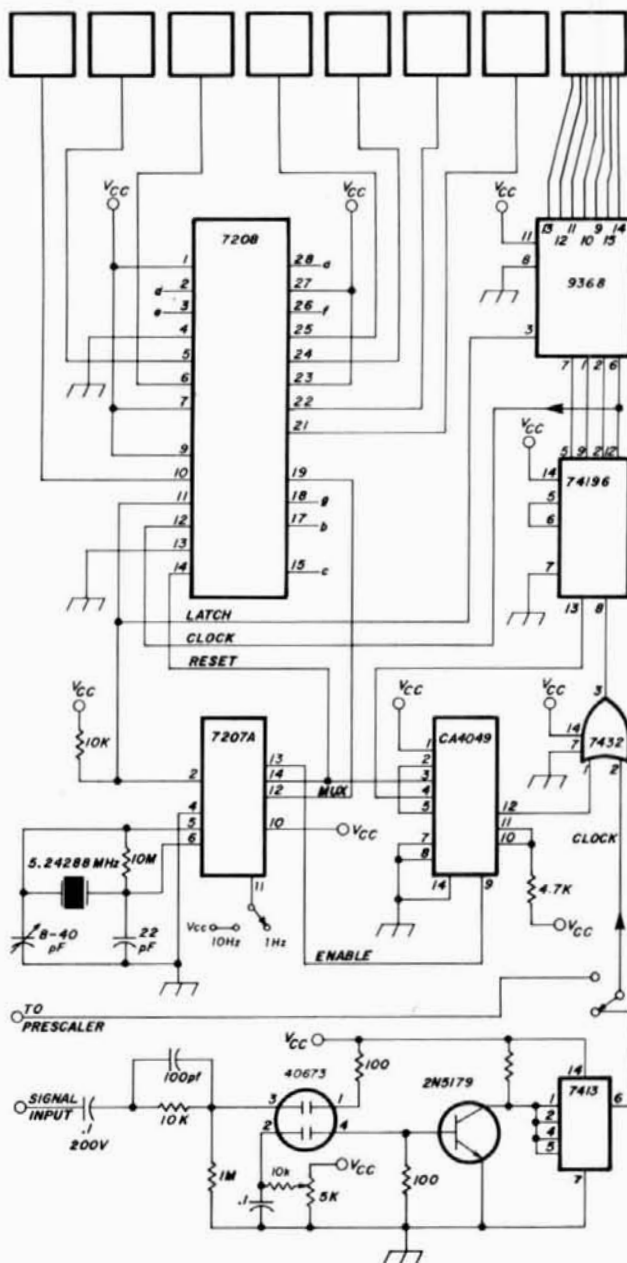
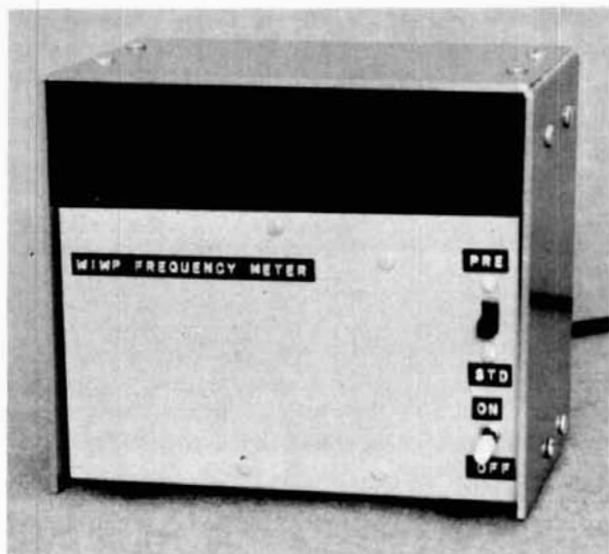
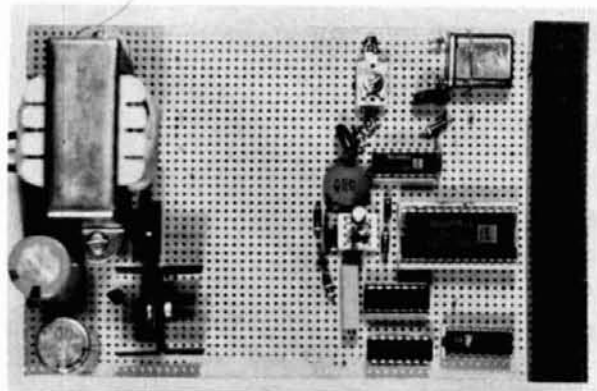


fig. 2. Schematic of a 50-MHz counter with 1-Hz resolution. In each counter example, the resolution is decreased by a factor of ten by taking pin 11 of the 7207A to V_{CC} .

resistor at the interstage in an effort to make the gating pulse rise and fall times equal. Equal rise and fall times are important, since the accuracy of the entire counter depends on the time the counting gate is open. The 7432 is an OR gate, which removes the





Pegboard construction of the counter shown in fig. 3. The power supply is shown at left. The two sections were cut apart for mounting into a cabinet.

input signal from the counter input except during the counting interval.

Circuit description. The 5.24288-MHz crystal and the oscillator portion of the 7207A form a high-precision time reference that determines counter accuracy. The crystal is a low drift, 5 ppm type. If higher precision is desired, an oven should be used.

The 7207A has a binary divider chain that divides the oscillator frequency by 2^{20} to yield a 0.5-Hz square wave that sets the counting interval to exactly 1 second. A higher frequency is picked off to synchronize the 7208 for multiplexing the output displays. Pulses are also generated to set the display latches and reset the counters at the appropriate times.

After passing through the 4049 buffer, which increases the power level to drive the TTL gates, the timing signal is applied to the 7432. For the 1 second during which the signal is low, the 7432 allows the input signal to pass through to the 74196 counter. The 74196 counts to 10, puts out a pulse to the 7208, then repeats. The 7208 has seven decade stages that similarly count successive decades.

After the 1-second counting period, the gate to the 7432 goes high and the counters stop. The latch pulse from the 7207A transfers this count into the latches, and the decoder/drivers in the 7208 convert the count to 7-segment form and pass it to the display lines. Another pulse from the 7207A through the 4049 then resets the 74196 and the seven decades in the 7208 to zero. Meanwhile the multiplexer in the 7208 energizes each LED display in sequence.

The low-frequency preamplifier was cribbed from Stark¹. An additional stage of amplification was added after the input fet to increase sensitivity. The circuit thus consists of a two-stage amplifier driving a Schmitt trigger. The Schmitt trigger turns on at one

level and turns off at another, much lower, level so that slowly rising signals can't cause jitter, and false triggering is avoided.

The prescaler to extend the range is likewise conventional. It was lifted from the excellent article by Kitchens². Why argue with success?

One point may appear puzzling. If you check the Intersil 7208 data, you'll find the multiplexer input brought into pin 16. This leaves two prior CMOS gates open, which is bad practice. I couldn't get the device to work at all with that connection. Bringing

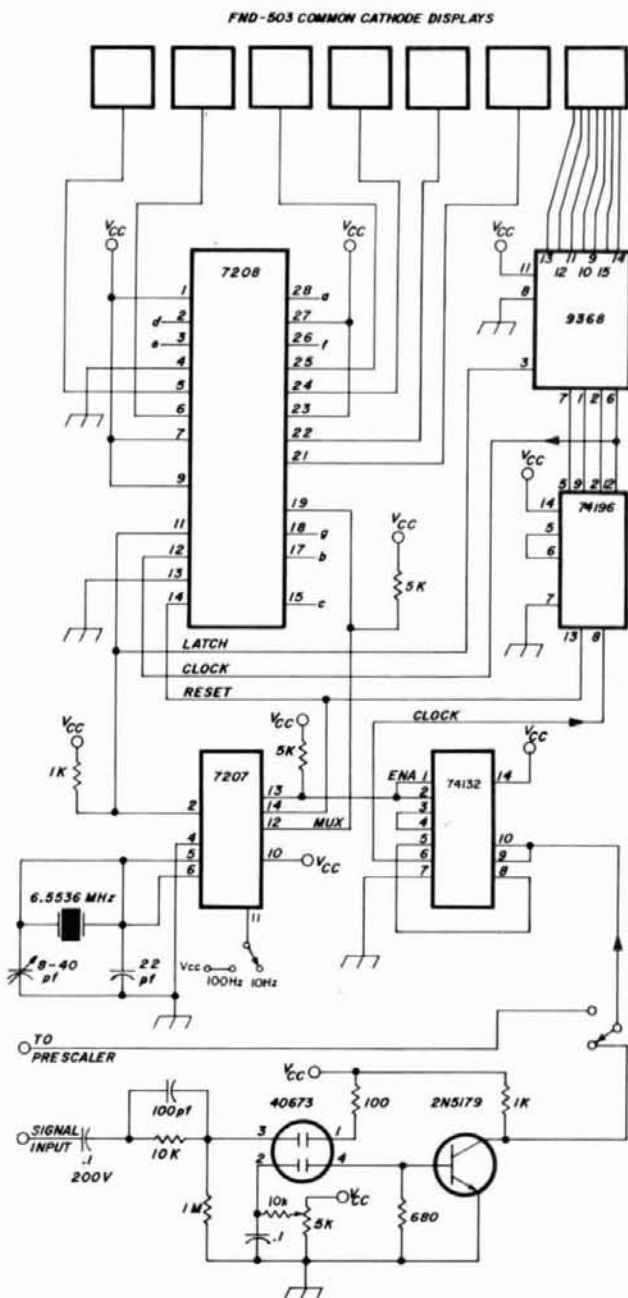


fig. 3. Schematic of a 30-MHz counter with 10-Hz resolution.

FND-503 COMMON CATHODE DISPLAYS

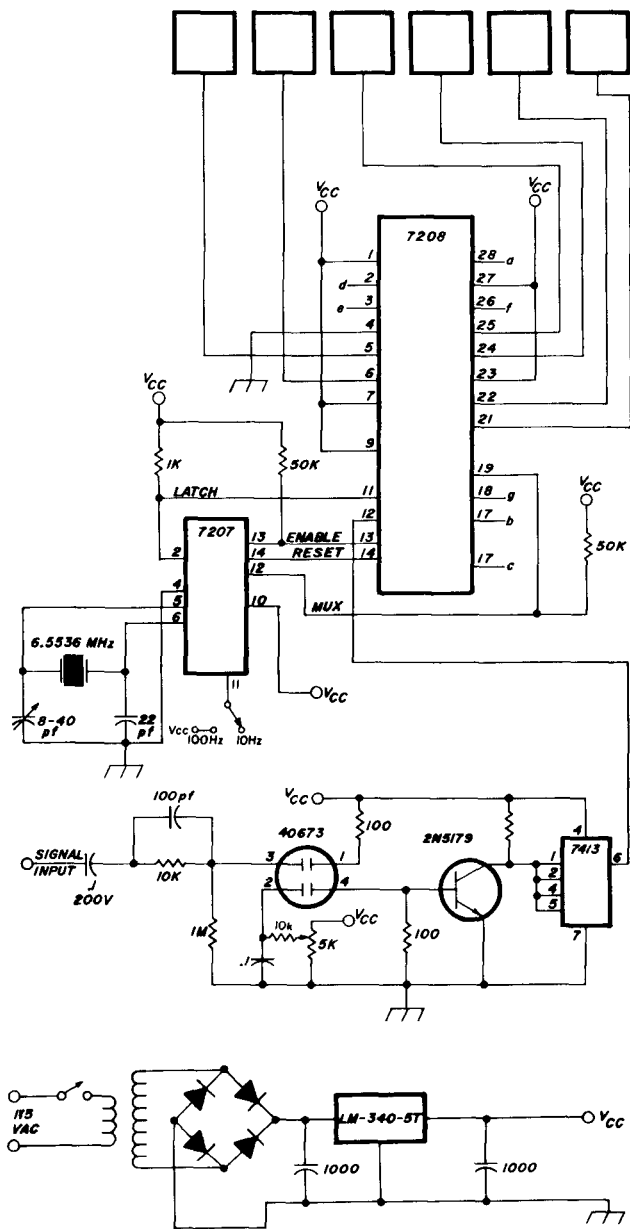


fig. 4. Simplified counter that operates up to 6 MHz with 10-Hz resolution.

the multiplexer input to pin 19, as shown in fig 2, ties everything down and cures the problem.

The new National 5881, 581, and 583 multiple display LEDs were used because when they first came out they were even cheaper than surplus standard units. If I had it to do over again, I'd use the conventional FND-507 even if more wiring is required.

construction

I built the circuit on pegboard using point-to-point wiring with a Vector wiring pencil. A more enter-

prising builder might wish to make a PC board. The only adjustments required are to tune the crystal to exact frequency using the 8-40 pF trimmer and to adjust the 5k pot in gate 2 of the 40673 prescaled transistor (fig. 1). This latter adjustment is easily made by connecting a voltmeter between the 7413 pin 1 and ground and adjusting the trimmer pot until the dc level is about 1.3 volt.

other forms of the counter

The one big disadvantage of the counter described above is that its sampling time is quite slow. Obviously, if a resolution of 1 Hz is required, the counting gate must be open for one full second. Another second is allowed for reset and latch, so that the total time to update the display is 2 seconds. Under some circumstances, this can seem like forever!

For most applications, a 1-Hz resolution isn't really necessary, and the circuit of fig. 3 can be used. Here the 7207 is used as the oscillator/timer. This device gives a counting interval of 0.1 second and a total period of 0.2 second. The counter thus updates 5 times per second instead of once every 2 seconds, and action seems much more normal. The resolution is, of course, only 10 Hz instead of 1 Hz, which is sufficient for most purposes.

The counter shown in fig. 3 includes further simplifications. The functions of the 7413 Schmitt trigger and the 7432 OR gate are combined in a single IC, the 74132. This device is a quad 2-input NAND Schmitt trigger, which does both jobs. This change could, of course, be made in the circuit of fig. 2.

Fairchild FND-503 displays are used here. The 100-ohm limiting resistors were eliminated for a brighter display, which increased total counter current from 200 to 300 mA.

A further simplification is possible if the highest frequency to be counted can be limited to 6-7 MHz. In this case the 74196, together with the 9368 and 7432, can be eliminated. The circuit is shown in fig. 4. The 7208 IC provides the counting function. Resolution is again 10 Hz, and the seventh digit in the readout is omitted, since it can never be used in this instance. However, the circuit makes a mighty simple counter.

I'd like to express my appreciation to my co-worker, Josh Schwartz, without whose excellent and timely suggestions this project couldn't have been completed.

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1. Peter A. Stark, K2OAW, "A Modern VHF Frequency Counter," 73, July, 1972, page 5.
2. Marion D. Kitchens, Jr., K4GOK, "Vhf prescaler for Digital Frequency Counters," ham radio, February, 1976.

ham radio

how to modify your frequency counter for direct counting to 100 MHz

Simple IC circuit
can be used in an
existing frequency counter
to extend its
direct counting range
to 100 MHz

When the prices of TTL integrated circuits first dropped to the point where the average amateur could use them to build a frequency counter, the maximum operating frequency was about 25 MHz. Then came the Fairchild Semiconductor 95H90 prescaler with its 350 MHz capability. This promised reliable measurements at 220 MHz, but some of the earlier homebrew counters were frequency limited and could not use the prescaler to its full advantage. Next came the Fairchild 11C90, which was rated at 650 MHz and the 1300 MHz counter from Hewlett-Packard. It will only be a matter of time before an inexpensive 1 GHz prescaler reaches the market. It hasn't arrived yet, but I decided to redesign my counter to adapt to such a prescaler, when it comes.

There has been a flood of second-generation TTL ICs appearing in the past few years that would both increase the speed and diminish the size of a modern frequency counter over one using the standard 7490 decade counter, 7475 latch, and 7447 LED decoder/driver. I decided that a complete rebuilding of my counter, while interesting, could not be justified. In this project described in this article a single board, containing a gate circuit and the first decade counter, is substituted for the original. This board can be used in any counter which has a positive gate-enable pulse, a positive or negative reset pulse, and enough room to sandwich in the modification.

circuit operation

The gate function is performed by one gate of a 74S00. When a positive enable pulse is applied to one NAND gate input (during the counting period), the gate will act like an inverter to a square wave ap-

By Ian MacFarlane, WA1SNG, 102 Columbus Avenue, Greenfield, Massachusetts 01301

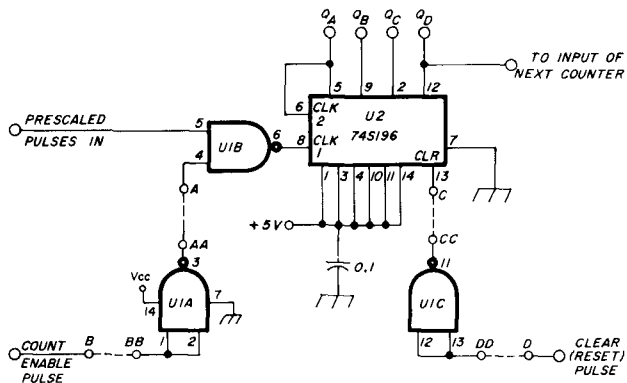
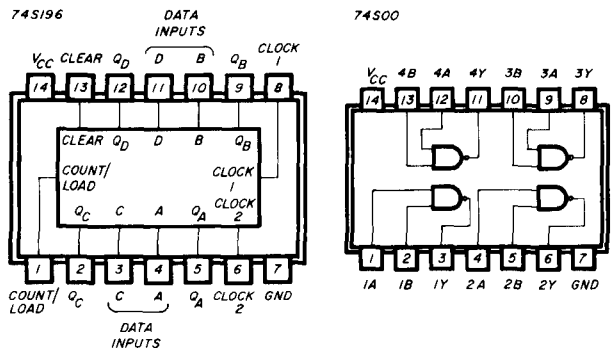


fig. 1. Schematic diagram of the 100 MHz counter. If your existing counter has a positive reset pulse, jumper C to CC and D to DD as shown here; if the reset pulse is negative, jumper C to D. If the count enable pulse is negative, jumper A to AA and B to BB as shown; if positive, jumper A to B. Not shown here but included on the circuit board are two bypass capacitors from the power supply to ground, C2 and C3.

plied at the second input. As the gate enable pulse falls to a logic zero, the NAND output rises to one, and is not affected by state changes at the second input. Consequently, the gate operates as an on/off switch to control the flow of the signal to be counted into the first decade counter. If your counter uses a negative enable signal, provisions have been made on the circuit board to use an extra NAND gate to invert the count-enable pulse.

A 74S196 presetable decade counter is used to extend the guaranteed count frequency to 100 MHz. The typical frequency limit is 140 MHz. To disable the preset feature, all presetable data inputs must be held at a logic one, along with the *Count/Load* input. Binary-coded-decimal counting is provided by connecting out Q_A to *Clock Input 2*, and injecting the signal to be counted into *Clock Input 1*. Counting occurs on the negative transitions of the count pulse. The reset (*clear*) pulse must also be negative, unlike that used for 7490 decade counters, so one NAND gate is used as an inverter. If your counter also uses a negative reset pulse, it is possible to directly drive the *clear* input, so long as the fanout of your clear line is not exceeded by the 74S196 requirements. Unlike a 7490, the clear input of the Schottky chip is two standard Schottky loads (or 2.5 standard 7400 series loads). If your counter has a maximum of eight digits, you should not encounter any difficulty. However, don't try to drive the reset line with an L or LS series device because it will not be able to supply sufficient current. The 74S196's four binary-coded-decimal outputs (Q_A through Q_D) are connected to the respective inputs of the latch.

Signetics Corporation manufactures a plug-in replacement for the 74S196 called the 82S90. This might be the easiest device to find for some people, but for me the difficulty of trying to locate a 74S196



was second only to finding an 82S90. Neither device is stocked by most suppliers, but I have been advised by Active Electronic Sales that they can supply SN74S196N ICs at \$3.45 each.* These are Texas

*Active Electronic Sales Corporation, Box 1035, Framingham, Massachusetts 01701. They have a minimum order requirement of \$10.00, plus a \$1.00 postage and handling charge.

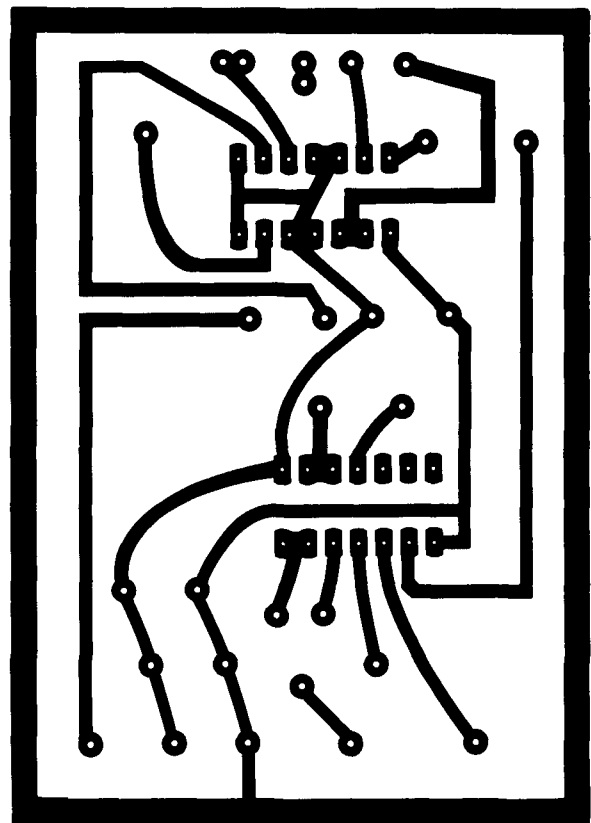


fig. 2. Full-size printed circuit board for the 100 MHz counter stage. Component layout is shown in fig. 3.

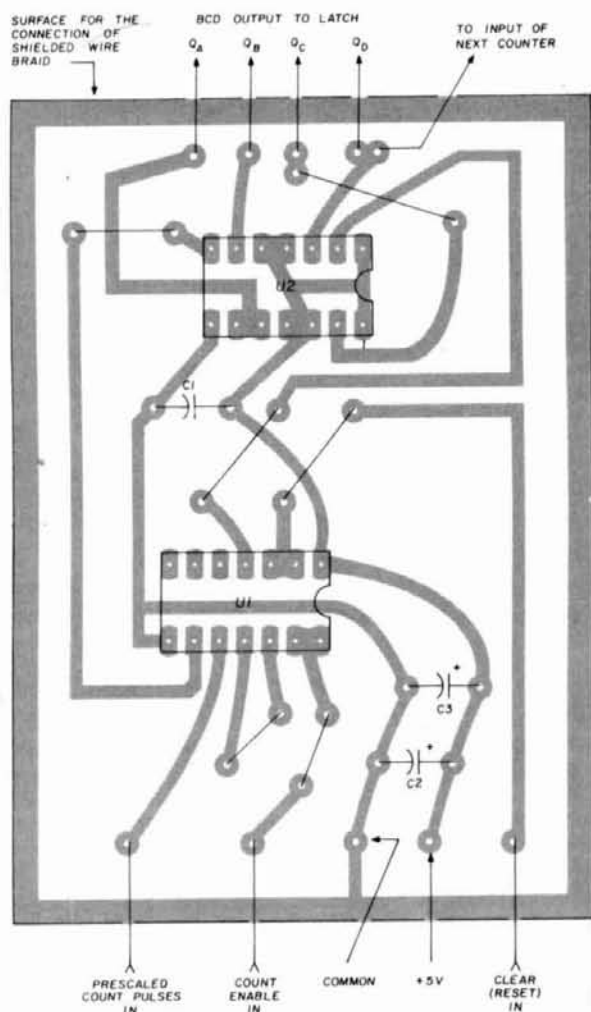


fig. 3. Component layout for the 100-MHz counter. Bypass capacitor C2 is a 0.1 μ F or larger disc ceramic; C3 is a 47 μ F, 10 volt, tantalum or electrolytic.

Instruments devices with current date code devices, but since they are not normally stocked, delivery time is three weeks.

construction and installation

Circuit details and IC pin-out diagrams are shown in fig. 1. The printed-circuit pattern and component placement information is given in fig. 2 and 3. The

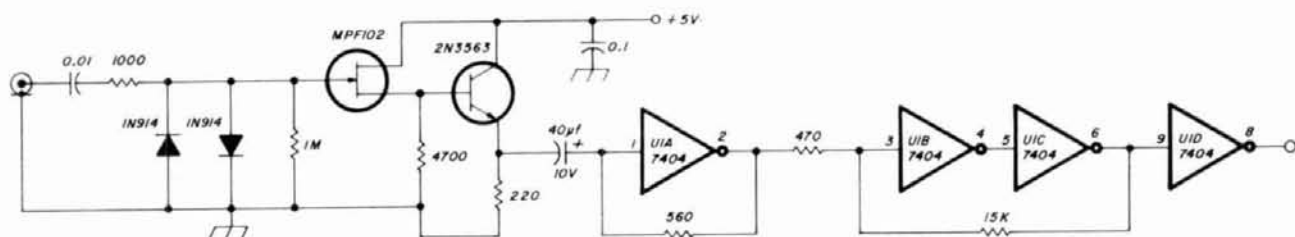


fig. 4. Counter input and shaper stage used in many homebrew frequency counters was originally used in a commercial instrument; it can be modified for use at 100 MHz as discussed in the text.

power supply lead should be a separate shielded cable (audio type is fine) connected directly at the power supply. Connect the shield to the common power supply ground point and attach the other end to *common* on the PC board. To prevent ground loops, do not ground the board where it is mounted.

Disc or rectangular ceramic bypass capacitors are used liberally to discourage transients but their values are not critical — the larger the better. One large tantalum or electrolytic capacitor is used to eliminate low-frequency transients. Shielded cable may also be used to bring the *Gate-Enable*, *Reset*, Q_A through Q_D , *Count Pulses In*, and *Count Pulses Out* signals into and out of the board. Shielded cable is not really needed in practice, but it will reduce the amount of rf floating around the counter, and may prevent jamming of the input circuit. If shielding is used, ground the shield at the board and trim back the braid at the other end.

modifying the input circuit

Extending the range of the input circuit to 100 MHz is not absolutely necessary because a prescaler will extend the counting range anyway. However, direct counting to 100 MHz is easy to accomplish in counters which use the two most popular input circuits.

One popular input circuit consists of a fet amplifier, often an RCA 40673, driving a Schmitt trigger; a typical circuit is shown in fig. 4.¹ Almost any dual-gate mosfet will operate well above 100 MHz, so no change is required there. If the Schmitt trigger is a 7413, it can be replaced directly with a 74S13. Discrete Schmitt triggers built up from 7400 or 7404 gates can usually be replaced with Schottky devices without changing any external resistors.

Another popular counter input circuit first appeared in *QST*² during a review of the HUA Electronics 1BC-1a frequency counter. This circuit (fig. 5) has been widely duplicated over the years with varying success. Input sensitivity estimates have been reported from 10 to 300 mV, so a few suggestions are in order for anyone having difficulty with the HUA circuit. Problems stem from two sources: the extreme sensitivity of the unit, and the substitu-

tion of one TTL sub-series for another (such as a 74LS04 for the 7404). Sensitivity can be so great that leakage from other counter circuitry jams the input, establishing a threshold that must be exceeded to trigger the counter. The solution is to shield the entire input circuit in a minibox, using coax to bring the signal in, and shielded wire, bypassed on both ends with 0.1 μ F and 47 μ F capacitors, for the +5 volt supply. The shield should be grounded at the power supply and connected as the common return on the circuit board.

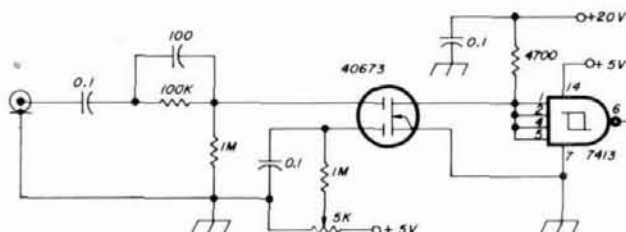


fig. 5. Popular counter input stage based on the 40673 mosfet which can be modified for operation to 100 MHz as described in the text.

The second problem concerns differing bias requirements between the standard and LS series. When substituting a 74LS04 for a 7404, it is impossible to simply interchange one for the other without changing the external resistors. This problem is not as acute when replacing the 7404 with a 74S04, because the difference in input current between the Schottky and standard series is slight. Only one resistor seems to be at all critical, and this is the 560-ohm feedback resistor across the first hex inverter. A sure way of obtaining the most performance from your particular combination of devices is to optimize this resistor. Connect a 1000-ohm potentiometer in place of the 560-ohm resistor and adjust it for maximum sensitivity. Using an ohmmeter, measure the resistance (after carefully removing the pot from the circuit) and replace it with a fixed value.

conclusion

For a total cost of less than \$20 for all materials, including printed-circuit drafting and etching supplies, this circuit can put new life into an otherwise outmoded counter. The 74S196 will typically extend counting to 125 or even 140 MHz, which should accommodate the output of prescalers for many years to come.

references

1. Peter Stark, K2OAW, "A Modern VHF Frequency Counter," 73, July, 1972, page 5.
2. Gerald Hall, K1PLP, "HUA Electronics Frequency Counter Model 1BC-1a," QST, April, 1972, page 60.

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simple front-ends

for a 500-MHz frequency counter

Basic front-end design is stressed in this adaptation of the Intersil seven digit CMOS frequency counter

The advent of the Intersil CMOS counter chip pair, the ICM 7208 and 7207A, has resulted in many designs which take advantage of the low cost and simplicity these chips make possible. I wanted a complete, 500-MHz counter, but with no bells and whistles. Since most hams do not need more than 100 Hz resolution at 500 MHz, the counter was restricted to seven digits. With most of the basic work already documented in Intersil application notes, the only real design problem was to fabricate a suitable front-end that interfaced between the prescaler and the CMOS integrated circuit.

As an engineer I have fought many debugging wars, enough to know that sure things don't always work the way they are designed. Accordingly, I researched the literature for the tried and true. As a result, what follows is not entirely original, but it has the redeeming virtue that it works.

50-MHz front end

The front end is composed of Q1 through Q7, with a sensitivity of 300 mV at 30 MHz, falling to 1 volt at

50 MHz (see **fig. 1**). Most of the credit for the design goes to Marvin Moss, W4UXJ, who adapted it from several other similar designs.

As a single device, the fet used for Q1 does not give satisfactory performance. Therefore, Q2 is used to bootstrap the voltage at the source of Q1 to more closely equal the voltage at the gate of Q1. This also greatly reduces the effect of Q1's input capacitance at high frequencies, thus maintaining the input impedance with increasing frequency. Capacitor C2 compensates for the small amount of rolloff that will never-the-less occur. R5 allows quiescent point adjustment for maximum sensitivity. Since the remainder of the front end is dc coupled, R5 also sets the operating point for the rest of the amplifier.

Q3 has a fixed gain of approximately 6.8, the ratio of R7 to R8. R8 also raises the input impedance of this stage to minimize loading on Q1-Q2. C3 tends to raise the gain of Q3 to compensate for the rolloff in gain brought about by the output capacitance of Q3 being in parallel with R7.

Q5 and Q6 form a high-gain amplifier with hysteresis. Basically, Q5 is an emitter follower driving the common base amplifier, Q6. The base of Q6 is held at approximately 6 volts by R12 and CR4. Because of the high voltage gain of the common base configuration, the signal alternately drives Q6 close to saturation and cutoff. However, R11 creates a small amount of positive feedback, or about a 0.6 volt hysteresis. This is necessary to avoid extraneous counts on low-frequency signals.

Transistor Q7 is used as an emitter follower to drive the TTL counter circuitry that follows. The low value of R14 is necessary, since TTL likes to see a low source impedance in the low state. A common-emitter stage might work here, but would require more components and would draw just as much current.

By John H. Bordelon, K4JIU, 1132 Research Drive, Marietta, Georgia 30066

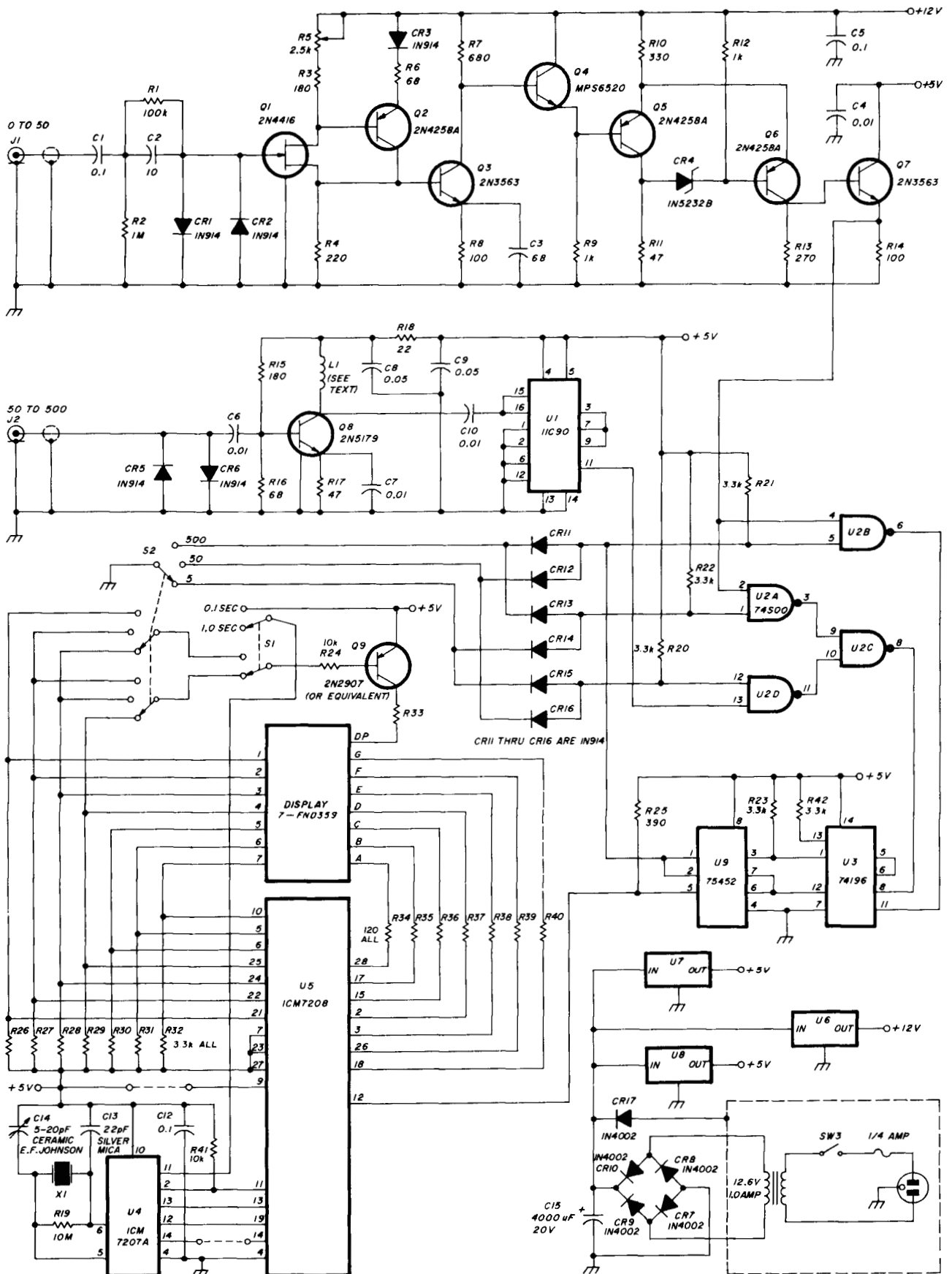


fig. 1. Schematic diagram of the complete frequency counter including the low-frequency (0-50 MHz) and high-frequency (50-500 MHz) preamplifiers. The displays have been wirewrapped into a set of sockets and are shown here as a multiplexed assembly. Other displays can be used but they must be common-cathode types. The crystal frequency is 5.242880 MHz (available from International Crystal). All resistors are 1/4 watt, 10 per cent tolerance. The 5 and 12 volt regulators are MC7805 and MC7812 ICs.

JOHN H BORDELON KEITH R HARRIS
• FREQUENCY COUNTER

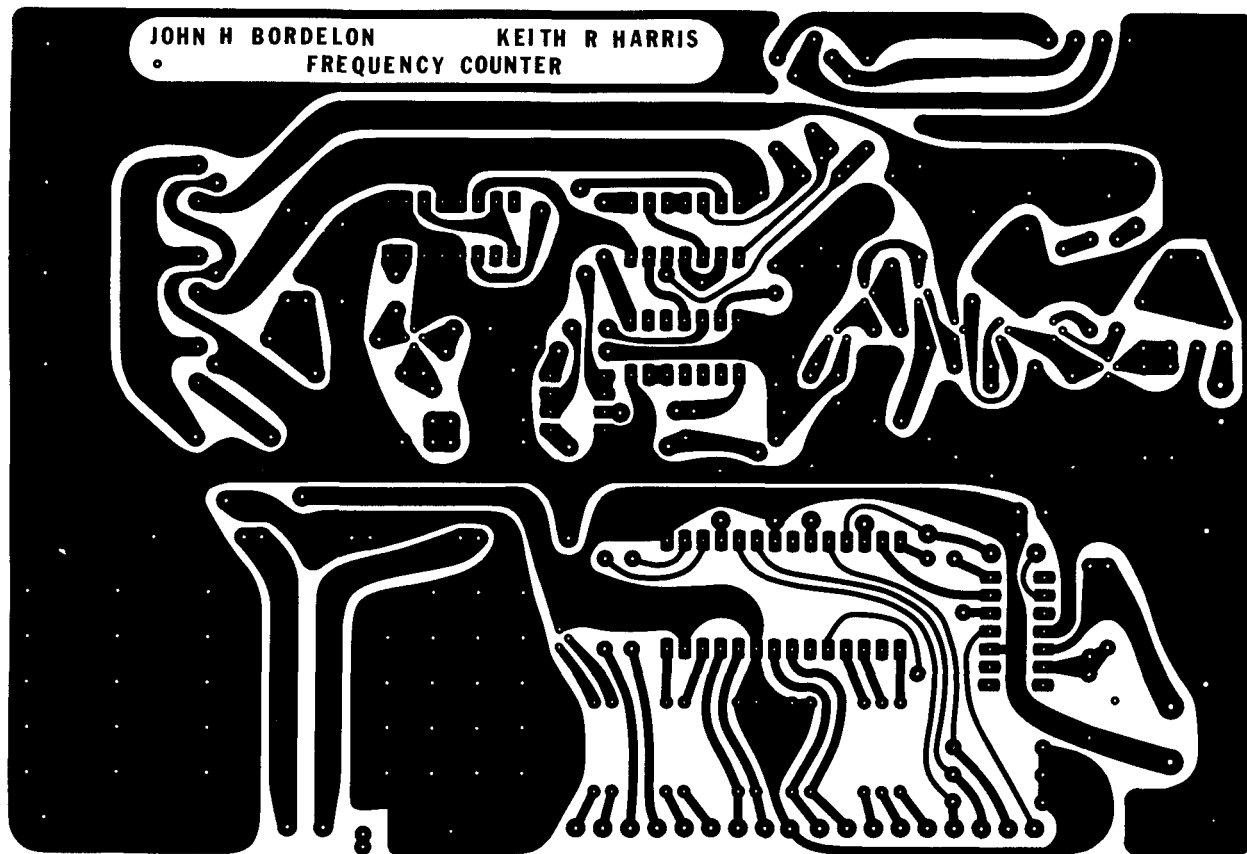


fig. 2. Foil pattern for the 500-MHz frequency counter.

The 500-MHz prescaler uses the popular and readily available Fairchild 11C90. This ECL device not only has a minimum count rate of 500 MHz, it directly drives TTL, and has built-in biasing networks to insure optimum sensitivity. It does not have the older 95H90's reputation for being picky about supply voltage.

The 2N5179 has got to be one of the most bang-for-the-buck transistors around. Reviewing articles for using it to drive 95H90s to 300 MHz led me to think it would almost certainly work well at 500 MHz, especially in view of its f_T of over 1 GHz. Such is the case, and the sensitivity at 500 MHz is 200 mV RMS, and 25 mV at 50 MHz. A 0.3m (1-foot) antenna yields solid counts on a 1-watt, 2-meter mobile 6.1m (20 feet) away. The low-input impedance is not a handicap in most cases, and a virtue in some. L1, in combination with the 2N5179's output capacitance, serves as a low- Q resonant circuit.

the complete counter

Referring again to the schematic, I chose a very common IC, the 74196, for the 50-MHz prescaler (U3). U2, a 74S00, routes the signals as directed by the range switch (S2). This also eliminates the need

for any front-panel rf switching. Q9 drives the appropriate display decimal point, under command of S1 and S2, to provide a display that always reads in kHz. The CMOS counter will not count to its specified maximum frequency when driven by TTL; the output voltage swing of TTL is too low. U9 cures this problem by providing a signal that swings from nearly ground to very near the supply voltage. Observant readers may notice that the signal goes through the 50-MHz prescaler, even when the 5-MHz range is in use. I've applied the signal to the Q_D data input; with the count/load control (pin 1) low, the Q_D signal will follow the data input. When pin 1 is high, inputs on pin 8 are accepted and prescaled by 10. Taking advantage of the architecture of the 74196 in this way eliminated the need for any extra gates.

Resistors R33 through R40 are current limiting resistors for the FND359 displays which I plugged into wire-wrap sockets. R26 through R32 are recommended by Intersil to preclude digit driver leakage from causing "ghosting" on the display.

The power supply is conventional. CR17 allows use of external battery power and protects the battery pack should ac be applied to the counter power supply. It should be noted that U8 supplies U4, U5,

and U9. Otherwise, U5 may be damaged by any voltage difference between the two regulators. I designed my circuit board (fig. 2) to take care of this.

construction

A printed-circuit board seemed the only reasonable way to construct the counter. * Fig. 2 shows the layout of the board. The two front ends use good high-frequency layout practices. Although two-sided boards are generally thought to be *de rigueur* for such situations, they are not always necessary. My counter does not talk to itself. One reason for this is the use of broad ground planes dividing and encircling some circuits. Note especially the layout of the 500-MHz preamp and prescaler.

The chassis is homemade, black anodized aluminum. Transfer letters were used for labeling, with clear spray enamel used to protect them. Enamel does not seem to cause smearing of the letters, but caution was exercised.

There are no adjustments to the 500-MHz front end. In use, a no-count condition may just as likely be an overload as well as indication of not enough signal.

The 50-MHz front end requires R5 to be adjusted for maximum sensitivity. The best way to adjust it is with the aid of a signal generator having a variable output attenuator. This way the input signal may be reduced in small steps, touching up the adjustment of R5, until maximum sensitivity is obtained.

Adjustment is most critical at 50 MHz, so final alignment should be done there. Because my 74196 was of questionable origin, the maximum count rate was just shy of 50 MHz. If operation at 50 MHz seems impossible, drop down in frequency and come up in small steps, touching up R5 each time. Don't forget that if the 74196 is not up to snuff, the 11C90 will be handicapped as a result. Check the former before blaming the latter.

conclusion

Two of these counters have been built and are operational. About twenty-two more are in various stages of construction around the southeastern United States. Minimum cost, exclusive of the circuit board, is about \$50, with a maximum cost of about \$85.

*Etched, drilled, and plated circuit boards, of G-10 epoxy fiberglass, with 7 pages of documentation, are available for \$15, postpaid, from the author.

reference

1. Marion D. Kitchens, Jr., K4GOK, "VHF Prescaler for Digital Frequency Counters," *ham radio*, February, 1976, page 32.

ham radio

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TRANSMITTERS

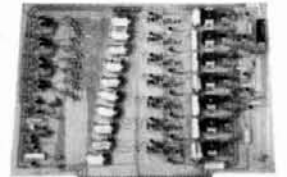
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precision temperature control for crystal ovens

The National LM3911 temperature-controller IC is featured in a circuit for precision temperature measurement and control in commercially available crystal ovens

Many amateur stations are equipped with frequency synthesizers and digital counters. These circuits require a highly stable frequency source. The most commonly used frequency standard uses the 100-kHz quartz crystal. Such standards exhibit drift with changes in temperature. More recently, the trend has been toward the use of high-accuracy AT-cut crystals operating in the 4- to 10-MHz range. These crystals also exhibit some temperature drift. They must be used in a *good* crystal oven to obtain a stability of better than one part per million over extended periods.

Crystal ovens fall into two categories: on-off (or bang-bang) and proportional control. In the former, a bimetallic strip makes and breaks the heater circuit. Most inexpensive ovens used by amateurs are of this type. Temperature variations frequently exceed 9°F (5°C). These ovens are somewhat erratic in operation, especially the older units that have worn and pitted strip contacts. With proportional control, the heater current is continuously and automatically varied to maintain constant temperature.

This article shows how the National Semiconduc-

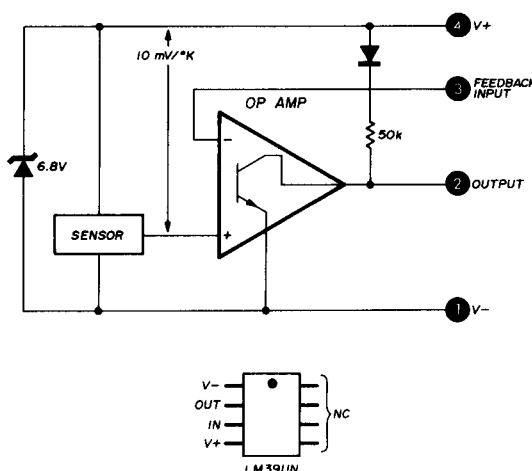


fig. 1. Block diagram of the LM3911N temperature controller. The pinout diagram is shown as a top view.

tor LM3911N temperature controller IC can be used for precision temperature measurement and control. This marvelous little device, selling for under \$2.00, makes it possible to maintain oven temperatures to better than 0.18°F (0.1°C). The LM3911N can be used to replace the thermostat in an existing oven or

By Fred Schmidt, K4VA, 3848 Parkcrest Drive, N.E., Atlanta, Georgia 30319

can be incorporated into an easily and inexpensively built oven.

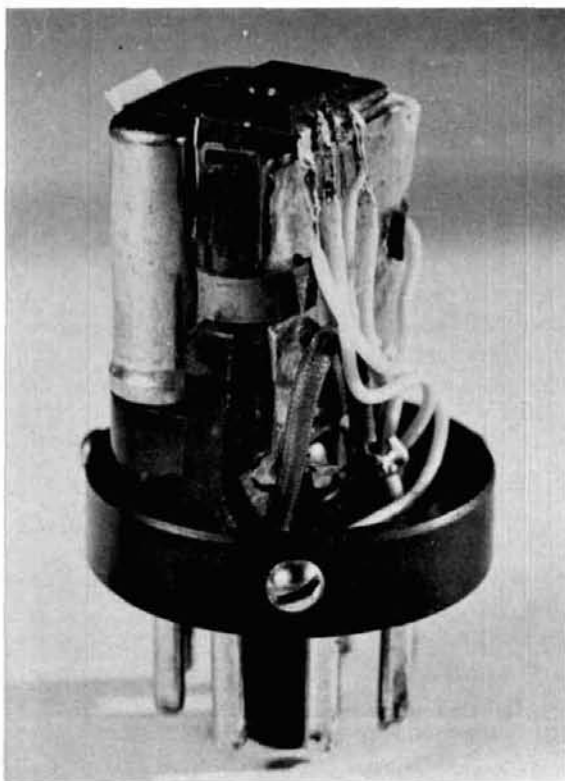
The LM3911N* consists of an operational amplifier, zener, and a temperature sensor in an 8-pin plastic DIP. The block diagram, taken from the data booklet is shown in **fig. 1**. The sensor develops 10 millivolts per degree Kelvin between the positive supply terminal on the op amp chip and its non-inverting input.

on-off and proportional control

Figs. 2 and 3, also from the device instruction booklet, show how the LM3911N can be used for simple on-off and proportional control respectively. In on-off control, the operational amplifier is used as a comparator. When the sensor temperature increases to the value at which the noninverting input voltage is equal to the inverting input voltage, the output switches from about 6 volts to a fraction of a volt. The output can be used to control oven heater current by means of a suitable power transistor. The voltage at the inverting input terminal determines the temperature at which the output switches.

The proportional-control circuit (**fig. 3**) generates a square wave at the output terminal. The duty cycle

*Available from Tri-Tek, Inc., 6522 North 43 Avenue, Glendale, Arizona 85301, Specifications and Applications Booklet, \$.80 (also, Radio Shack RS3911, Catalog 276-1706, \$2.19).



Installation of the LM3911 IC in a commercial oven.

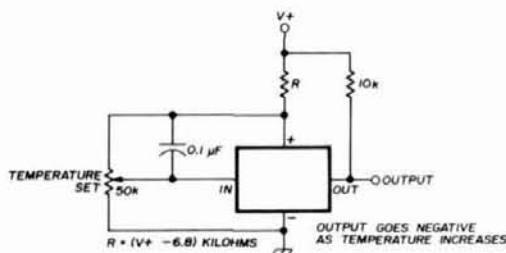
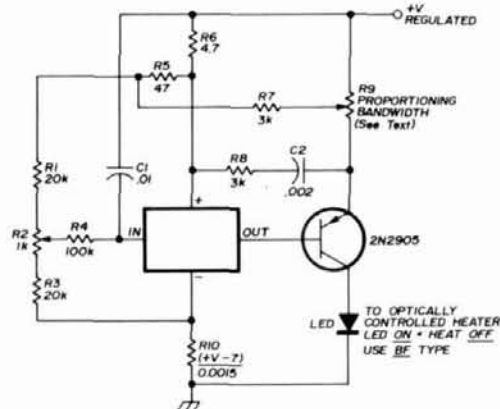


fig. 2. Circuit for on-off temperature control.

(ratio of OFF to ON time) is determined by the sensor temperature and the voltage at the inverting input terminal. Any departure of temperature from the desired value causes the duty cycle to change. This action is used to change the average heater current in such a way as to bring the temperature back toward the desired value. Proportioning bandwidth refers to the temperature range over which the output is a square wave. When the temperature is above



1. ALL RESISTORS 1/2 WATT COMPOSITION EXCEPT R1 AND R3 WHICH SHOULD BE CARBON FILM OR WIREWOUND FOR TEMPERATURE STABILITY. R2 IS WIREWOUND.
2. VALUES OF R1, R2 AND R3 ARE FOR 75C OVEN TEMPERATURE.

fig. 3. Circuit for proportional temperature control.

the bandwidth the output will remain low, and when the temperature is below the bandwidth output will remain high. The square-wave frequency is determined by the time constant $R4C1$. The frequency varies somewhat over the bandwidth and is maximum at the center.

The complete circuit for proportional control of an oven is shown in **fig. 4**. The 4N30 is a 6-pin dual in-line IC containing a light-emitting diode, which is optically coupled to a photo-sensitive darlington transistor. It drives a power transistor. The oven heater is in the power-transistor collector circuit. During the ON intervals of the square wave, the power transistor is driven to saturation. During the OFF intervals it is cut off.

The photo shows how the LM3911N is installed in an oven of the ON-OFF variety. The oven is a Bliley

TC0-1A surplus unit with a 6.3-volt, 0.85-A heater element. The thermostat was unsoldered and discarded.

The LM3911N was cemented to the inverted-U copper strip, which is used to conduct heat from the heating element to the crystal. The LM3911N die is on the base of the package. Therefore, it's important that the base be coupled as closely as possible to the heat source.

Pins 4-8 are not used electrically and should be bent in such a way as to make contact with the copper strip to help conduct heat into the package. Pins 1-4 are connected to pins in the oven socket.

testing

Performance can be checked by connecting a voltmeter across the heating element. When first turned on, the heater voltage will nearly equal the supply voltage. As the oven heats up, the heater voltage will go through several damped oscillations about a final value of about three volts. (The damped oscillation is caused by the thermal lag between the heating element and the copper heat conducting strip.)

The maximum temperature overshoot is about 1 degree C. Bandwidth control R9 should be set about two-thirds of the way down from the end connected to the supply line. R1 and R3 should be wirewound or metal-film resistors. (Composition resistors have a large temperature coefficient and would cause the oven temperature to change somewhat with changes in ambient temperature.)

R1 and R3 values can be changed if it's desired to operate the oven at a temperature other than 75°C. The maximum operating temperature of the LM3911N is 85°C.

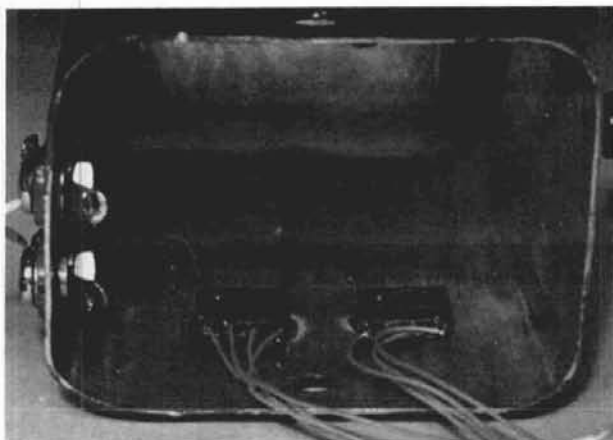
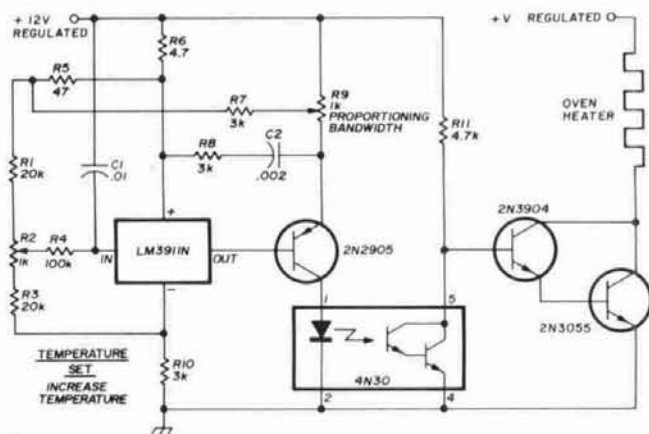


Photo of can inside showing the placement of the LM3911 ICs — one for temperature control and the other for temperature measurement.

Temperature calibration can be performed by drilling a small hole in the metal cover just above the crystal socket and inserting a thermometer into the



- NOTES:
1. ALL RESISTORS 1/2 WATT COMPOSITION, EXCEPT R1 AND R3, WHICH SHOULD BE METAL FILM OR WIREWOUND.
 2. VALUES OF R1, R2 AND R3 ARE FOR OVEN TEMPERATURE OF 75°C.
 3. +V SHOULD BE 5-6 VOLTS FOR 6.3 VOLT SURPLUS OVEN; 9-12 VOLTS FOR HOME BREW OVEN.

fig. 4. Complete circuit for a proportional temperature-control crystal oven.

oven. The thermometer will take a minute or more to reach final temperature.

homebrew oven

If a surplus oven isn't available you can make one easily and inexpensively. A homebrew oven offers several advantages:

1. It can be made large enough to include the complete oscillator circuit for enhanced stability.
2. There is room for an additional LM3911N for temperature measurement.
3. It can be constructed with very close thermal coupling between the heating element and the controller to avoid temperature oscillations (hunting).

The photo shows how an oven can be made with a surplus i-f transformer can. The unit measures 1-1/2 by 2 by 3-1/2 inches (38x51x89mm).

The heating element consists of no. 28 (0.3mm) enamelled copper wire wound directly onto the aluminum can. The wire should be close wound over a length of about 2-1/2 inches (64mm). After winding, check for short circuits to the can then paint with Red X Corona Dope (General Cement Catalog no. 50-2), or with Dipping Varnish (General Cement Catalog no. 56-2).

Connections are brought out to two screws,

which are insulated from the can by fiber shoulder washers. The oscillator circuit is built onto a PC board, which is held in place by two metal brackets. At the top of the board is a ceramic trimmer capacitor for frequency adjustment. A hole in the top of the can allows screwdriver adjustment for frequency trimming.

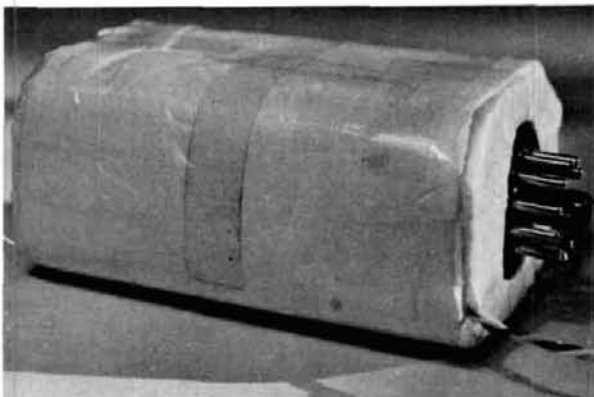
Two LM3911N chips are cemented to the inner surface of the can as shown in the photo. The second chip is used as an electronic thermometer. After assembly, the oven is insulated with two layers of styrofoam. A convenient source of this material is the 16-ounce (472ml) styrofoam drinking cup.

In the unit shown, the cold resistance of the heater winding is about 6 ohms. The regulated supply for the heater should deliver between 9 and 12 volts. For a smaller-size can it may be necessary to use smaller wire for the heater. A cold resistance of 6 to 8 ohms is about right.

thermometer circuit

The thermometer circuit is shown in **fig. 5**. Resistance values are for a 0-1 mA meter and a temperature range of 70° - 80°C (158-176°F). Other temperature and meter ranges can be obtained by changing the resistance values according to the equations in the appendix. It's not necessary to install a milliammeter permanently in the unit: a pair of terminal posts can be installed to allow temperature to be monitored with your multimeter.

Many AT cut crystals exhibit a turning point in the neighborhood of 60°C (140°F). Advantage should be taken of this by operating the oven at that temperature. The turning point can be found by observing the crystal frequency as the oven warms up. If the frequency initially drops then starts to rise, adjust the oven temperature to obtain the minimum frequency. The oven temperature should be at least 5°C above



The assembled oven in its styrofoam jacket.

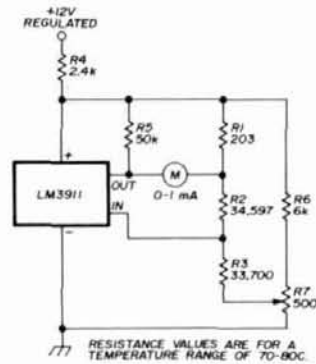


fig. 5. Circuit of the LM3911N IC as an electronic thermometer.

maximum ambient in the equipment in which it will be installed.

The performance of either oven leaves little to be desired. Temperature stability is reached in about 5 minutes. Temperature variation is too small to be detected on the thermometer calibrated from 70°-80°C (158°-176°F).

acknowledgement

I wish to thank Jim Bell, K4FUP, for suggesting the use of the LM3911N in this project.

appendix

The resistances for the thermometer circuit are calculated by the following equations:

$$R_1 = \frac{(6.85)(0.01)(\Delta T)}{I_M (6.85 - 0.01T_0)} \quad (1)$$

$$R_2 = \frac{0.01 T_0 - I_0 R_1}{I_0} \quad (2)$$

$$R_3 = \frac{6.85}{I_0} - R_1 - R_2 \quad (3)$$

$$R_4 = \frac{V^* - 6.85}{0.001 + I_M + I_0} \quad (4)$$

$$R_5 = 50K \quad (5)$$

$$R_6 = 6k \quad (6)$$

$$R_7 = 500 \text{ ohms}$$

where

ΔT = Meter temperature span (degrees C)

I_m = Meter full scale current (amperes)

T_0 = 5 + meter zero temperature (degrees K)

I_0 = Current through R_1 , R_2 , and R_3 at zero meter current (use 0.0001 amperes)

Notes:

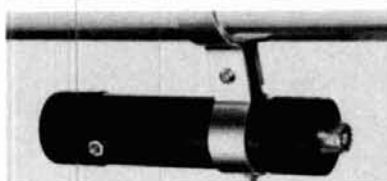
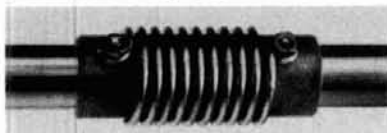
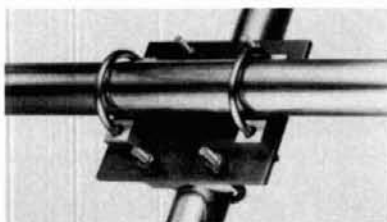
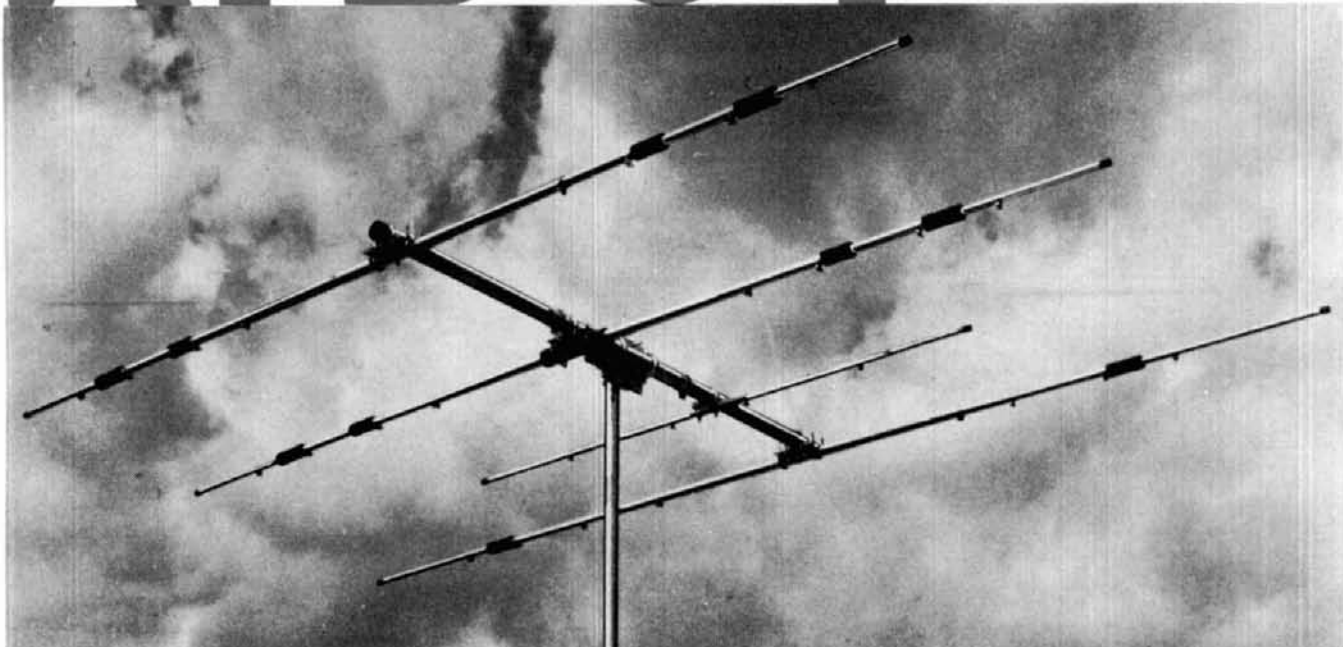
1. °K = °C + 273.

2. Resistors should be metal film.

3. If no thermometer is available for calibration set potentiometer to its midpoint.

ham radio

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VSWR -	1.5-1	WIND SURVIVAL -	90 MPH.
POWER HANDLING -	2000 WATTS PEP		
BOOM LENGTH/DIA. -	18' x 2 1/8"		
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TURNING RADIUS -	18'9"		

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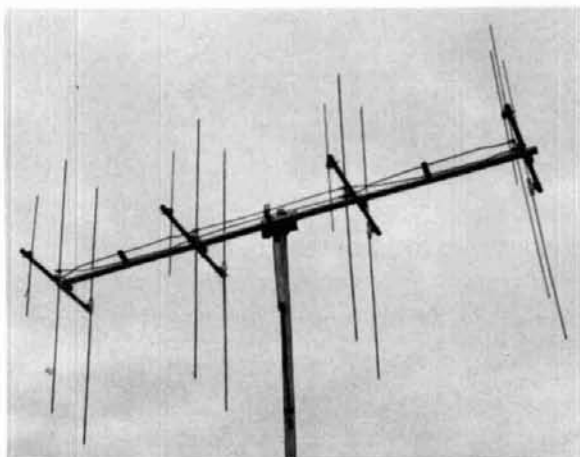


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VHF DX

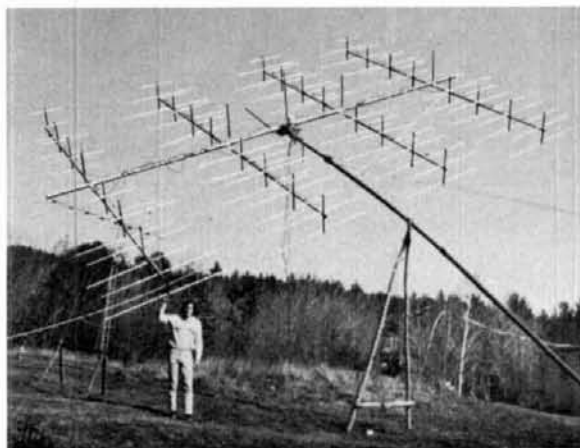
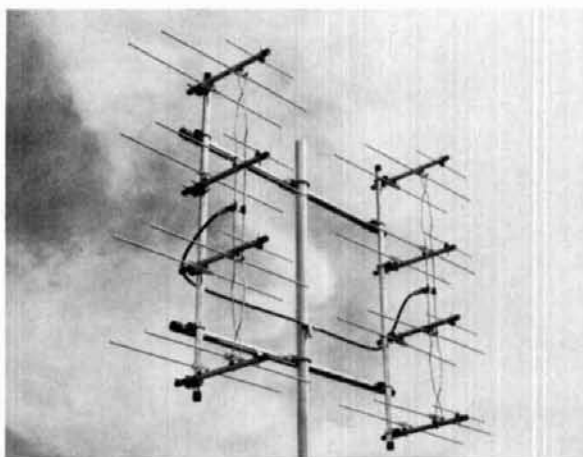


FM —

Enjoy the thrill of dependable long distance contacts on simplex or thru remote repeaters. The 20 element co-linear DX-Array offers a precise pattern with large capture area. This vertically polarized, horizontally stacked array provides a narrow beamwidth for the discriminating FM user. Wide impedance and gain bandwidths make the DX-Array a natural choice for the serious FM'er. A vertical polarization bracket, model DX-VPB, is required (support boom and mast not supplied). Seek out new horizons with DX-Array!

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

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20 Element DX-Array	DX-120	\$42.95	DX-220	\$37.95	DX-420	\$32.95
Frame & Harness (40 E.)	DXK-140	\$59.95	DXK-240	\$54.95	DXK-440	\$39.95
Frame & Harness (80 El.)	DXK-180	\$109.95	DXK-280	\$89.95	DXK-480	\$79.95
1-1 52-ohm Balun	DX-1BN	\$12.95	DX-2BN	\$12.95	DX-4BN	\$12.95
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satellite tracking —

pointing and range with a pocket calculator

How to use
a pocket calculator
to calculate
antenna pointing angles
and range
for an earth-bound
or satellite station

A frequent question in long-distance radio communications is where to point the beam. This article contains algorithms for RPN pocket calculators to calculate distances and headings between stations on the earth, and pointing angles and slant ranges to an earth satellite or the moon.

earth stations

Given the longitude λ_1 (west longitudes are +) and latitude ϕ_1 (north latitudes are +) of station 1 (home) and the longitude λ_2 and latitude ϕ_2 of station 2; the algorithm below* calculates A , the initial heading or pointing angle (north reference clockwise

azimuth) from station 1 toward station 2, and D , the great-circle distance between stations.

Alternatively, if λ_2 and ϕ_2 are the coordinates of the sub-satellite point and h is the height of a satellite over the surface of the earth, then the algorithm also gives E , the elevation look angle and r , the slant range or straight-line distance from station 1 to the satellite. If E comes out negative, then the satellite is invisible below the horizon.

On an HP-45 calculator, the longitudes and latitudes can be keyed in the format DD.MMSS (degrees, minutes, and seconds of arc):

```
( $\phi_2$ : DD.MMSS) [G] {D.MS →} 1 [G] {→R}
( $\lambda_1$ : DD.MMSS) [G] {D.MS →}
( $\lambda_2$ : DD.MMSS) [G] {D.MS →} [-] [x ↔ y]
[G] {→R} [x ↔ y] [↑] [R ↓] [R ↓] [→P]
[x ↔ y]
( $\phi_1$ : DD.MMSS) [G] {D.MS →} [-] [x ↔ y]
[G] {→R} [R ↓] [x ↔ y] [R ↓] [→P] [x ↔ y]
(if negative: 360 [+]; see  $A$  in degrees)
[R ↓] [x ↔ y] [→P] [x ↔ y]
```

At this point choose one of the following two options:

- A. 69.1 [x] (see D in miles).
- B. (h , miles) [↑] 3958 [+] [G] {→R} 3958 [-] [x ↔ y] [→P] (see r in miles) [x ↔ y] (see E in degrees).

For an HP-21 calculator, the longitudes and latitudes should first be converted to decimal degrees. Select DEG mode, then:

```
( $\phi_2$ , degrees) [↑] 1 [B] {→R}
( $\lambda_1$ , degrees) [↑]
( $\lambda_2$ , degrees) [-] [x ↔ y] [B] {→R} [x ↔ y] [↑]
[R ↓] [R ↓] [B] {→P} [x ↔ y]
( $\phi_1$ , degrees) [-] [x ↔ y] [B] {→R} [R ↓] [x ↔ y]
[R ↓] [B] {→P} [x ↔ y] (if negative: 360
[+] ; see  $A$  in degrees) [R ↓] [x ↔ y]
[B] {→P} [x ↔ y]
```

By John A. Ball, Oak Hill Road, Harvard, Massachusetts 01451 (Mr. Ball is a radio astronomer at the Center for Astrophysics in Cambridge, Massachusetts)

*These algorithms are based on similar algorithms in my "Algorithms for the HP-45 and HP-35," Center for Astrophysics Technical Report, Cambridge, Massachusetts, March 9, 1975; and in Appendix A.7 of *Algorithms for RPN Calculators*, to be published by John Wiley & Sons, New York.

At this point choose one of the following two options:

- A. 69.1 [x] (see D in miles).
 B. (h , miles) [↑] 3958 [STO] [+] [B] { -R }
 [RCL] [-] [x → y] [B] { -P } (see r in
 miles) [x → y] (see E in degrees).

For a Corvus-500 calculator:

- (ϕ_2 , degrees) [ENT] [SIN] [y → x] [COS]
 (λ_1 , degrees) [ENT]
 (λ_2 , degrees) [-] [y → x] [INV] [G] { -POL }
 [y → x] [ENT] [R↓] [R↓] [G] { -POL }
 [y → x]
 (ϕ_1 , degrees) [-] [y → x] [INV] [G] { -POL }
 [R↓] [y → x] [R↓] [G] { -POL } [y → x] (if
 negative: 360 [+]; see A in degrees) [R↓]
 [y → x] [G] { -POL } [y → x]

At this point choose one of the following two options:

- A. 69.1 [x] (see D in miles).
 B. (h , miles) [ENT] 3958 [+] [INV] [G]
 { -POL } 3958 [-] [y → x] [G] { -POL }
 (see r in miles) [y → x] (see E in degrees).

On some Corvus-500 calculators, [R↓] is just [↓], and [y → x] is [x → y].

If you prefer to work in kilometers (km), change the constant 69.1 miles/° to 111.2 km/° (this is the length of 1° on the earth's surface) and change 3958 miles to 6370 km in two places (this is the radius of the earth). Also change the units of D , h , and r . To work in decimal degrees on the HP-45, replace the first two appearances of [G] {D.MS→} by [↑], and drop the last two [G] {D.MS→}. On an HP-25 calculator, use the HP-45 algorithm but change [G] {D.MS→} to [g] {→H}, [G] {→R} to [f] {→R}, and [→P] to [g] {→P}. This algorithm is approximate because it uses a spherical earth and E is not cor-

note on notation

Keystroke symbols in brackets (e.g., [+]) are printed on the top of the key, those in braces (e.g., { -R }) on the side of the key or on the land area above or below the key. [G] represents an unlabelled gold-colored key, [B] an unlabelled blue key, and [↑] stands for [ENTER]. The symbol :| is analogous to a musical repeat symbol and means loop back to the last preceding colon (:): not in parentheses. Parameters to be keyed or read and comments or optional sequences are in parentheses.

The symbol DD.MMSS means degrees, minutes, and seconds of arc, with two digit locations for each. The decimal point after DD must be keyed. Any digits following SS will be taken for a decimal fraction of a second. Similarly HH.MMSS means hours, minutes, and seconds of time. Use [CHS] for negative numbers. For example -5.420202 would mean -5°42'02.02" with DD.MMSS or -5h42m02s02 with HH.MMSS. For details, see the instruction booklet with the calculator.

rected for refraction (which can be as much as ½°).
Example. $\lambda_1 = 71^\circ 03' = 71.05$, $\phi_1 = 42^\circ 22' = 42.367$ (Boston), $\lambda_2 = 74^\circ$, $\phi_2 = 40^\circ 42' = 40.7$ (New York City); get $A = 234^\circ$ (southwest by west), $D = 191$ miles away. If a satellite is 900 miles directly over New York City, then $r = 925$ miles and $E = 75^\circ$ as seen from Boston. If $\lambda_2 = 70^\circ 40' = 70.667$, $\phi_2 = -33^\circ 25' = -33.417^\circ$ (Santiago de Chile on the west coast of South America); get $A = 179.97$ (slightly east of south) and $D = 5237$ miles from Boston.

As an exercise, compare the distance from Los Angeles to London with the distance from Los Angeles to Rio de Janeiro.

the moon and other celestial objects

This algorithm, slightly modified, also gives pointing angles for the moon. Substitute the local sidereal time T for λ_2 , the moon's right ascension α for λ_1 , and the moon's declination δ for ϕ_2 . Multiply by 15 to change the units of α and T from hours to degrees. For the moon, h should be about 235,000 miles, but an error of less than 1° in E comes from taking $h = \infty$. With these changes, the HP-45 algorithm becomes:

- (δ : DD.MMSS) [G] {D.MS→} 1 [G] {→R}
 (α : HH.MMSS) [G] {D.MS→}
 (T : HH.MMSS) [G] {D.MS→} [-] 15 [x]
 [x → y] [G] {→R} [x → y] [↑] [R↓] [R↓] [→P]
 [x → y]
 (ϕ_1 : DD.MMSS) [G] {D.MS→} [-] [x → y] [G]
 {→R} [R↓] [x → y] [R↓] [→P] [x → y] (if
 negative: 360 [+], see A in degrees) [R↓]
 [→P] [x → y] (see E in degrees).

For an HP-21, α and T should first be converted to decimal hours, and δ and ϕ_1 to decimal degrees, then:

- (δ , degrees) [↑] 1 [B] {→R}
 (α , hours) [↑]
 (T , hours) [-] 15 [x] [x → y] [B] {→R} [x → y]
 [↑] [R↓] [R↓] [B] {→P} [x → y]
 (ϕ_1 , degrees) [-] [x → y] [B] {→R} [R↓] [x → y]
 [R↓] [B] {→P} [x → y] (if negative: 360
 [+]; see A in degrees) [R↓] [B] {→P}
 [x → y] (see E in degrees).

For a Corvus 500:

- (δ , degrees) [ENT] [SIN] [y → x] [COS]
 (α , hours) [ENT]
 (T , hours) [-] 15 [x] [y → x] [INV] [G]
 {→POL} [y → x] [ENT] [R↓] [R↓] [G]
 {→POL} [y → x]
 (ϕ_1 , degrees) [-] [y → x] [INV] [G] {→POL}
 [R↓] [y → x] [R↓] [G] {→POL} [y → x] (if

HP-25 Program

HP-25 Program

SWITCH TO PROGRAM MODE, PRESS [P], THEN KEY IN THE PROGRAM

LINE	CODE	KEY ENTRY	X	Y	Z	T	REMARKS	MEMORY REGISTERS
00	00	GM					in HH.MMSS	R0 GMT, hours
01	15 00	g +H					in hours	
02	23 00	STO 0						
03	15 01	g FRAC					* GMT mod 1	
04	24 02	RCL 2						
05	61						** change in α	
06	24 01	RCL 1					α at GMT	
07	51	+						
08	24 00	RCL 0					α GMT	
09	24 07	RCL 7					δ	
10	51	+					$\alpha + \delta$	
11	24 06	RCL 6					GMT \rightarrow α	
12	61	*					δ in degrees	
13	41	\rightarrow					in degrees	
14	24 00	RCL 0					$\alpha - \delta$	
15	15 01	g FRAC					* GMT mod 1	
16	24 04	RCL 4					$\alpha - \delta$	
17	61	*					δ change in δ	
18	24 03	RCL 3					δ on the hour	
19	51	+					δ at GMT	
20	01	1					unit vector	
21	14 09	f \rightarrow R					$\cos \delta$	
22	21	\rightarrow Y					$\sin \delta$	
23	22	R \rightarrow					$\cos \delta$	
24	14 09	f \rightarrow R					$\sin \delta$	
25	22	R \rightarrow					$\sin \delta$	
26	22	R \rightarrow					$\sin \delta$	
27	22	R \rightarrow					$\sin \delta$	
28	15 09	g \rightarrow P					δ at GMT	
29	21	\rightarrow Y					$\sin \delta$	
30	24 05	RCL 5					ϕ_1	
31	51	+						
32	21	\rightarrow Y					$\sin \delta$	
33	14 09	f \rightarrow R					$\cos \delta$	
34	21	\rightarrow Y					$\sin \delta$	
35	22	R \rightarrow					$\sin \delta$	
36	15 09	g \rightarrow P					δ at GMT	
37	22	R \rightarrow					$\sin \delta$	
38	22	R \rightarrow					$\sin \delta$	
39	22	R \rightarrow					$\sin \delta$	
40	21	\rightarrow Y					$\sin \delta$	
41	15 09	g \rightarrow P					δ at GMT	
42	22	R \rightarrow					$\sin \delta$	
43	13 00	GTO 00						

fig. 1. HP-25 program to calculate antenna pointing angles for the moon or other celestial objects.

negative: 360 [+]; see A in degrees) [R!] [G] { \rightarrow POL } [y \rightarrow x] (see E in degrees).

This is a general pointing algorithm for any distant celestial object. The α and δ of the moon and other objects are in *The American Ephemeris and Nautical Almanac* or *AENA*.¹

sidereal time

If your shack is not equipped with a sidereal clock, the *AENA* shows how to calculate T , or on an HP-45:

```
(Day-Number) [!] 24 [x] [!] [!]
(GMT: HH.MMSS) [G] {D.MS  $\rightarrow$ } [ + ]
1.0027379 [x] [x  $\rightarrow$  y] [ - ]
( $\lambda_1$ : DD.MMSS) [G] {D.MS  $\rightarrow$ } 15 [ + ] [ - ]
(S, hours) [ + ] (if negative: 24 [ + ], if greater than 24: 24 [ - ]) [G] {  $\rightarrow$ D.MS } (see  $T$  in HH.MMSS).
```

Day-Number means day of the year, GMT is Greenwich mean time or universal time (UT), and S is from the table below.

Year, AD	S, hours
1976	6.5865
1977	6.6363
1978	6.6204
1979	6.6044
1980	6.5885
1981	6.6383
1982	6.6224
1983	6.6065
1984	6.5906

SWITCH TO PROGRAM MODE, PRESS [P], THEN KEY IN THE PROGRAM

LINE	CODE	KEY ENTRY	X	Y	Z	T	REMARKS	MEMORY REGISTERS
00	00	GM					in minutes	R0 δ , degrees
01	23 00	STO 7					save 1	
02	04	4						
03	23 71 02	STO \rightarrow 7					$1/\delta$ to R7	R1 $\lambda_1 - \lambda_1 + 90$, degrees
04	24 04	RCL 4						
05	14 21	f \rightarrow R					$360/\rho$	
06	14 04	f \rightarrow R					$180-1$	
07	14 09	f \rightarrow R					$\sin \delta$	
08	14 21	f \rightarrow R					$360/\rho$	
09	14 05	f \rightarrow R					$\cos \delta$	
10	15 09	g \rightarrow P					δ at GMT	
11	21	\rightarrow Y					$\Delta \lambda - 1/4$	
12	24 07	RCL 7					$\Delta \lambda - 1/4$	
13	51	+					$\cos \delta$	
14	24 01	RCL 1					$\Delta \lambda$	
15	51	+					$\cos \delta$	
16	21	\rightarrow Y					$\cos \delta$	
17	14 09	f \rightarrow R					$\sin \delta$	
18	31	f \rightarrow R					$\sin \delta$	
19	22	R \rightarrow					$\sin \delta$	
20	22	R \rightarrow					$\sin \delta$	
21	15 09	g \rightarrow P					δ at GMT	
22	21	\rightarrow Y					$\sin \delta$	
23	24 05	RCL 5					ϕ_1	
24	41	\rightarrow						
25	21	\rightarrow Y					$\sin \delta$	
26	14 09	f \rightarrow R					$\cos \delta$	
27	22	R \rightarrow					$\sin \delta$	
28	21	\rightarrow Y					$\sin \delta$	
29	22	R \rightarrow					$\sin \delta$	
30	15 09	g \rightarrow P					δ at GMT	
31	21	\rightarrow Y					$\sin \delta$	
32	23 00	STO 0						
33	22	R \rightarrow					$\sin \delta$	
34	21	\rightarrow Y					$\sin \delta$	
35	15 09	g \rightarrow P					δ at GMT	
36	21	\rightarrow Y					$\sin \delta$	
37	24 06	RCL 6					3956+h	
38	14 09	f \rightarrow R					$\sin \delta$	
39	24 05	RCL 5					395h	
40	41	\rightarrow						
41	21	\rightarrow Y					$\sin \delta$	
42	15 09	g \rightarrow P					δ at GMT	
43	24 00	RCL 0					α	
44	13 00	GTO 00						

fig. 2. HP-25 program to calculate antenna pointing angles and range for earth satellites.

S is the Greenwich sidereal time on January 0.0 of the indicated year, and is in the *AENA*. For an HP-21, key GMT in decimal hours and delete the following: [G] {D.MS \rightarrow }; key λ_1 in decimal degrees and replace the following: [G] {D.MS \rightarrow } by [!]; and delete the final [G] { \rightarrow D.MS } and see T in decimal hours. For a Corvus 500, follow the HP-21 scheme but replace [!] by [ENT] and [x \rightarrow y] by [y \rightarrow x]. The precision of this algorithm is about ± 1 second. A sidereal clock runs one day per year or 3^m57^s per day faster than an ordinary clock.

Example. On 1976 December 25 (Day-Number = 360) at GMT = 21h, the moon was at $\alpha = 22^h25^m37^s564$, $\delta = -5^{\circ}21'33''33$ (from the 1976 *AENA*, page 187). The local sidereal time at Boston was $T = 22^h33^m46^s$, and the moon was visible at $A = 182.07$ (almost south), and approximately $E = 42.0$ (just under halfway from horizon to zenith).

The program in fig. 1 combines the celestial pointing and sidereal-time algorithms for the HP-25.

Key in the program, then initialize:

```
( $\alpha$ : HH.MMSS) [g] {  $\rightarrow$ H } 15 [x] [STO] 1
( $\Delta \alpha$ , seconds/hour) [!] 240 [ + ] [STO] 2
( $\delta$ : DD.MMSS) [g] {  $\rightarrow$ H } [STO] 3
( $\Delta \delta$ , arcseconds/hour) [!] 3600 [ + ] [STO] 4
( $\phi_1$ : DD.MMSS) [g] {  $\rightarrow$ H } [STO] 5
(S, hours) [!] 15 [x]
( $\lambda_1$ : DD.MMSS) [g] {  $\rightarrow$ H } [ - ] 15.04106864 [STO] 6 [ + ]
```

(Day-Number) [1] 24 [x] [+] [STO] 7 [f] {PRGM}
 Calculate: (GMT: HH.MMSS) [R/S] (see *E* in
 degrees) [x → y] (see *A* in degrees) :|.

If *A* is negative, you may want to add 360°. For this program, α is the right ascension on the hour, $\Delta\alpha$ is the change in α in one hour, δ is the declination on the hour, and $\Delta\delta$ is the change in δ in one hour. Don't forget minus signs [CHS] where needed. The rates $\Delta\alpha$ and $\Delta\delta$ are called first differences and are tabulated for the moon in the *AENA*.

This program interpolates linearly through the hour; α and δ should be for the preceding hour. For a planet such as Jupiter, tabulated daily rather than hourly in the *AENA*, change the [g] {FRAC} in lines 03 and 15 both to [g] {NOP} and key α and δ for 0h. The tabulated first differences in this case need to be divided by 24 to get the right units (per hour rather than per day). Jupiter is an interesting object for amateur radio astronomy. For a star with constant α and δ , set $\Delta\alpha$ and $\Delta\delta$ to zero.

Example. Repeat the preceding test case on the moon but for GMT = 21h20m. From the 1976 *AENA*, $\Delta\alpha = 122^{\circ}767$ and $\Delta\delta = 615''44$. Get $A = 189.2^{\circ}$ ($A = -170.78 + 360 = 189.2^{\circ}$ and $E = 41.9^{\circ}$).

satellites

Given q , a satellite's period in minutes of time; i , the inclination in degrees of the satellite's orbit to the equator; τ , the time since an ascending node (northbound equator crossing) in minutes; and λ_0 , the longitude of the ascending node ($\tau = 0$) in degrees;* the HP-45 algorithm below^{2,3} calculates λ_2 and ϕ_2 for the satellite:

```
(τ, minutes) [1] [1] (q, minutes) [÷] 360 [x]
[1] [SIN]
(i, degrees) [x → y] [G] { -R } [x → y] [G]
{ SIN-1 } (see φ2 in degrees)
[R ↓] [x → y] [COS] [ -P ] [R ↓] [x → y] 4 [÷]
[ - ]
(λ0, degrees) [x → y] [ - ] (see λ2 in degrees).
```

This algorithm is approximate because it uses a circular orbit; h is taken to be constant.

For an HP-21, select DEG mode, change [G] to [B] and change [-P] to [B] { -P }. For an HP-25, change [G] to [g] or [f], [SIN] to [f] { sin }, [COS] to [f] { cos }, and [-P] to [g] { -P }. For a Corvus 500, change [1] to [ENT], [x → y] to [y → x], [G] { SIN⁻¹ }

*The orbital elements of Oscar satellites are available from AMSAT, Post Office Box 27, Washington, D.C. 20044. For Oscar 7: $q = 114.95$ minutes, $i = 101.7$ degrees, $h = 908$ miles, $e = 0.001$; but these numbers change slowly with time. Also see the *Satellable* available from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048. For other satellites, see *Satellite News*, 12 Barn Croft, Preston PR1 OSX, England.

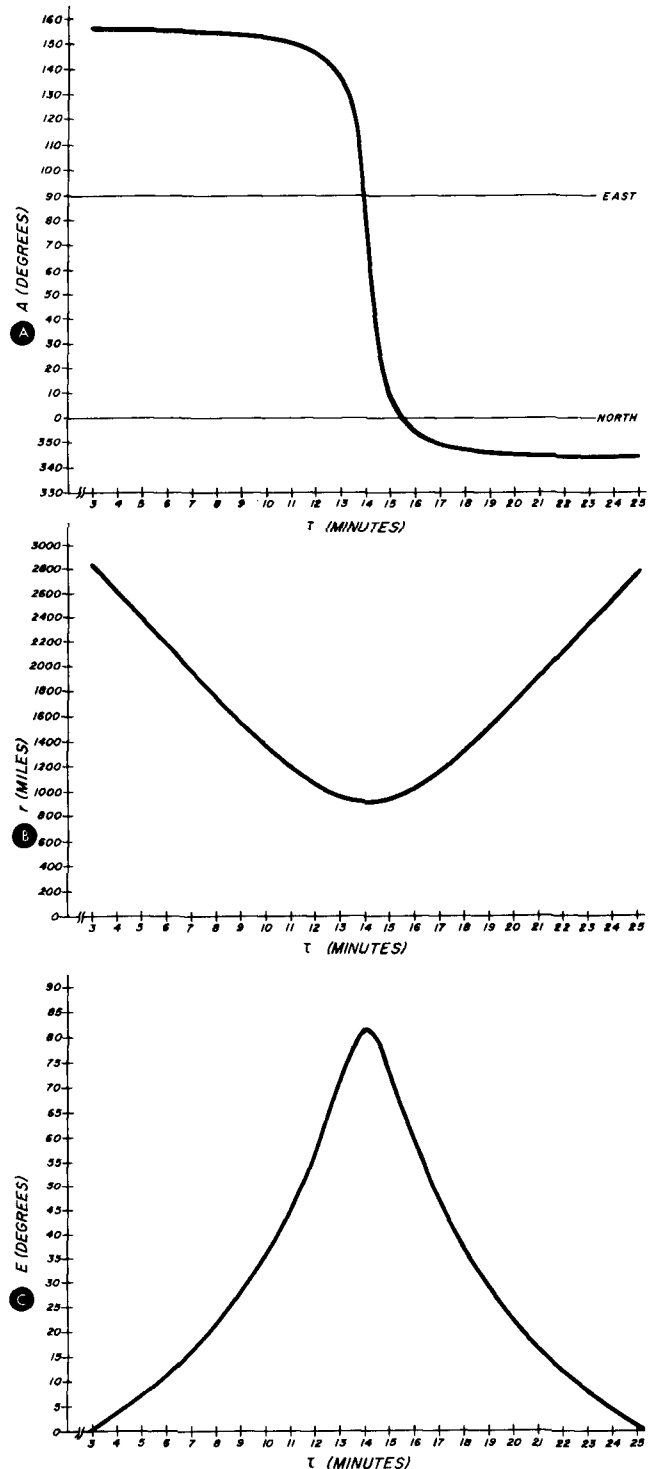


fig. 3. Azimuth, range, and elevation for Oscar 7 orbit 10590 as seen from Boston.

to [INV] [SIN], [-P] to [G] { -POL }, and [G] { -R } to [INV] [G] { -POL }. These algorithms are also easy to convert for the HP-19, HP-27, HP-29, HP-46, HP-55, HP-65, HP-67, HP-91, HP-97, HP-9815,

summary of symbols

Station 1 is normally home; station 2 is sometimes a subsatellite point.

A	heading (north reference clockwise azimuth) from station 1 toward station 2 or toward a celestial object. $A = 0$ is north, $A = 90^\circ$ is east, $A = 180^\circ$ is south, and $A = 270^\circ$ is west. Negative A means counterclockwise, so $A = -90^\circ$ is also west.	S	Greenwich sidereal time on January 0.0 (tabulated in the article).
c	speed of light; $c = 11,177,000$ miles/minute.	T	local sidereal time at station 1.
D	great-circle distance from station 1 to station 2.	v	Doppler velocity of a satellite relative to station 1.
Day-Number	day of the year (GMT) available on some desk calendars.	α	right ascension of a celestial object.
E	elevation look angle from station 1 toward a satellite or celestial object. $E = 90^\circ$ is straight up toward the zenith, $E = 0$ is the horizon, and negative E means that the object is invisible below the horizon.	δ	declination of a celestial object.
e	eccentricity of a satellite's orbit; $e = 0$ is a circle.	Δf	change in f due to a Doppler shift.
f	frequency.	$\Delta\alpha$	change in α per unit time (first difference).
GMT	Greenwich mean time or universal time (UT or UTC). Ephemeris time differs from GMT by about 48 seconds.	$\Delta\delta$	change in δ per unit time (first difference).
h	height of a satellite over the surface of the earth.	λ_0	longitude of the ascending node at $\tau = 0$ in a satellite's orbit. West longitudes are positive (+), east longitudes are either negative (-) or greater than 180° .
h_a	the average h over an orbit.	λ_1	longitude of station 1.
i	inclination angle of a satellite's orbit to the equator.	λ_2	longitude of station 2 or a subsatellite point.
r	slant range or straight-line distance from station 1 to a satellite.	π	3.1415926536.
		ϱ	period (time for one complete orbit) of a satellite.
		τ	time since an ascending node (northbound equator crossing) in a satellite's orbit.
		ϕ_1	latitude of station 1. North latitudes are positive (+), south latitudes are negative (-).
		ϕ_2	latitude of station 2 or a subsatellite point.

National Semiconductor 4640, APF Mark 55, and Omron 12-SR calculators. Converting for non-RPN calculators is more difficult.

The program shown in **fig. 3** combines the satellite and pointing algorithms for the HP-25. Key the program, then initialize:

```
( $\lambda_0$ , degrees) [1] ( $\lambda_1$ , degrees) [-] 90 [+] [STO] 1
( $\phi_1$ , degrees) [STO] 2 ( $\varrho$ , minutes) [1] 1440 [+ ]
[STO] 3 180 [1] ( $i$ , degrees) [-] [STO] 4 3958 [STO]
5 ( $h$ , miles) [+] [STO] 6 [f] {PRGM}
Calculate: ( $\tau$ , minutes) [R/S] (see  $A$  in degrees)
[R↓] (see  $r$  in miles) [R↓] (see  $E$  in
degrees) :|.
```

If A is negative, you may want to add 360° . The constant 1440 is the number of minutes in a day.

Example. Oscar-7 orbit 10590 on March 10, 1977, $\tau = 14^m$ (which corresponds to GMT = $00^h08^m + 14^m = 00^h22^m$), $\varrho = 114^m945$, $i = 101.7^\circ$, $\lambda_0 = 54.95^\circ$, $h = 908$ miles; get $\phi_2 = 42.971^\circ$, $\lambda_2 = 69.902^\circ$, $A = 76^\circ$, $r = 916$ miles, and $E = 82^\circ$, almost overhead as seen from Boston.

Fig. 2 shows A , r , and E as a function of τ for this orbit over Boston. For a pass almost overhead such as this, A is nearly constant except for a couple minutes around closest approach. The slope v of the r curve is the Doppler velocity (in miles per minute) and can be calculated by subtracting two y s a minute apart. The largest possible v is $2\pi(h + 3958 \text{ miles})/\varrho$ or about 266 miles/minute for Oscar 7. Convert v to a Doppler frequency shift using $\Delta f = -fv/c$, where f is the frequency, Δf is the shift in frequency, and c is the speed of light ($c = 11,177,000$ miles/minute). The Oscar-7 beacon at 145.975 MHz, for example, shifts as much as 0.00347 MHz and so appears somewhere between 145.972 and 145.978 MHz.

When the satellite is approaching, v is negative, Δf

is positive, and the beacon appears at a higher frequency; when the satellite is receding, v is positive, Δf is negative, and the beacon appears at a lower frequency on the dial. If you are working through a repeater on a satellite, then two different Doppler shifts need to be calculated and added.

Kepler's third law relates ϱ and h_a , the average h , (ϱ , minutes) [1] [x] 8720351 [x] 3 [1/x] [G] { y^x } 3958 [-] (see h_a in miles).

The first number is Kepler's constant for the earth, 8,720,351 miles³/minute². With a circular orbit, h_a and h are the same. For Oscar 7 with $\varrho = 114^m945$, get $h_a = 908$ miles.

accuracy

These algorithms employ several approximations that cause inaccuracies in the answers. The earth's eccentricity causes an error up to 0.2 per cent (2 miles in 1000 miles) in D . The eccentricity of a satellite orbit has two effects: first, h is not constant. The maximum variation in h is $e(3958 \text{ miles} + h_a)$ where e is the eccentricity and h_a is the average h . So if $e = 0.01$ and $h_a = 925$ miles, then the variation in h is 49 miles and h varies from 876 to 974 miles.

Another effect of e is on the speed of the satellite, which can be as much as about $e\varrho/\pi$ ahead or behind in the orbit compared to a circular orbit. If $e = 0.01$ and $\varrho = 115$ minutes, then the satellite can be up to 0.37 minutes early or late. Also the "4" in the orbital algorithm, which converts τ from minutes of time to degrees, should be $4/1.0027379 = 3.9890783$ because the earth's rotation speed is one degree per 4 *sidereal* minutes. The error due to using 4 is noticeable only for large τ .

Finally, the orbital elements of earth satellites change slowly due to influences such as the sun and

moon, air resistance, and the uneven distribution of mass in the earth. For Oscar 7, the orbit regresses just about enough to cancel the error caused by using 4 minutes per degree; such orbits are called sun-synchronous.

I wish to thank George Rybicki for showing me the spherical-trigonometry trick used in these algorithms; Dick Ellis, W5YCK, for helping research articles on Oscar satellites; Tom Bates and Fritz Mans-

velt-Beck for loaning me their calculators; and R.C. Vanderburgh for sending me copies of his programs.

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2. P. D. Thompson, Jr., "A General Technique for Satellite Tracking," *QST*, November, 1975, page 29.
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appendix RPN portable calculators

The table below lists all the available or recently available RPN portable calculators. The algorithms in this article can be easily converted to work on any calculator that has "yes" in the R → P column in this table. Non-RPN calculators are not listed; they are awkward for this type of problem.

The Corvus 500, the APF Mark 55, and the Omron 12-SR are internally identical, but the Corvus 500 has a better keyboard and case. The instruction booklet with the Corvus 500 is very poor; if you have this calculator, you should also get the book *Everything You've Always Wanted to Know About RPN but were Afraid to Pursue — Comprehensive Manual for Scientific Calculators* available for \$7.50 plus postage from T. K. Enterprises, 16611 Hawthorne Boulevard, Lawndale, California 90260. This book does not live up to its name and is not as good as the instruction books with HP calculators, but is the best book available for the Corvus 500, APF Mark 55, and Omron 12-SR.

For an HP calculator, you should get the appropriate HP applications book. Some dealers will throw these in with the calculator; otherwise they are about \$10 from Hewlett-Packard, 19310 Pruneridge Avenue, Cupertino, California 95014. Some HP calculators have several such books.

Generally speaking, HP calculators are better made and will probably last longer than the others in this table. Don't overlook the possibility of finding a used or surplus calculator of a model listed as no longer available.

"Start extravagant, and you'll never finish. Get the cheap tool first, see if it feeds your life. If it does, then get a better one. Once you use it all the time, get the best. You can only grow into quality. You can't buy it."

— from *The Last Whole Earth Catalog*

Table 1. List of portable calculators which use RPN architecture.

manufacturer	model	stack of	scientific notation	number of storage registers	number of program steps	R → P	D.MS
Hewlett-Packard	HP-19C	4	yes	30	98	yes	yes
Hewlett-Packard	HP-21	4	yes	1	0	yes	no
Hewlett-Packard	HP-22	4	yes	10 + 5	0	no	no
Hewlett-Packard	HP-25	4	yes	8	49	yes	yes
Hewlett-Packard	HP-25C	4	yes	8	49	yes	yes
Hewlett-Packard	HP-27	4	yes	10 + 5	0	yes	yes
Hewlett-Packard	HP-29C	4	yes	30	98	yes	yes
Hewlett-Packard	HP-35*	4	yes	1	0	no	no
Hewlett-Packard	HP-45*	4	yes	9	0	yes	yes
Hewlett-Packard	HP-55*	4	yes	20	49	yes	yes
Hewlett-Packard	HP-65*	4	yes	9	100	yes	yes
Hewlett-Packard	HP-67	4	yes	26	224	yes	yes
Hewlett-Packard	HP-70*	4	yes	2 + 5	0	no	no
Hewlett-Packard	HP-80	4	yes	1	0	no	no
Hewlett-Packard	HP-91	4	yes	16	0	yes	yes
Hewlett-Packard	HP-92	4	yes	30	0	no	no
Hewlett-Packard	HP-97	4	yes	26	224	yes	yes
National Semiconductor	Novus 4510	3	no	1	0	no	no
National Semiconductor	Novus PR4515/4615	3	no	1	100	no	no
National Semiconductor	Novus 4520	4	yes	1	0	no	no
National Semiconductor	Novus PR4525	4	yes	1	100	no	no
National Semiconductor	NS 4640	4	yes	3	0	yes	yes
Corvus	500	4	yes	9	0	yes	no
APF	Mark 55	4	yes	9	0	yes	no
Omron	12-SR	4	yes	9	0	yes	no

*No longer available

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Simple circuit for
improving your counter
from dc to
over 60 MHz using
readily available devices

Many of the inexpensive frequency counters on today's market, as well as some of those built by homebrew enthusiasts, could use some improvement in the preamp and pulse-shaper circuit (sometimes called the trigger circuit). This is the circuit that brings the input signal waveform to TTL level (3.5 volts peak-to-peak). Often counters potentially capable of counting to 40-50 MHz don't produce best results because of an inefficient input circuit. On the other hand, if the trigger works fine at high frequencies, it shows some limitation in squaring low-frequency signals. Kritter¹ solved this problem by using two pulse shapers in his frequency counter. Although this approach is quite satisfactory, it can be impractical and expensive.

The circuit we're going to examine can process any frequency between dc to over 60 MHz, providing

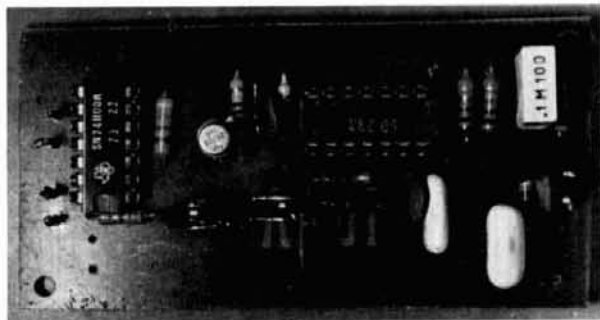
at the output a perfect square wave at TTL level. Furthermore, it's not very expensive, because it uses transistors and ICs available at most discount outlets.

circuit description

The input circuit (fig. 1) is a balanced amplifier using two field-effect transistors. For dc stability, the stages are dc coupled. These fets must be selected for the same ID_{SS} .² The absolute ID_{SS} value isn't critical but must be nearly equal for the two devices.

The first stages are source followers, so the input impedance is extremely high. At low-input signal levels, when the two back-to-back protective diodes don't conduct, overload protection is provided only by the 2.2-meg polarization resistor. It would be possible to increase the input impedance even more by increasing this resistor value or by using bootstrap polarization. However, I felt that this value was more than adequate for amateur purposes.

The input stage drives a 733 IC which, according to reference 3, is a differential video amplifier with a



Top view of the counter preamp PC board showing component layout.

bandwidth of over 100 MHz. The gain of this amplifier is selectable by proper connection of pins 3, 4, 11, and 12 (the DIP package).

Three different gain values are possible without adding external components: x10, x100, and x400. The last has been chosen for this application. Even if, in this case, the amplifier bandwidth is reduced

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I-41012, Carpi, Italy

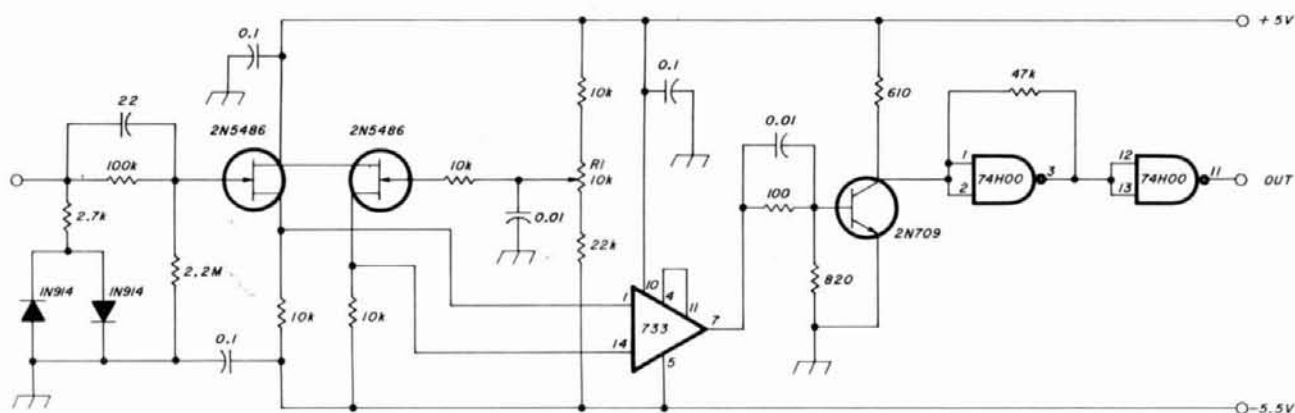


fig. 1. Input circuit for improving inexpensive counters. Circuit requires a dual-polarity supply that delivers at least 63 mA.

slightly, it remains well over 50 MHz — more than sufficient for our use.

A 2N709 switching transistor follows the preamplifier and squares the signal. If you want to reach the maximum frequency that the pulse shaper can handle, this transistor is highly recommended. It's fast. Some attempts that have been made to use other types of transistors (for example, 2N914, 2N2368) for processing 60-MHz signals haven't been very satisfactory. So I highly recommend the 2N709 if you don't want problems. Of course, I haven't tried all the available switching transistors, so there are many substitution possibilities if you like to experiment. If you have some computer transistors, try them; you might obtain even better results.

The last stage is quite conventional. It is a TTL translator using two sections of a high-speed quadruple NAND gate (type 74H00). For this application I've tried several ICs made by different manufacturers. Those that gave the best results were a Texas Instruments SN74H00 and a National Semiconductor DM74H00. I also tried some Schottky devices (74S00). All performed well, even at higher frequencies.

The circuit requires a supply that delivers plus and minus 5 volts. Without an input signal, the input transistor drain current is 18 mA from the negative

side and 45 mA from the positive side, which increases to 63 mA with a strong input signal. Tests showed that the circuit was not too sensitive to an unregulated supply, so the filtering doesn't have to be elaborate. Of course devices such as the 320- and 340-series voltage regulator ICs can be used to solve a power-supply problem at a reasonable price. The only recommendation is to avoid any possibility of false counting by bypassing the positive supply as closely as possible to pin 14 of the TTL IC.

construction and alignment

The prototype was constructed on a 3 x 1.5-inch (8.2x4cm) board (fig. 2). The component layout shown in fig. 2B is slightly different from that in the photo because of some improvements made after the photo was taken. The second trimmer resistor on the PC board was a former polarization control for the level translator, which was disconnected after some tests.

The only alignment required for the circuit is the regulation of the trimmer pot, R1. With the counter connected at the output and a signal source (grid-dip meter or signal generator) at the input, adjust R1 for maximum sensitivity at the highest measurable frequency. With proper alignment, the input sensitivity must be better than 100 mV in the range from a few

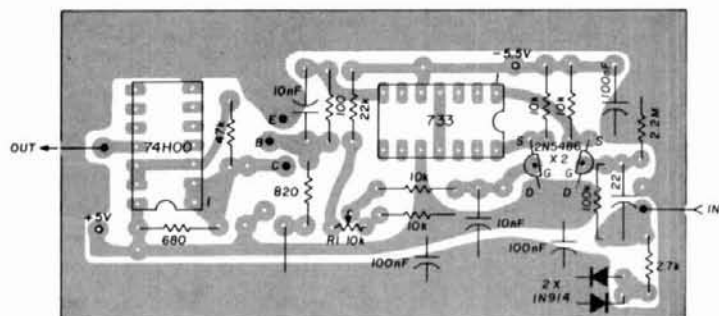
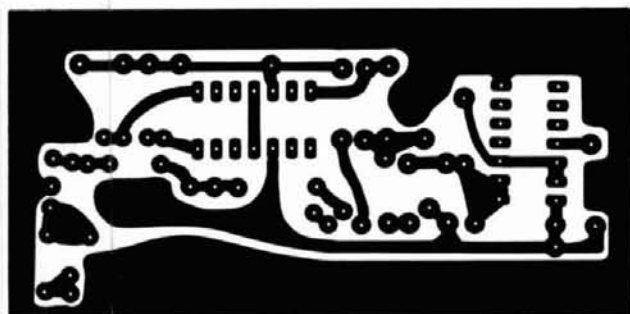


fig. 2. Circuit-board layout and component arrangement for the counter preamp.

MHz to at least 40 MHz, decreasing to about 150 mV at the lower (DC) and upper limit (66 MHz in the prototype).

counter improvements

Some counters using 7400-series ICs may not cover frequencies higher than 30 MHz. It's now time to update them using faster ICs provided by modern technology. Substitute the input gate of the counter (it may be a 7400 or a 7410, perhaps a 7420) with the Schottky TTL equivalent: 74S00, 74S10, or 74S20. The substitution was direct and required no wiring change. Then substitute the first decade divider (7490) with the faster type 74196 or 74S196. Here the connections are different, so you must have a little patience and, following the schematic of fig. 3, make the modifications needed on the PC board. The 74196 IC requires a reset pulse inverted with respect to that required by the 7490. If you have an unused NAND or inverter circuit in one of the other ICs, use it to invert the reset pulse; otherwise use one of the two NAND gates not used in the pulse-shaper circuit. With these simple modifications, your frequency counter can now display at least 50 or 60 MHz. Don't forget that some ICs are now available that permit scalars for GHz frequencies. With a pair of these ICs

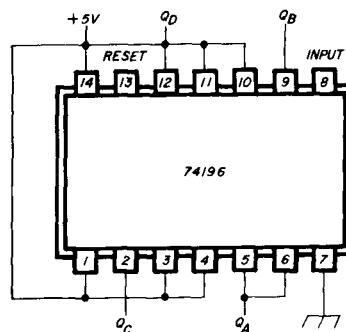


fig. 3. Pin layout for the 74196 decade divider, which may be used in your counter for faster response. Modifications shown must be made on the PC board.

and your improved counter, it will be possible to make precise measurements well into the microwave region.

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features

With the instrument described you can apply a polarized voltage of a few volts to the capacitor under test, one side of which is grounded. A single unregulated 9-volt power supply is used. Capacitance readout is linear.

Three testing modes are available: low capacitance (to 1 μF), high capacitance (to 2500 μF), and a logarithmic indication of the test-capacitor leakage current with up to 8 volts applied. Let's take a look at the circuit (**fig. 1**).

low-capacitance measurement mode (to 1 μF)

U1, an NE555, operates as a clock, which provides negative-going pulses at about 350 per second to

trigger U2, also an NE555. This action unclamps the test capacitor, allowing it to charge through a switch-selected resistor, until it reaches half the supply voltage. At this point U2 resets, discharging the capacitor through pin 7. During the charging period, U2 pin 3 is high (about 8 volts), and the duration of this high state is directly proportional to the test capacitance.

The resulting rectangular waveform can be used to drive a 1-mA meter directly through a 5k trimpot for a simplified circuit. In this instrument, the high signal is attenuated to 0.6 volt across silicon diode CR1 at the noninverting input of U3, a CA3140 op amp with mos input. U3 operates as a unity-gain buffer, which feeds the meter through calibrating trimpot R6. Meter deflection is proportional to the average value of the rectangular waveform output from U3 and is therefore proportional to the capacitance.

Supply voltage is noncritical because:

1. U1 clock frequency is, for practical purposes, independent of voltage.
2. The reset level of U2 pin 6 is at one-half the supply voltage, which compensates for voltage-change effects on the charge rate of the test capacitor through the switched resistors.
3. CR1 operates as a simple regulator, limiting the high signal input to U3 to 0.6 volt.

With no test capacitor applied to the circuit, about 30 pF of internal capacitance exists at U2 pins 6 and 7 (plus strays). To prevent this capacitance from causing a substantial residual reading on the lower ranges, a *negative* capacitance is used to cancel the internal capacitance. This unlikely device is simulated by C3, a 100-pF capacitor, which is connected not to

By P. H. Mathieson, M.I.E. Australia, JAARS, Box 248, Waxhaw, North Carolina 28173

ground but to Q1/Q2 output, which has a small positive voltage gain and a low output impedance. The current through C2 is equal and opposite to that through the residual 30-pF capacitance.

To calibrate, use a known accurate capacitor giving a high deflection on the 0.01 or 0.1 μF ranges. First adjust negative-capacitance trimpot R1 so that the source of Q1 is at the top end of R1. Then calibrate by adjusting R6. There may be a small residual zero error caused by the minimum pulse width from U2, which can be offset by the addition of R7, a 470 k in my instrument. Now switch to the 100-pF range with a known capacitor of 10-20 pF connected and adjust negative cap trimpot R1 until the meter reads the correct value, showing that the strays have been cancelled. The readings are then accurate to a few pF.

high-capacitance measurement mode (to 2500 μF)

This mode uses a single-shot method. U1 is not required. When +9 volts is switched on, U2 is triggered by the momentary low on pins 2 and 4 caused by the uncharged capacitor, C2. As in the previous mode, a high of 0.6 volts is applied to the noninverting input of U3 until the capacitor under test is charged to half the supply voltage. During this period, U3 behaves as an accurate integrator using low-leakage capacitor C4 in the feedback loop. At the end of the high input period, U3 output voltage will be proportional to the duration of that period and therefore to the test capacitance.

Accurate high-value capacitors are difficult to find, so calibration using trimpot R5 is best done with values around 1 μF . Leakage resistance of the test capacitor extends the charging time, causing a false high reading.

After the integrating period the meter should remain stationary while the reading is taken. If drift is a problem (assuming feedback capacitor C4 is not leaky) it may be minimized by correcting the offset in the U3 input stage. Try a 10k resistor R3 from U3 pin 1 or 5 to ground and adjust for minimum drift.

leakage mode

This mode produces a logarithmic indication of the test capacitor leakage current with up to 8 volts applied. The lower end of the capacitor is disconnected from the supply minus, and the leakage current now flows through limiting resistor R8 and diode CR1. The voltage across CR1 bears an approximately logarithmic relationship to the current flowing through it. U3 is again used as a unity-gain buffer, and trimpot R4 is set to produce full-scale deflection with a short circuit across the test terminals. U1, U2, and the negative capacitance amplifier are disconnected from the negative supply line.

Electrolytic capacitors that have remained unused for some time can be reformed in the leakage test mode before their capacitance is measured.

The logarithmic readout can be interpreted by observing the readings obtained with known resistors across the test terminals, ranging from a few kilohms to hundreds of megohms.

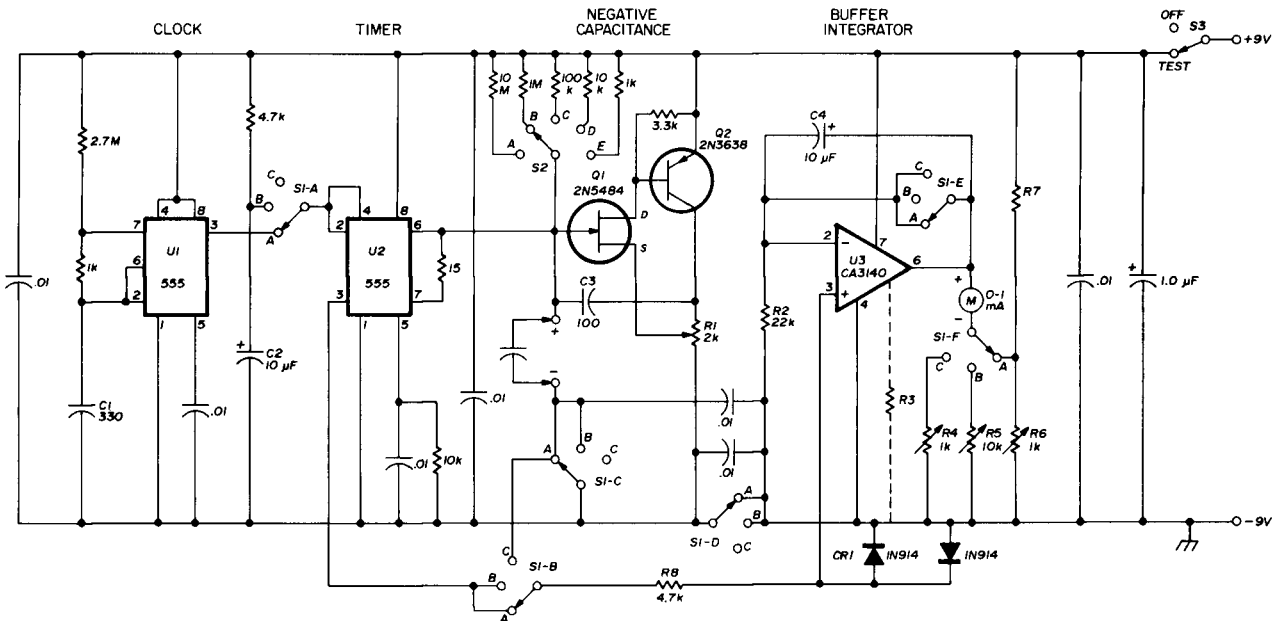


fig. 1. The three-mode capacitor tester. Circuit features linear readout and common devices. Author's original was built on perf board but with a little ingenuity an etched board could be used. See table 1 for switch configurations and positions.

table 1. Switch arrangement for the portable capacitance meter.

switch	configuration	position		
		mode A	mode B	mode C
S1	3-pole, 3-position, 2 section	A (low-capacitance mode)		
		B (high-capacitance mode)		
		C (leakage mode)		
		μF	μF	
S2	1-pole, 5-position	A 0.0001	0.25	
		B 0.001	2.5	
		C 0.01	25.0	
		D 0.1	250.0	
		E 1.0	2500.0	leakage
S3	SPST (test)			

The capacitance ranges are in steps of 1:10. Intermediate ranges are most economically obtained as follows. In mode A, (table 1) switching to a larger clock timing capacitor, C1, will permit larger capacitances to be read on scale. In mode B, switch resistor R2 to a higher value, which slows the integrating rate and thus allows a greater full-scale capacitance reading. Ten-thousand μF is probably a practical limit because of leakage.

Use accurately scaled values for the switched range resistors connected to S2 for best accuracy.

construction notes

The CA3140 op amp was chosen because of its very high input impedance and because both inputs and the output can be swung down to the negative supply line, eliminating the need for a separate negative supply. The positive supply may be varied from +6 to +12 volts with little effect on calibration. Current drain is 20-30 mA for capacitance measurements and a few mA for leakage measurements.

The original circuit was built on matrix board with the layout approximately following this schematic. Because fast rise times are involved, several 0.01 μF bypass capacitors are included on the board, and leads should be kept short and neat. The op amp is zener protected — keep a shorting ring around the pins while soldering! If a shorting-type wafer switch is used for S1, insert a 470-ohm current-limiting resistor in series with U2 pin 3.

The test points can be alligator clips on short flexible leads fed through grommets in the front panel. Take care to minimize stray capacitance to ground from the plus test point.

Finally, never apply reversed power-supply voltage to the circuit unless you want to buy three new ICs. A reverse-biased 1-amp diode across the 9-volt supply will provide protection.

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solid-state vhf-uhf transmit/receive switch

New PIN diode device
provides good isolation
and low vswr
at frequencies
up to 1000 MHz

Microwave Associates recently introduced a solid-state T/R switch for vhf-uhf applications. The device, designated the MA8334, makes use of PIN diodes in a hybrid rf circuit which is small and easy to use. This spdt switch is rated at 50 watts CW and has a nominal 50-ohm impedance. Frequency of operation is from 20 to 1000 MHz. Specifications list typical insertion loss at 0.2 dB with 1.2:1 vswr from 20 to 500 MHz.

After evaluating the MA8334 on the test bench, I decided to replace the conventional relays in a solid-state 2-meter transverter I had recently built. The circuit of **fig. 1** was used. To operate a switch path, approximately 50 mA forward bias is applied; removal of the bias releases the switch path. Capacitors C1 and C2 provide isolation of the dc bias and the rf source feeding the switch. These capacitors should have low rf loss and be able to handle the power used. I used button micas in my design. The inductance of the rf choke is not critical, and anything around 3 μ H should work satisfactorily. Capacitors C3 and C4 are feedthrough bypass types. The 220-ohm resistors provide the correct bias current when using a 12-volt power supply; their values

will have to be adjusted accordingly if another supply voltage is used.

Operation of the MA8334 has been completely satisfactory. With the circuit of **fig. 1**, the measured insertion loss was 0.25 dB, and the swr is 1.23:1 when operated at 50 ohms. Isolation between ports has not been measured, but from observation I would judge it to be at least the 37 dB specified by the manufacturer.

To conclude, this device provides interesting possibilities for vhf-uhf switching at power levels up to

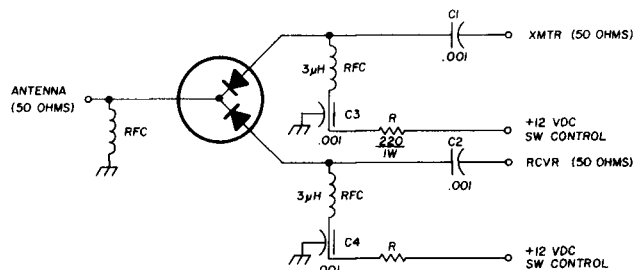


fig. 1. Circuit diagram of a solid-state T/R switch for 144 MHz. The MA8334 can be used at other frequencies up to 1000 MHz by proper choice of circuit constants. The lead marked with an asterisk is identified by a diagonally cut end; the other two leads may be used interchangeably for either transmitter or receiver.

50 watts CW. With proper component selection, the switch should perform well on 432 MHz. When compared with conventional switches, the MA8334's compact size, rugged construction, and reliability give it a definite advantage. The present single-quantity cost is \$19.00, but with increased production the price is expected to drop significantly. It may be purchased through any dealer handling Microwave Associates products.

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digital scanner for 2-meter synthesizers

Complete
construction details
plus other unusual ideas
for integrating
a digital scanner
with a 2-meter synthesizer

After operating a synthesizer, you begin to realize the number of 2-meter repeater channels that exist and also how long it takes to turn the switches through all channels. The digital scanner presented in this article was designed to permit easy, hands-off, monitoring of the 2-meter fm band. It can also serve as a good indicator of 2-meter band conditions by listening for repeaters outside your local area.

The features incorporated in the scanner were based on several months of on-the-air operation of a prototype in an area heavily populated with repeaters. It scans all 2-meter repeater input and output frequencies between 146.01 MHz and 147.99 MHz, in 30 kHz steps; all 67 frequencies are scanned in about 6 to 8 seconds. The frequency is also read out directly by five, 7-segment LED displays.

Operating features include three modes of scanning, A, B, and C. Mode A scans until a signal is received, at which time the scanner stops, listens for about 3 seconds, and then continues. This mode allows rapid scanning of all channels to determine activity. Mode B scans until a signal is received and waits until the signal is gone before continuing the scan. Mode C is the same as Mode B except that a 3-second delay oc-

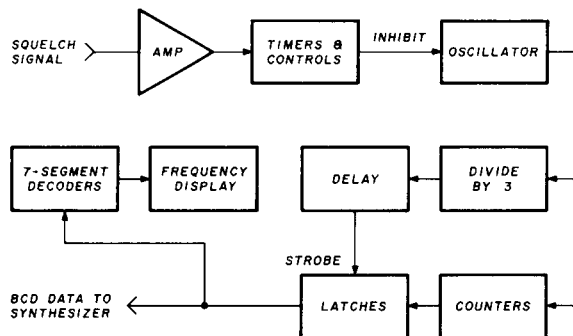


fig. 1. Functional block diagram of the digital scanner.

curs before continuing the scan. Mode C allows monitoring of repeaters that require the repeater carrier to drop between transmissions. To prevent the scanner from locking-up on a very active repeater, a timer is incorporated to ensure that the scanner does not stay on any frequency more than 3 minutes.

A latch function is provided to permit locking a synthesizer receiver to the scanner's receive frequency by command. If the synthesizer has automatic transmit offset capability, this feature can be used to good advantage. In my case, I've wired the function to a momentary type toggle switch but it's readily adaptable to a push-to-talk (PTT) mike switch. Wire

By Marion D. Kitchens, K4GOK, 7100 Mercury Avenue, Haymarket, Virginia 22069

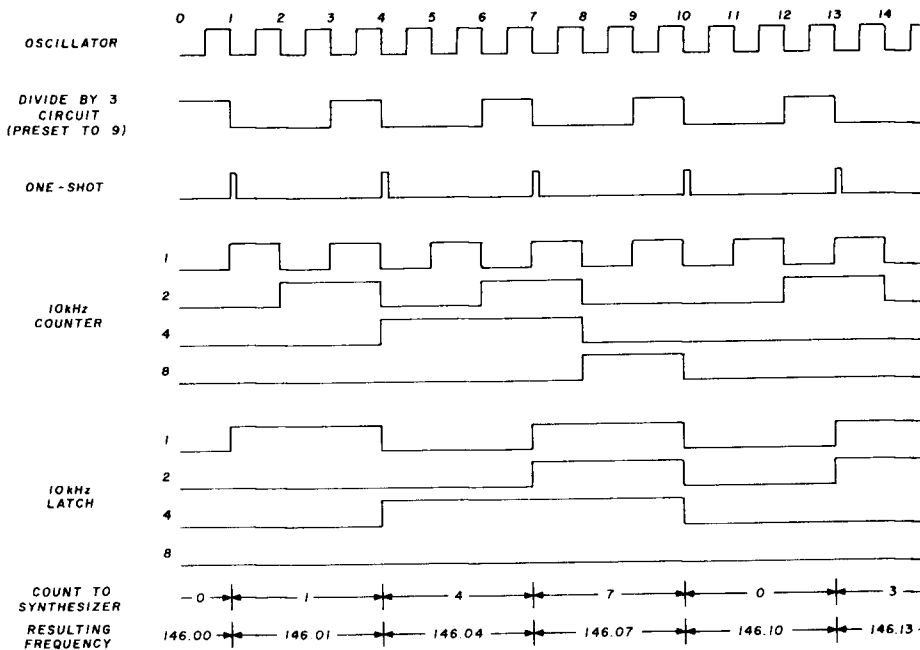


fig. 2. Timing diagram of the scanner. The divide-by-three circuit sets the channel spacing at 30 kHz.

the latch function to the PTT line and when a QSO is to be joined, just push the mike button and the synthesizer is on frequency. The scanner is latched or locked on this receive/transmit frequency, even when the PTT is released. Scanning will continue when manually commanded; the continue command will cause the scanner to run continuously, even though a signal may be present, until the command is removed. If the scanner has stopped on an undesired frequency, a quick tap of the continue switch will move the scanner one or two channels up frequency to where it will resume scanning.

circuit description

The circuit diagram of the scanner is shown in fig. 1. A 555 timer is used as an oscillator to drive three 7490 decade counters. The BCD outputs of the

counters are applied to three 7475 latches. The oscillator output is also applied to another 7490, connected as a divide-by-three circuit, which provides a strobe to the latches on every third count. A time delay is provided by a 74121 one-shot multivibrator to insure that the inputs to the latches have settled.

The basic timing diagram is shown in fig. 2 that indicates the signal relationship between the oscillator, the first decade counter, the divide-by-three counter, and the one-shot delay circuit. The output from the 7475s is a BCD output occurring in steps of three; 1, 4, 7, 10, 13, etc. The latch outputs are also applied to 7446 BCD to seven-segment decoders that drive the displays, and to the synthesizer inputs in place of the frequency control switches. This section of the scanner is standard for TTL counting and display circuits.

Two things are sufficiently different, however, that

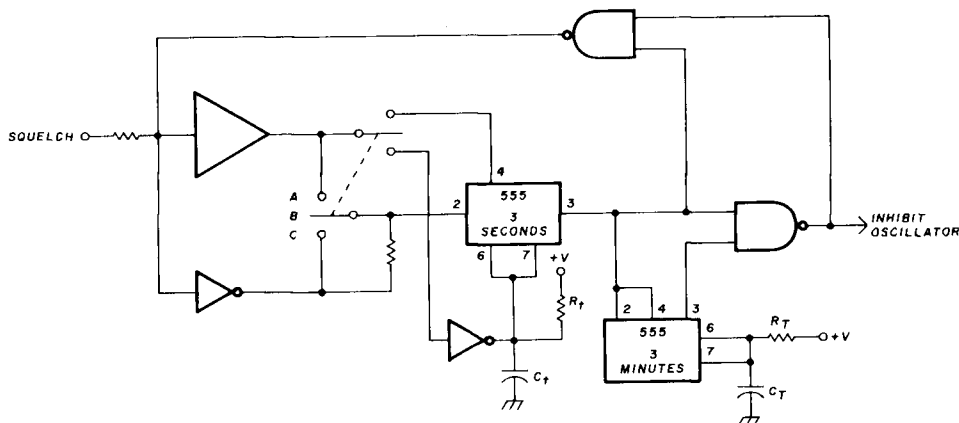
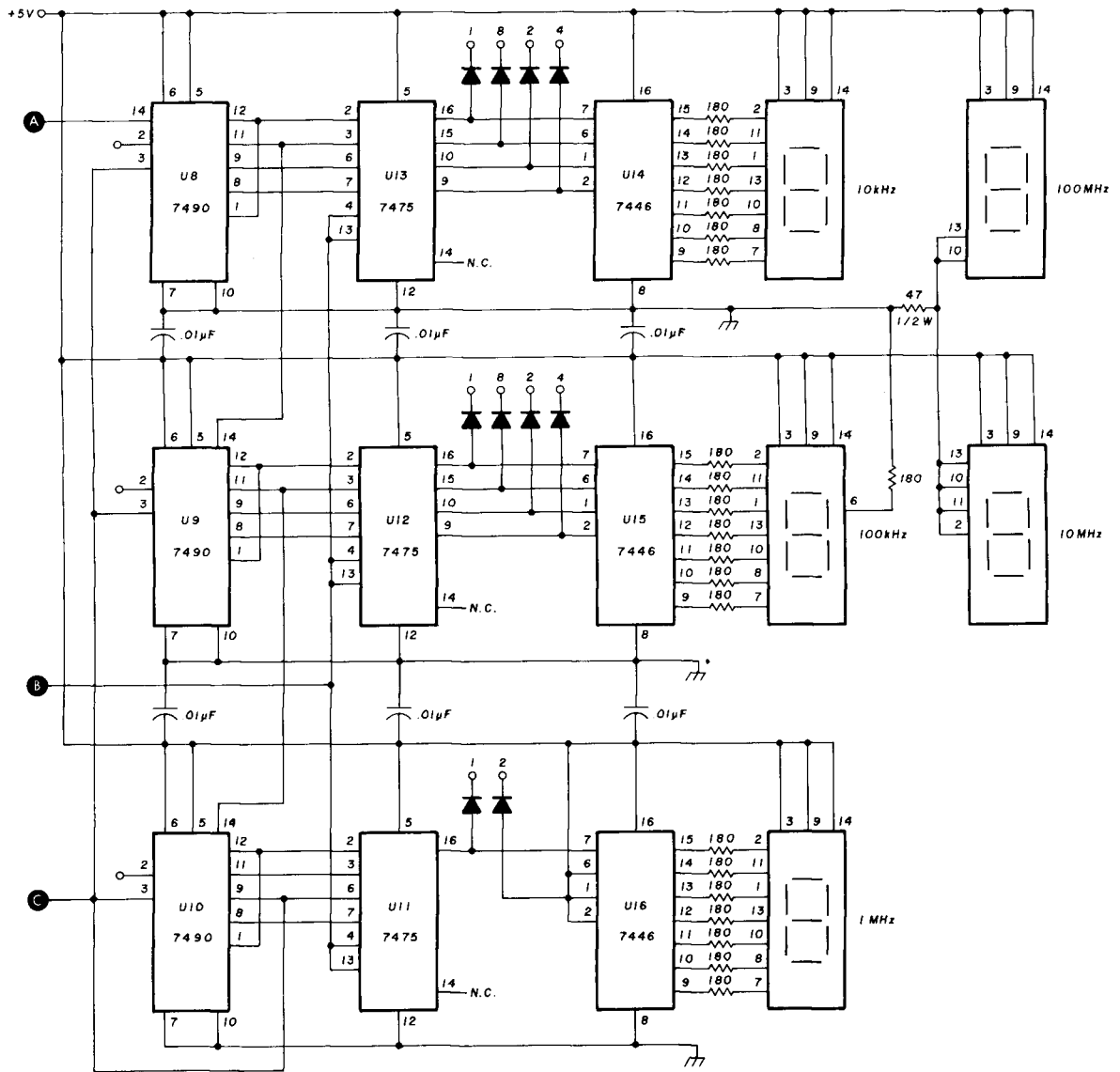


fig. 3. Functional diagram of the 3-minute and 3-second timers. They control the time that the oscillator runs and when it is inhibited.



A and B of U6 to be high and the 7490 resets itself to zero. At this point, the output from the diode OR circuit goes low and strobes the latches. The 7475s now have BCD data for 146.04 MHz stored in them.

Three more pulses from the oscillator will again trigger the 74121 and the 7475s, providing BCD data for 146.07 MHz. The sequence continues for frequencies in 0.03 MHz steps until 148.00 MHz is reached, whereupon the divide-by-three circuit is reset to nine and the sequence starts over.

Two timing circuits are used to provide the three operating modes described earlier. One 555 timer provides a 3-second delay and another timer provides about 3-minutes delay. A receiver squelch circuit is sensed by the 339 voltage comparator whose output controls the 3-second timer. Both true and inverted outputs are obtained from the 339. Fig. 3 shows a partial schematic of the timer circuitry. With the dpdt switch in position A, pins 2 and 4 of U2 are con-

nected to the voltage comparator. An output pulse will be generated if the pins receive a positive going signal from the comparator. The duration of the pulse from U2 is determined by the values of R_t and C_t . The output pulse will terminate after this time even though the input is still present. If the input signal is less than the $R_t C_t$ determined pulse dura-

table 2. Outputs from the 7490 showing the divide-by-three operation.

inputs	7490 outputs				circuit output
	d	c	b	a	
Reset to 9	1	0	0	1	1
Oscillator input count	1	0	0	0	0
	2	0	0	1	0
	3	0	0	0	1
	4	0	0	0	0
	5	0	0	0	1
	6	0	0	1	0
	7	0	0	0	0

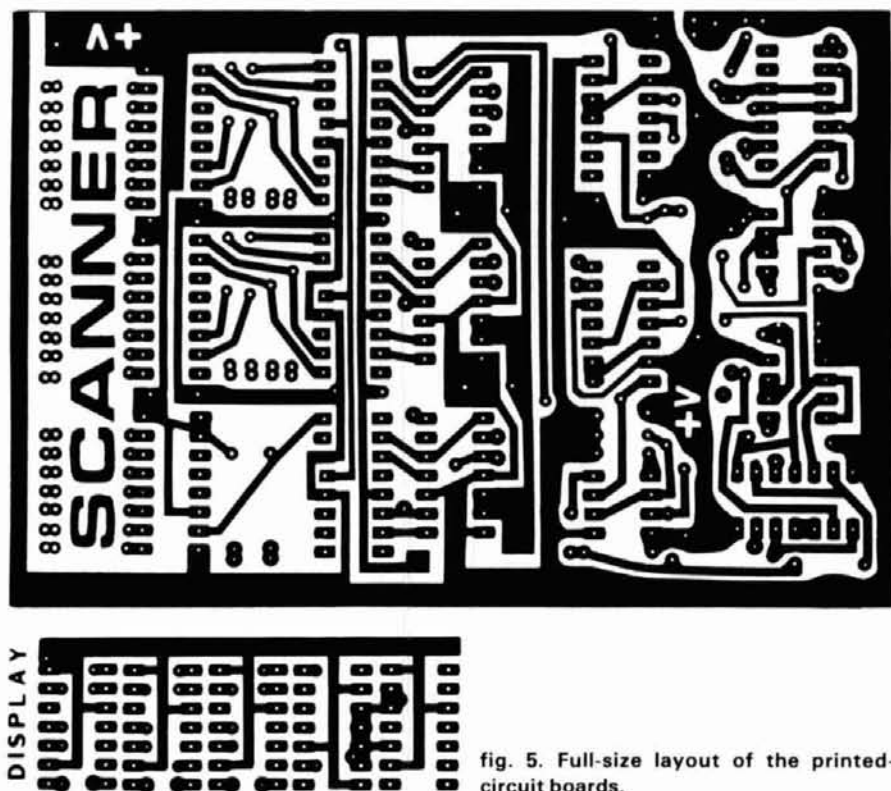


fig. 5. Full-size layout of the printed-circuit boards.

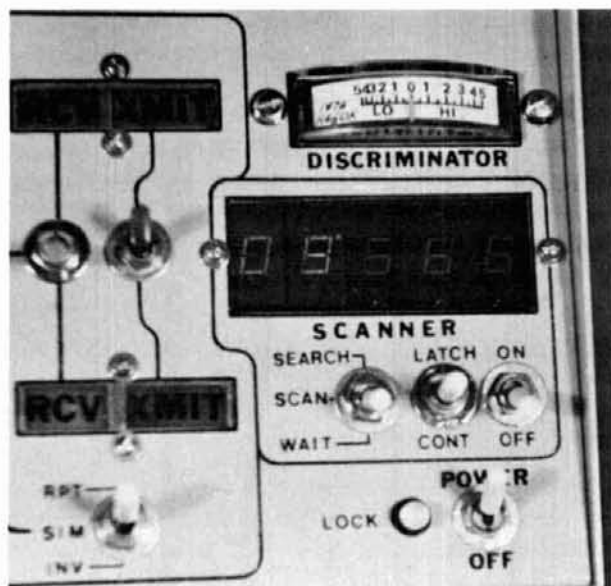
tion, the output will terminate with the input. The output from the timer is used to inhibit the oscillator; this causes the scanner to stop on the frequency that opened the transceiver squelch. In this mode the scanner stops only for a duration determined by R_1 and C_1 , in this case 3 seconds (or less if the signal is present for less than 3 seconds).

With the dpdt switch in position B, the inverted squelch signal is applied through the 1-meg resistor to only pin 2 of the 555. In this mode, the 555 output follows the squelch signal, without delays, as long as the squelch signal is longer than 3 seconds. The 555 output exists for at least 3 seconds even though the squelch input may be shorter. This is the mode used when it is not desired to wait for a return call after a repeater drops.

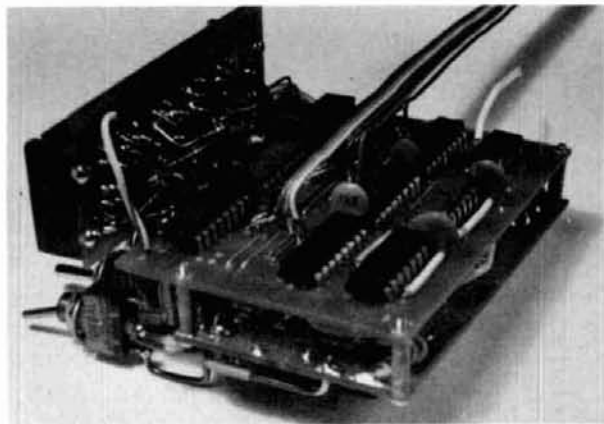
One section of the 339 voltage comparer is used as an amplifier and connected across the timing capacitor of U2. When the dpdt switch is in position C, the amplifier has an input and becomes active. A signal from the squelch drives the amplifier output to ground potential, thereby maintaining C_1 in a discharged state until the squelch is present, and for a period afterwards that is determined by C_1 and R_1 . The receiver listens as long as the squelch is open and then for an additional period, 3 seconds in this case, waiting for a return call. This mode is used with repeaters that require a repeater carrier drop between transmissions. The timing circuits used in the scan-

ner were based on reference¹ and the data in the Signetics data books.²

The 3-minute timer is used to prevent the scanner from locking up indefinitely on very active repeaters. The 3-minute 555 timer (U3) senses the output from the 3-second timer, as shown in fig. 4. It operates just as the 3-second timer does when in mode A. A



Close-up view of the scanner portion of the synthesizer showing the operating controls.



The main circuit board has been split into two separate portions with the display board connected by the 180-ohm current-limiting resistors.

positive-going signal at its input (pins 2 and 4 tied together) causes initiation of its 3-minute output pulse. The 3-minute timer's output is terminated when the input is removed or after 3 minutes, whichever is less. The output of the 3-minute timer is applied to one input of a 2-input NAND (U4B) gate to control passing or inhibiting of the 3-second timer output, which is connected to the second input of

the NAND gate. The output from the NAND gate (pin 6) controls the scanner oscillator. The squelch input to the amplifiers is inhibited by a second NAND gate (U4A) when the 3-minute timer has expired and the 3-second timer's output still exists, thereby resetting both timers.

construction

Construction of the scanner is rather straightforward with use of the printed circuit board. A full-size circuit-board layout and parts placement diagram are shown in **figs. 5** and **6**, respectively. The circuit board can be built as a single unit, or can be cut and assembled into a compact unit as shown in the photographs. Since IC sockets usually cost more than the ICs themselves, soldering directly to the circuit board is recommended.

The circuit board should be built and tested in sections. The recommended sequence is to install the squelch amplifier, timers, and latch/continue chain, and then follow with the 555 oscillator, divide-by-three counter, and one-shot chain. A 1-meg pot should be temporarily installed in place of R_x . This pot will later be adjusted to suit the builder's transceiver and then replaced with a fixed-value

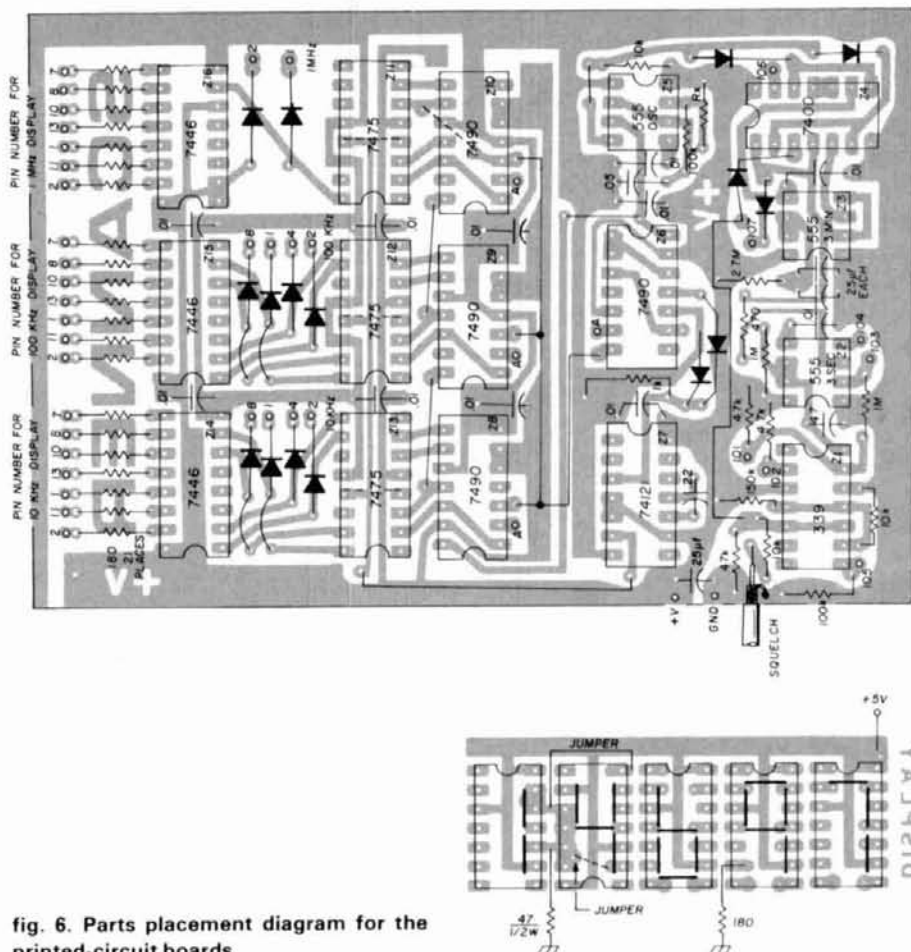


fig. 6. Parts placement diagram for the printed-circuit boards.

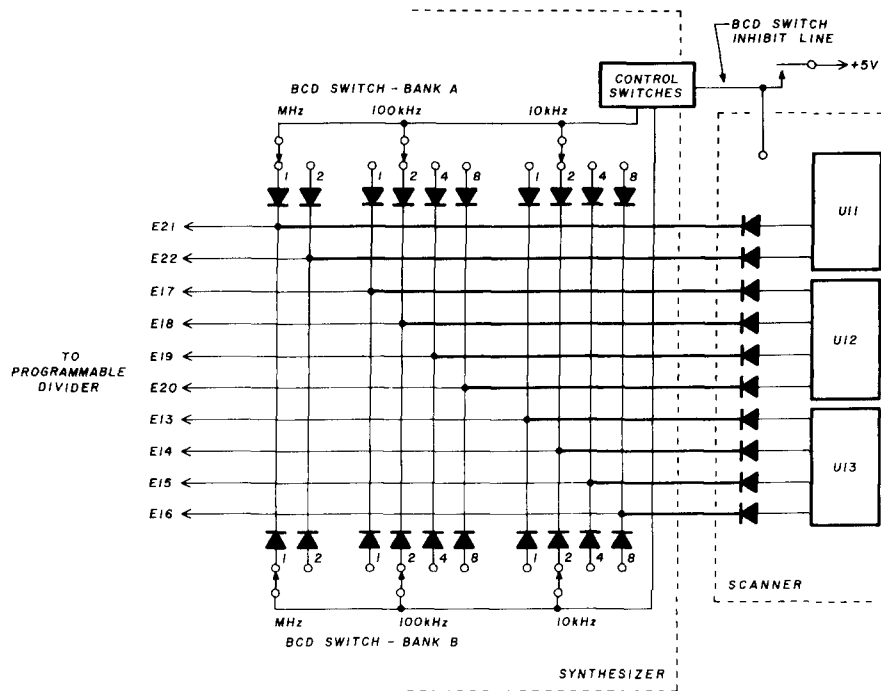


fig. 7. Scanner wiring to synthesizer. The control switches must be disabled when the scanner is in use.

resistor. At this point in construction, a 30-Hz signal should be observed at the oscillator (U5, pin 3) and a 10-Hz signal at the output from the one shot (U7, pin 6). The oscillator should also respond to squelch input signals. The 7490 counters and the 7475 latches can now be installed. Pin 14 of the 7475s must be removed from the package. The pin was removed to facilitate circuit board layout; jumpers on the foil side of the board connect pins 4 and 13 of the latches and between pins 3 and 9 of U10. Check that the

7490s are counting down and data is being transmitted through the latches. Install the decoders and verify the correct information exists at the outputs.

The seven-segment LED displays are mounted on a small, separate printed-circuit board. Note that the 10-MHz display (the number 4) is mounted inverted. The display circuit board will accommodate most of the SLA-1, MAN-7, or 707 displays with a common anode. Be sure to install the jumper wires on the foil side of this circuit board. The displays should first be mounted to the circuit board, then make the connections between the display circuit board and the main circuit board. The main circuit board can be cut near the 7476s and the 180-ohm resistors used to mount the display board to the main circuit board. Make these connections to suit your particular installation.

The control switches should now be wired and the completed scanner checked out before connection to the synthesizer. Use a regulated 5-volt power supply capable of providing about 750 mA. An independent LM309K or similar regulator, supplying power only to the scanner, is recommended.

The scanner outputs from the 7475s are wired to the synthesizer as shown in fig. 7. The synthesizer BCD transmit, BCD receive, and scanner outputs are OR wired through the diodes to the synthesizer programmable divider. When the scanner is in use, neither the synthesizer BCD transmit or receive diodes can have voltage applied; their anodes must be either open or grounded. Fig. 8 shows the circuit used in my homebrew 2-meter TTL synthesizer.

Another method is to replace the synthesizer

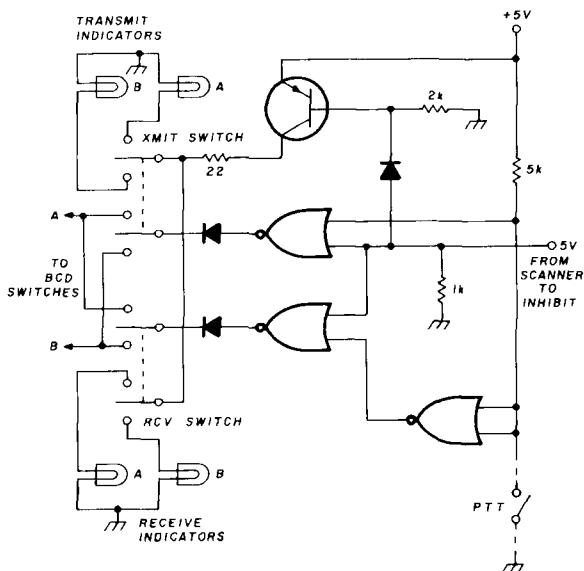


fig. 8. Control switching for the BCD lines in the synthesizer. The normal switches on the synthesizer are automatically inhibited when using the scanner.

receive select switch (used for selecting between the two sets of BCD switches) with a double-throw, center-off switch. This switch *must* be in the off position when the scanner is in use. An alternative to both of the above methods is to set the BCD switches for all zero outputs, usually 144.00 MHz. The squelch input to the 339 comparator (U1) should be connected, with shielded cable, to a point in the transceiver where the voltage goes high when the squelch opens.

Power can be applied to all portions of the scanner and correct operation of the control switches, display, etc. should be verified. The 1-meg pot temporarily installed in place of R_x should initially be set at its maximum resistance. It should then be adjusted for the maximum scan rate, as dictated by the lock-up time of the external synthesizer. After satisfactory operation is obtained, the pot should be replaced by the next largest, fixed value resistor. A 470k resistor can be used for R_x with most transceivers and synthesizers if maximum scan rate is not of particular importance.

circuit variations and additions

A number of practical and interesting circuit and functional variations are possible. This section will present several variations and additions that have occurred to me. Some have been tried, while others are only ideas that you may want to develop to suit your own particular needs.

One variation is to have the digital display indicate both the scanner frequency and the synthesizer BCD switch frequency. This can be accomplished by taking the BCD inputs to the 7446 seven-segment decoders from the synthesizer at the programmable divider inputs. The display will indicate the scan frequency when the scanner is in operation; when off, the display will indicate the synthesizer receive frequency when receiving and the transmit frequency when transmitting. Pull-down resistors may have to be added at the 7446 inputs; the 7446s and displays must, of course, have power when the scanner is off.

Scanning in 10-kHz steps can be obtained by inhibiting or bypassing the divide-by-three circuit. This feature will provide nearly continuous coverage of the 2-meter band.

Fig. 9 shows simple circuit changes that will eliminate scanning of the 147-MHz repeater input frequencies. This change will reduce the scan time by about 2 seconds. No comparably simple way was found to eliminate scanning of the 146-MHz repeater input frequencies.

Other variations include elimination of the scan mode control switch and the latch/continue switch. The scanner will be in mode B (does not wait for a

return call) if the mode switch is simply omitted. Connecting a jumper between points 102 and 105 results in mode C operation (wait for a return call). The latch/continue function is useful and its elimination is not recommended unless minimization of panel space is desired.

Those builders needing the absolute minimum panel space could eliminate the control switches and use 0.1-inch (2.5mm) high displays. Only the 100 kHz and 10 kHz digits need displays. A discrete LED could be used to indicate 147 MHz. The synthesizer

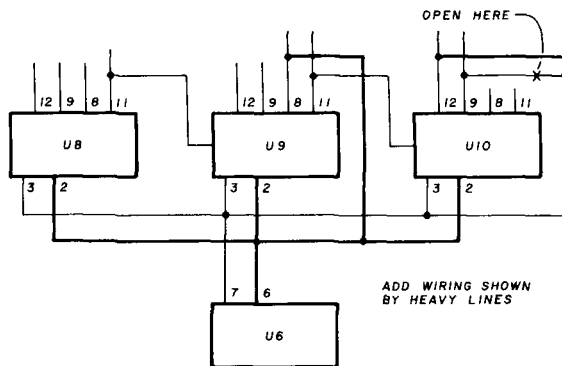


fig. 9. Circuit changes to prevent the counters from covering the 147-MHz repeater receive frequencies.

ON/OFF switch could be a center-off dpdt switch supplying power only to the synthesizer when in one on position, and power to the synthesizer and scanner when in the other on position.

Another addition that could lead to some interesting possibilities is an automatic transmit frequency offset feature. Digital subtraction of 600 kHz could be provided for 146 MHz repeater frequencies and 600 kHz addition provided for 147 MHz repeater frequencies. A center-off, double-throw switch, marked REPEAT-SIMPLEX-INVERT could be used. The latch/continue control could then be used to place the synthesizer on frequency and latch it there. The bulky BCD transmit and receive switches could be eliminated, and a very compact, highly functional synthesizer/scanner could be built.

These variations and additions are presented to encourage building and experimenting among amateurs. I hope that others will build upon and add to these efforts, and eventually present the results for the benefit of all amateurs.

references

1. Mike Connor, WA0BMP, Bob Henson, WB0JHS, "Super COR," 73, June, 1976, page 16.
2. *Signetics Digital, Linear, MOS Data and Applications Book*, 1974, Signetics Corporation, Sunnyvale, California.

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

Specifications

Amateur Bands: 160-10 meters

plus WWV (receive only)

Modes: USB, LSB, CW

Antenna Impedance: 50-75 Ohms

Frequency Stability: Within ± 1 kHz during one hour after one minute of warm-up, and within 100 Hz during any 30 minute period thereafter

Tubes & Semiconductors:

Tubes 3

(5Z001A x 2, 12BY7A)

Transistors 52

FETs 19

Diodes 101

Power Requirements: 120/220 V

AC, 50/60 Hz, 13.8 V DC

(with optional DS-1A)

Power Consumption: Transmit:

280 Watts Receive: 26 Watts

(with heater off)

Dimension: 333(13 1/4) W x 153 (6-0)

H x 335(13-13 3/16) D mm(inch)

Weight: 16.0 kg(35.2 lbs)

TRANSMITTER

RF Input Power: SSB: 200 Watts

PEP CW: 160 Watts DC

Carrier Suppression: Better than -40 dB

Sidband Suppression: Better than -50 dB

Spurious Radiation: Better than -40 dB

Microphone Impedance: 50k Ohms

AF Response: 400 to 2,600 Hz

RECEIVER

Sensitivity: 0.25 μ V for 10 dB (S+N)/N

Selectivity: SSB: 2.4 kHz/-6 dB, 4.4 kHz/-60 dB

Selectivity: CW: 0.5 kHz/-6 dB, 1.5 kHz/-60 dB (with optional CW-520 filter)

Image Ratio: Better than 50 dB

IF Rejection: Better than 50 dB

AF Output Power: 1.0 Watt (8 Ohm load, with less than 10% distortion)

AF Output Impedance: 4 to 16 Ohms



DG-5 (Digital Display)

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

VFO-520S

The VFO-520S is a solid state remote VFO designed to match the TS-520S. It allows VFO controlled cross channel operation when connected to the transceiver. A built-in RIT circuit, with an LED indicator, permits receiver incremental tuning.

SP-520

The SP-520 is an external speaker designed for use with the TS-520S in place of the transceiver's built-in speaker for added clarity.

AT-200

Here's a new and versatile accessory from Kenwood that belongs in every station. The AT-200 is an antenna tuner, but it's also much more. It's an antenna switch, an SWR bridge and an in-line wattmeter. The AT-200 reduces the clutter and increases the operating efficiency of your station... and at a surprisingly moderate price.

TV-506

An easy way to get on the 6 meter band with your TS-520/TS-820/T-599D series and most other excitors. Simply plug it in and you're on... full band coverage with 10 watts output on SSB and CW.

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modifying the Collins 51J receiver for ssb reception

If you're lucky enough
to have one of the
51J-series receivers,
here's an easy way
to update it
for single-sideband reception

One of the most popular surplus receivers is the Collins 51J series, available in limited quantities through MARS and some surplus stores. Designed in the mid 1950s, the stability, readout accuracy, and general excellence of this receiver literally revolutionized receiver design, setting the trend for most of the modern ssb receivers and transceivers. The immediate fallout from the 51J design was the well-known Collins 75A series of amateur-band-only receivers, followed by the present S-line.

The many virtues of the 51J series receivers do not include good ssb reception. An important modification is the inclusion of a product detector and alteration of the automatic gain-control loop to accommodate ssb signals. This article covers these modifications as well as other minor changes that make the 51J into a first-class receiver suitable for

amateur service, including general-coverage operation.

Five models of the 51J receiver are available. The 51J-1 is quite rare; probably the quantity made was small. The 51J-2 and 51J-3 are fairly common on the surplus market; differences between the receivers are minor. The military R-388/URR is similar to the 51J-3. The 51J-4 was the latest production model and incorporates mechanical filters in the i-f system. A choice of three filters may be made with a panel switch.

At one time Collins made an adapter (Collins part number 354A-1) for the 51J-2 and 51J-3 that would modify the receivers for inclusion of crystal filters. The adapter is no longer in production.

The first job for the owner of a 51J is to align it correctly and test all the tubes. Complete alignment information is included in the Collins receiver manual and also in the military technical manual, *Radio Receiver R-388/URR*, TM-11-854, sometimes obtainable through MARS or surplus dealers.

receiver sensitivity

A common fault in most 51J receivers I've inspected is that overall gain is low and the receiver seems dead above about 15 MHz. Investigation has shown that receiver gain is reduced because of an uncommonly high bias voltage applied to the rf tubes. Bias is obtained from a voltage divider in the negative side of the high-voltage power supply (fig. 1). Normal bias voltage is -1.4 volts and, in the receivers tested, has usually run from -1.6 to -3.0 volts. This high

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negative voltage lowers the gain of the rf stages, leaving the receiver lifeless. Bias voltage is developed across resistor R149, which is 820 ohms, 1/2 watt. In many receivers, this resistor looks to be overheated or measures abnormally high in resistance. The cure is to remove R149 (which is located on a terminal strip on the inside wall of the receiver, near the line cord) and replace it with a 2-watt resistor of the proper resistance, which will develop a voltage drop of 1.4 volts across it. You'll find the value will run between 700 and 1000 ohms, depending upon your receiver.

receiver PTO

On occasion a 51J may be picked up for a song because the PTO (permeability tuned oscillator) "doesn't work." The usual cause of malfunction is a collection of matchstick capacitors in the PTO (C005, C006 and C008), which tend to short circuit after a few years of service. These are 0.01- μ F, 400-volt capacitors of a design no longer made. Replacing these capacitors with 0.01- μ F, 600-volt disc ceramic capacitors will usually restore the PTO to operation.¹

the new product detector

Once the 51J has been aligned and is operational, the ssb modification may be added. The circuitry to

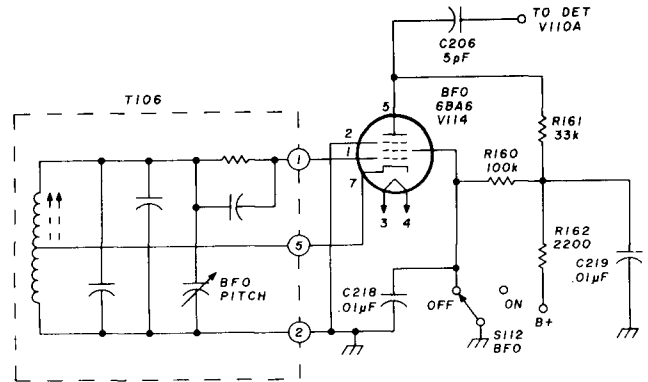


fig. 2. Original 51J BFO circuit. A 6BE6 tube is substituted for the original 6BA6 (V114) to provide a product detector. Tube is turned off by switch S112, which short circuits the screen voltage to ground. (See reference 2 for more details on the tube substitution.)

be modified is shown in **figs. 2, 3, and 4**. The major alteration is in the beat-frequency oscillator (**fig. 2**), which is changed to perform as a product detector. To make this change, the receiver panel may have to be removed, as a new beat oscillator switch (S112) may be required. A 6BE6* is substituted for the 6BA6 BFO tube, and various circuit changes are made beneath the chassis. The final circuitry, after modification, is shown in **fig. 4**.

The first step is to start work on the BFO tube socket (XV114). Most Collins 51Js are wired with high-quality wire having a thin plastic coating, which can be easily damaged by a soldering iron. I suggest, therefore, that you use a small iron with a long, narrow point and proceed carefully so that you don't inadvertently burn any insulation on adjacent wires. Referring to **fig. 2**, remove the following components: R161 (33k), R160 (100k), R162 (2.2k), C218 (0.01 μ F) and C219 (0.01 μ F).

Next, capacitor C206 (5 pF) must be carefully disconnected from XV114 pin 5 and reconnected to pin 7. A 10k, 1/2-watt resistor is then connected between pin 7 and the adjacent ground lug. XV114 pin 2 is ungrounded and reconnected to the BFO transformer pin 5 (center pin) through the 220-ohm resistor and 0.01 μ F combination.

The next step is to solder the 0.05- μ F disc ceramic capacitors in place. One capacitor connects between pin 6 and the nearby socket ground post. The other, in the plate circuit, is attached to an existing terminal stud, which is screwed to the bolt holding the main filter capacitor socket. The 10k, 1-watt resistor is connected between the high-voltage terminal (pin 5 of C217B socket) and the terminal stud. The 47k, 1/2-watt resistor is placed between the stud and pin 5 of socket XV114.

*For additional information on the use of the 6BE6 product detector, see reference 2.

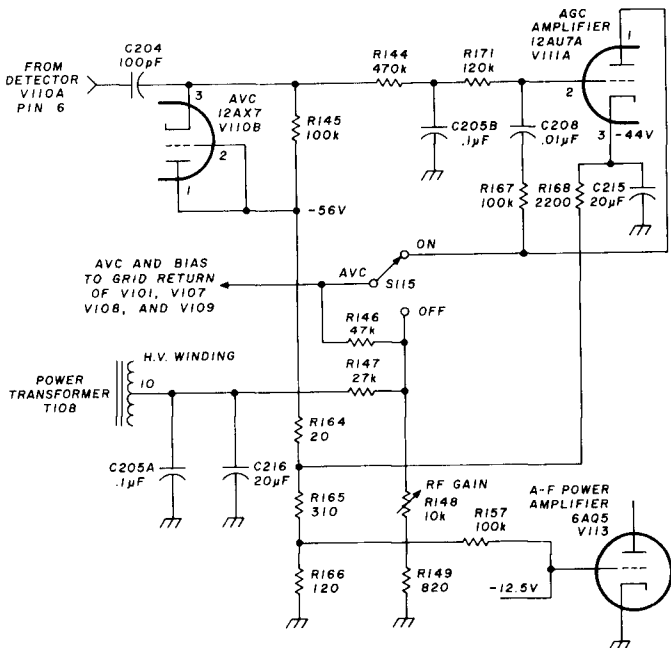


fig. 1. Agc and bias control portion of the 51J receiver. R149 establishes control-bias level. For a negative control voltage V110B and V111A operate below ground. Agc time constant is determined by R144 and C250B. External cathode-to-grid circuit (V111A) should be below 2 megohms after modification to prevent stray "gas current" in the 12AU7 from blocking the agc action. Audio amplifier bias is obtained from the negative source across R166.

The final modifications at this point are to place the 0.1 μF filament bypass capacitor on the socket and revise the audio and agc circuitry.

audio-stage mods

The remainder of the modified circuitry is shown in **fig. 4**. The plate circuit filter components (two 470-pF capacitors and a 47k, 1/2-watt resistor) are mounted on a two-terminal strip placed under one bolt of coaxial socket J104 (marked **if output**). The 0.05 μF coupling capacitor is connected between this assembly and XV114 pin 5.

The 51J panel must now be removed to get at selector switch S112 (BFO OFF-ON) (**fig. 5**). If not, the switch will have to be replaced with the proper type (dpdt). The **A** section shorts the 6BE6 screen supply for am service. The **B** section switches the audio section of the receiver from the product detector to the diode detector, through limiter tube V112A. The audio takeoff point is XV112A pin 3.

To make the interconnections, three coax cables must be run from switch S112 to the rear of the receiver. For ease of wiring, the small-diameter RG-179/U is suggested. The outer braids of the three cables are grounded to the switch assembly on the panel. The cables are dressed into position and run to the respective termination points, at which place the shields are again grounded.

agc mods

To complete this step, capacitor C205A-B-C should be temporarily unbolted from the chassis and moved out of the way.

The agc loop in the receiver is designed to adjust the rf and i-f gain automatically for a-m signals. It must be modified for ssb reception. Pappenfus *et al*³ recommends an attack time of about 0.002 second and a release time of 0.2 to 2 seconds. This time constant can be closely approximated within the limita-

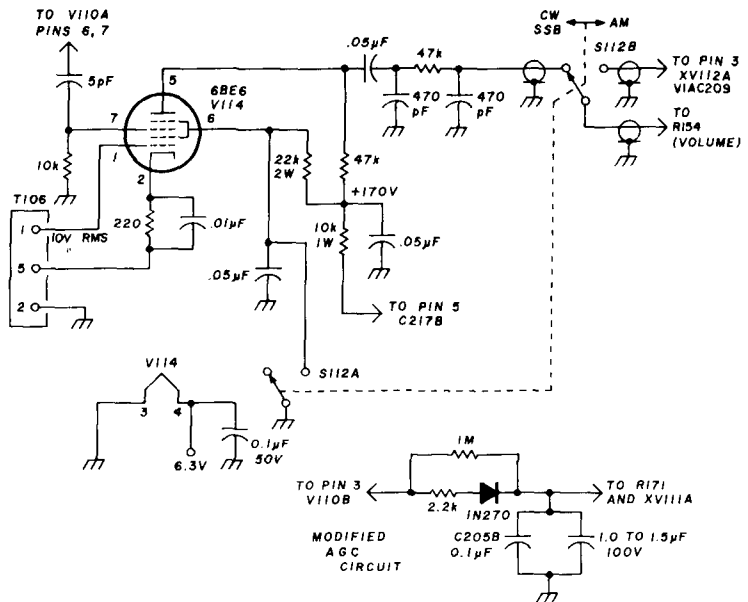


fig. 4. Revised product detector and agc circuit. Caps are ceramic except for the time-constant cap, which is Mylar (see text). BFO injection, measured at pin 1 XV114 socket, should not be more than 10V rms. Oscillator voltage can be set by varying the 22k, 2W screen resistor. Signal injection level is set by the value of the resistance between XV114 pin 7 to ground.

tions imposed by the 51J circuitry. The agc circuit is shown in **fig. 1**. The agc time constant, as the receiver stands, is about 0.06 second, determined by capacitor C205B and resistor R144.

It's theoretically possible to increase the time constant by increasing R144; however, there's an upper limit to the value of this resistance, as pointed out by my friend and colleague, W6PO, who reminded me that oxide cathode tubes such as the 12AU7 are restricted as to the maximum value of grid resistance, which should run less than two megohms.

The reason for this restriction is that a combination

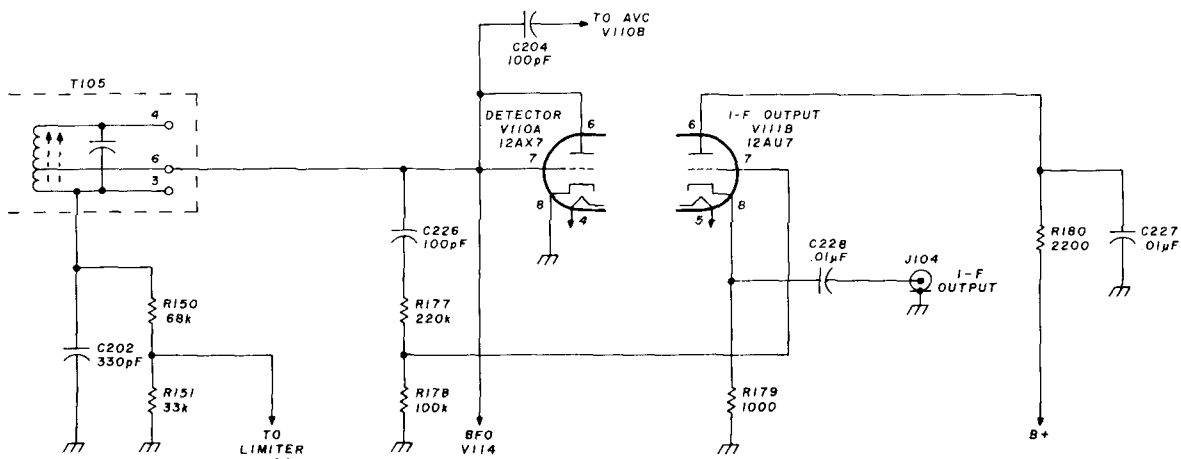


fig. 3. Detector and i-f output amplifier schematic. V110A is connected as a diode detector. Audio is recovered across R151.

of the Edison effect and the migration of oxide from the cathode to the grid as the tube ages can lead to grid emission. An electron flow caused by grid emission (even if only a microampere or so) can seriously disrupt the bias level when the grid resistor is an unreasonably high value. One microampere, for example, flowing through a 2-megohm resistor produces a 2-volt drop, enough to alter the operating characteristics of the 12AU7 agc amplifier tube. The flow of grid current can block the agc line, rendering the receiver inoperative. W6PO recommended that not more than 2 megohms, and preferably less, be used in the agc time constant circuit.

To achieve the desired results capacitor C205B (0.1 μ F), which is part of the timing circuit, must be increased to at least 1 μ F. The use of a low leakage, Mylar capacitor at this point is recommended. The capacitor can be placed directly from the center terminal of C205B to an adjacent ground lug. The resistive portion of the timing circuit is made up of a germanium diode and two resistors. The attack time is set by the 2.2k, 1/2-watt resistor and the release time by the 1 megohm, 1/2-watt resistor. The 1N270 diode disconnects the attack resistor during the discharge portion of the agc cycle. This tiny network is made up and then placed between pin 3 of socket XV110B and the adjacent terminal of capacitor C205B (fig. 4).

testing

After the wiring is checked, the receiver should be tested on a-m to make sure that all original circuits are working. When the BFO switch is turned on, the BFO may be adjusted for good ssb reception. Once satisfied the receiver is working properly, you can check out ssb operation.

The first step is to check for BFO harmonics. With the antenna off, tune the receiver to 1 MHz, 1.5 MHz, and 2 MHz. The BFO harmonics should be heard weakly at the lower frequency and should be in the receiver noise level above 3 MHz. If the harmonics are loud enough to be troublesome, the BFO level should be reduced by increasing the value of the 22k, 2-watt screen resistor on the 6BE6. Once the BFO harmonics have been reduced to your satisfaction (about 2 or less divisions on the S meter at 2 MHz), you can check the product detector for signal overload.

With the constants shown, the signal from the product detector will be somewhat less than that from the a-m detector. The receiver has ample audio gain, so this presents no difficulty. You should be able to tune in a needle-banging ssb signal and receive it crisp and clean. If audio distortion shows up as a growl on speech, this indicates that the product detector is being driven too hard by the i-f signal. The remedy is to reduce the value of the 10k,

1/2-watt resistor in the rf input leg of the 6BE6 XV114 pin 7. In some cases, this resistor value will be as low as 1.2k for low intermodulation distortion.

The 51J receivers vary a bit from one production run to another, and changes in harness layout affect the oscillator level injection, oscillator harmonics, and intermodulation distortion. However, the values

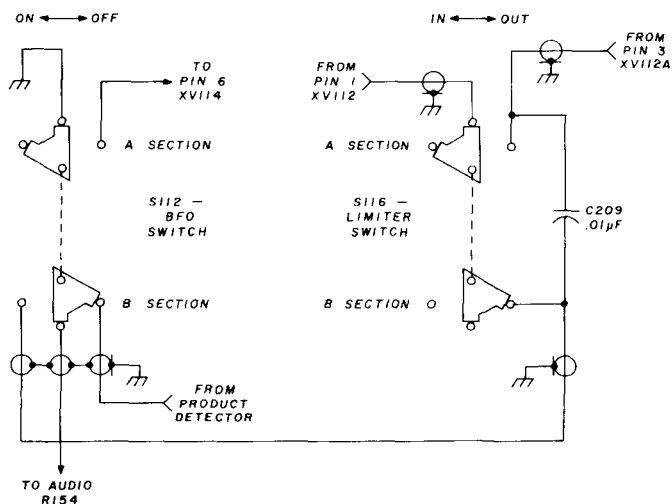


fig. 5. Many 51Js can be wired in this fashion for proper switching. If S112 and S116 are single-pole switches, they must be replaced with double-pole, 2-position, shorting switches. Note that section B of S116 is used only as a tie point for C209.

given in the schematic are representative and are a good place to start from.

parting thoughts

One baffling 51J receiver, after modification, overloaded on even the weakest ssb signal. A painstaking check revealed that some previous owner, anxious of wringing every decibel of gain out of the receiver, had changed the detector tap on transformer T105 from pin 6 to pin 4 (fig. 3). This upset the gain level of the receiver so that overload was inevitable. Changing the modification back to the original circuitry cured the trouble.

The modified 51J, especially if equipped with mechanical filters and a reduction tuning knob, is the equal of the best of today's ssb receivers. How many items of equipment, designed in the mid-1950s can equal that?

references

1. William I. Orr, W6SAI, "Collins 51J PTO Restoration," *ham radio*, December, 1969, page 36.
2. Lee, "The Single-Tube Pocket Detector," *CQ*, April, 1961, and Scherer, "More on Updated Improvements for the 51J Receivers," *CQ*, December, 1968.
3. Pappenfus, Bruene, and Schoenike, *Single Sideband Principles and Circuits*, McGraw-Hill Book Company, New York, 1964.

ham radio

discrete operational amplifier active filters

The functions of
the hybrid active filter
can be replaced
by using individual
operational amplifiers —
the quad op amp
now permits a
package-by-package
replacement

Two recent excellent articles have described the construction of active filter circuits for CW and ssb receiving applications, using the Kinetic Technology type FX-60 hybrid, integrated circuit.^{1,2} As mentioned in K6SDX's article,² the KTI FX-60 "Universal Active Filter" is a basic building block, incorporating three micro-power op-amps with internal chip resistors and capacitors forming multi-loop negative feedback transfer functions. By the addition of external resistors and/or capacitors, the nominal center frequency may be changed, and the outputs modified to simulate a variety of classic filter characteristics.

However, the FX-60 is not always easy to come by since it is a "cull" or production reject of the commercial series FS-60. The FS-60 rejects have one or more tolerances out of limits, but are perfectly ac-

table 1. Comparison of multiple op amp integrated circuits.

manufacturer	type	number of op amps	supply voltage
RCA	CA3401E	four	single (+)
Motorola	MC3301P	four	single (+)
National	LM3900N	four	single (+)
National	LM-324	four	single (+)
RCA	CA3060E	three	dual (+ & -)
National	LM148	four	dual (+ & -)

Note: The CA3401E, MC3301P, and LM3900N are pin-for-pin compatible.

ceptable for experimental and amateur applications; they carry the designation FX-60. The FX-60 is only available directly from the manufacturer, the supply is limited due to a small rejection rate, and the commercial grade FS-60, at a five times higher price, is proportionately less attractive for amateur projects.

Fortunately, there are now a number of inexpensive, multiple op amp ICs which can be used to adequately simulate the basic functions of the FX-60. The multiple op amps can be substituted in most of the circuits for which the FX-60 is specified. A partial list of suitable ICs for this purpose is shown in table 1.

Single supply voltage types require only a positive supply in the range of 5 to 25 volts, and have a built-in center-signal reference. Dual supply types more commonly require both a positive and negative voltage with respect to ground. Some may be found at bargain prices at surplus supply houses.

While all of the ICs listed in table 1 are suitable for active filter applications, I chose the LM324 for further consideration. Though not classified as "micropower," it has relatively low power drain (approximately 700 μ A/amp), low internal noise (allowing use in low-level signal circuits), incorporates four independent op amps, and requires only a single positive supply voltage.

basic universal active filter

Fig. 1 illustrates the basic circuit of the FX-60 with its internal negative feedback loops, and connections for the DIP configuration (viewed from the bottom). The internal resistor and capacitors (R1A/R2A and C1A/C2A) set the nominal bandpass output center frequency of 230 Hz and also the cutoff frequency (f_c) of the lowpass and highpass outputs. This frequency (230 Hz) can be increased by connecting external shunt resistors, R1B/R2B, across pins 1 and 2, and pins 10 and 12. The external resistors are always of equal value for a specific frequency above 230 Hz, and can be calculated from the formula

$$R = \frac{455 \times 10^5}{f_c}$$

where f_c is the desired frequency above 230 Hz. If

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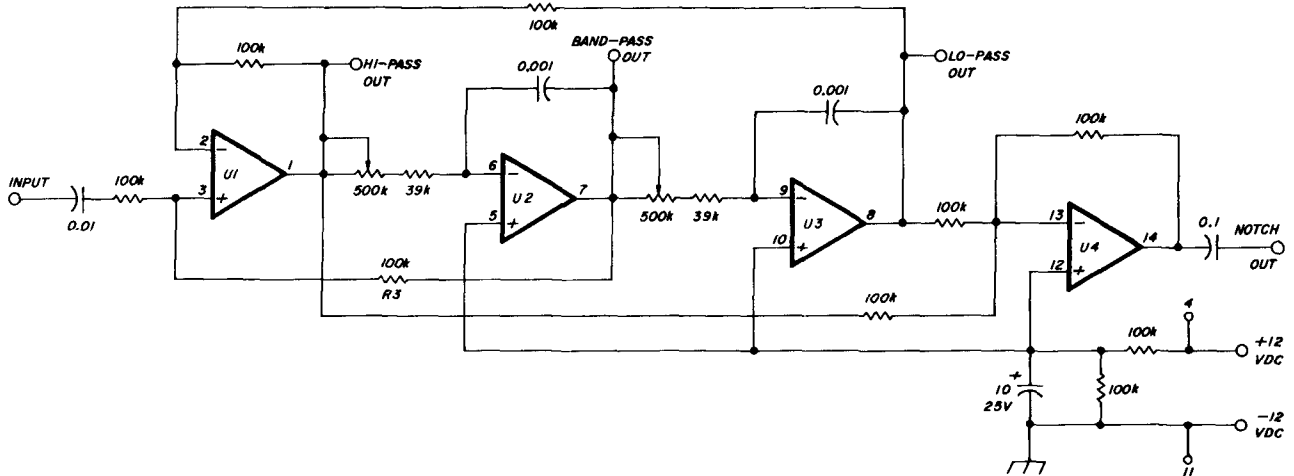


fig. 4. Schematic diagram of a tunable active filter. The highpass and lowpass outputs have been summed in the fourth op amp to provide a notch output. The potentiometers *must* have a reverse log taper.

A fully tunable active filter, covering the range of 300 Hz to 3000 Hz is shown in fig. 4. In addition to the previous highpass, bandpass, and lowpass outputs, the fourth op amp (U4) is used to sum the highpass/lowpass outputs which, being 180 degrees out of phase, result in a tunable notch at the output of U4. The tuning potentiometers are ganged, *reverse log taper*, 500k carbon, 2 watt units. Although exact tracking between the potentiometers is not critical, high quality components are recommended to minimize noise and frequency jumps. A notch of -35 dB can be attained using fixed components with 5 per cent tolerance. This circuit is similar to that used for audio notching in the new Atlas 350-XL transceiver, and is most useful for nulling out unwanted CW signals or broadcast hetrodynes in the 3.8 and 7 MHz bands. The low internal noise of the LM324 permits inserting this circuit between the product detector and first audio amplifier stages of a receiver.

Resistor R3 establishes the *Q* for a notch width of 200 Hz at the -3 dB points. While the notch may be

narrowed by increasing the value of R3, tuning for maximum notch depth becomes increasingly difficult; 200 Hz is about optimum, for ease of adjustment. The notch output has unity gain with respect to the input signal, and any variations due to component tolerances can be adjusted for by trimming the value of the 100k resistor between pins 13 and 14 of U4.

This tunable active filter is generally useful for amateur receiver applications since the choice of high, low, bandpass, or notch outputs may be switched. It should be noted, however, that the highpass, lowpass, and bandpass outputs have gain with respect to the input. A resistive attenuator (minimum 1 megohm) coupled through a 0.1 μ F capacitor to each of these outputs, is required to adjust the levels for unity gain.

An alternate, fixed-frequency notch filter, using only three op amps, is shown in fig. 5. Other than reduction of components, this circuit has no inherent advantage, but lends itself to triple op amp ICs. Not easily adaptable to tuning, this circuit is useful for discrete frequency notching.

general considerations

Reasonable care must be taken when laying out any circuit that uses multiple outputs and feedback loops. The LM324 is particularly well suited to minimizing stray coupling, since the output terminal of each op amp is located at the four corners of the DIP IC. Stray coupling between the input and output of the separate op amps must be avoided to prevent instability or performance degradation. This is particularly *important* in the notch filter circuits where stray coupling may limit the attainable notch depth.

If you wish to use one of the suggested devices other than the LM324, for a filter, I recommend that

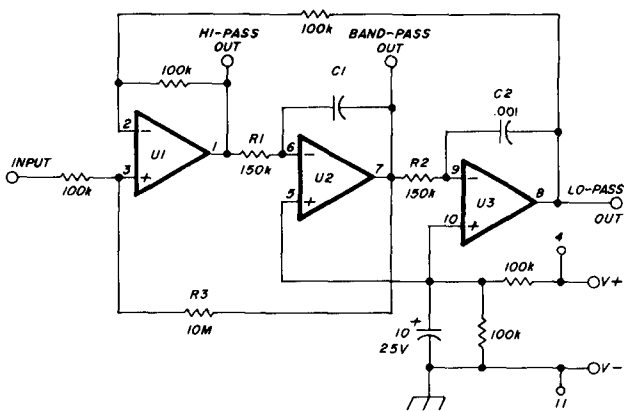


fig. 3. A practical fixed-frequency active filter using the LM324. The center frequency is 1 kHz, with a *Q* of 50.

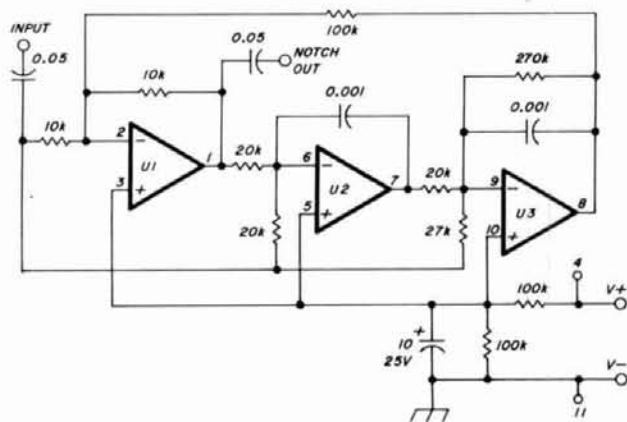


fig. 5. A fixed frequency notch circuit can be formed by using three sections of an LM324. This arrangement does not lend itself well to adjustable notch frequencies. The center frequency for this circuit is 3 kHz.

you consult the manufacturer's specifications regarding supply voltages. For single-supply voltage types, the biasing requirements can be uniquely

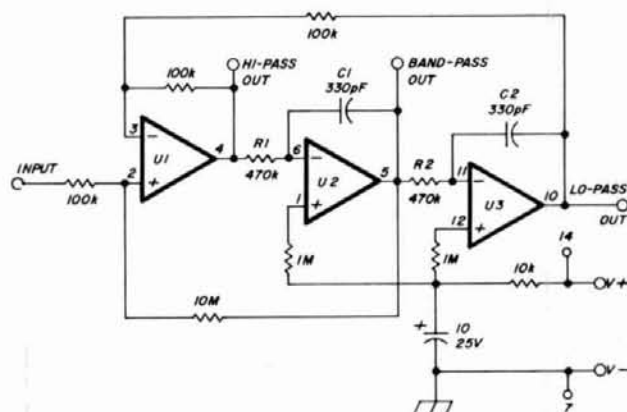


fig. 6. A basic fixed frequency active filter that uses the compatible CA3401E, MC3301P, and LM3900N. The center frequency is 1 kHz. The 1 megohm resistor in the non-inverting lead is used to limit the input current.

different. As an example, fig. 6 shows a fixed frequency filter using the CA3401E, MC3301P, or LM3900. These pin-compatible devices employ an internal "current mirror" transistor for the single polarity supply. Compared with the previous circuits, you can see that these devices require a different biasing arrangement, including high-value series resistors for the non-inverting inputs to limit bias current to between 10 and 100 μ A.

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1. Ken Hollady, K6HCP, "Tunable Audio Filter for Weak-signal Communications," *ham radio*, November, 1975, page 28.
2. M. A. Chapman, K6SDZ, "Audio Filters for Improving SSB and CW Reception," *ham radio*, November, 1976, page 18.

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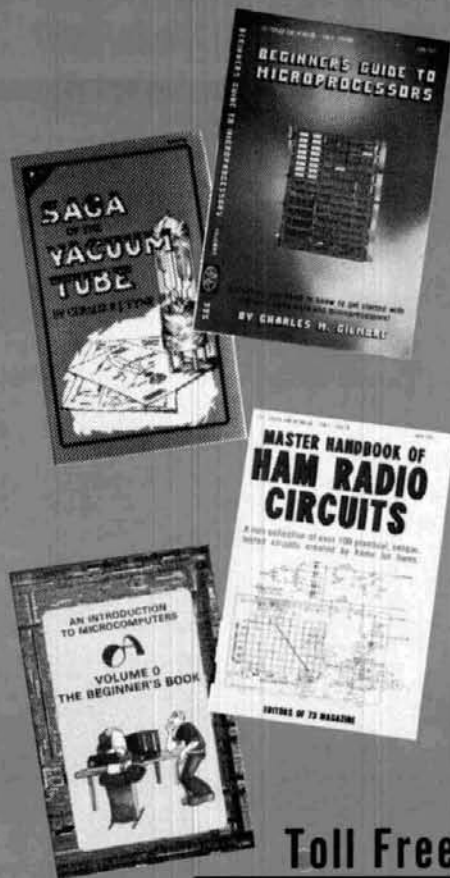
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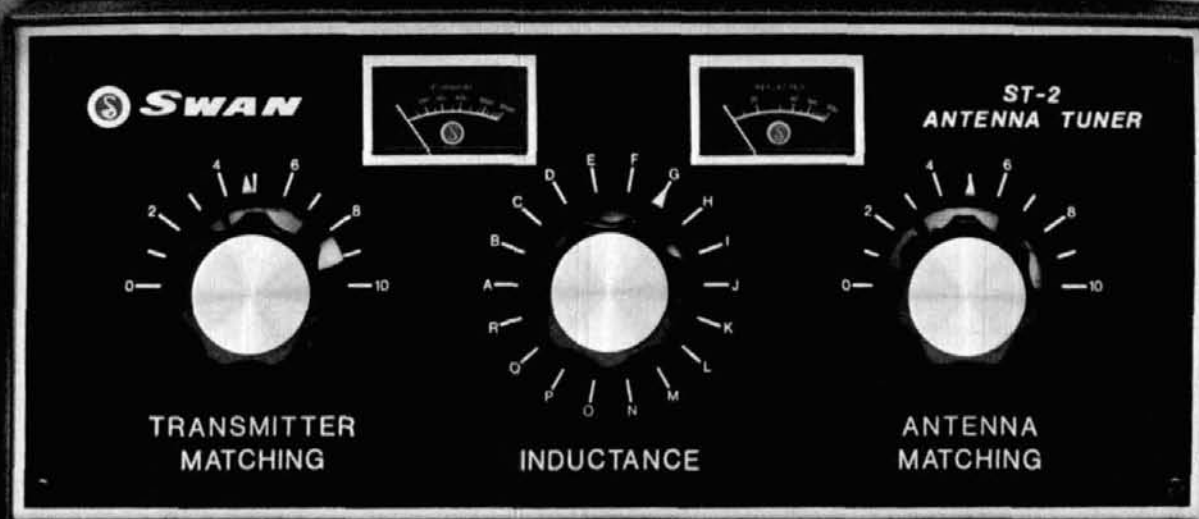
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For both balanced and unbalanced output, Swan's new ST-1 and ST-2 Antenna Tuners put a reliable 50 Ohm source between your 160-10 meter transmitter... and virtually any type antenna system.

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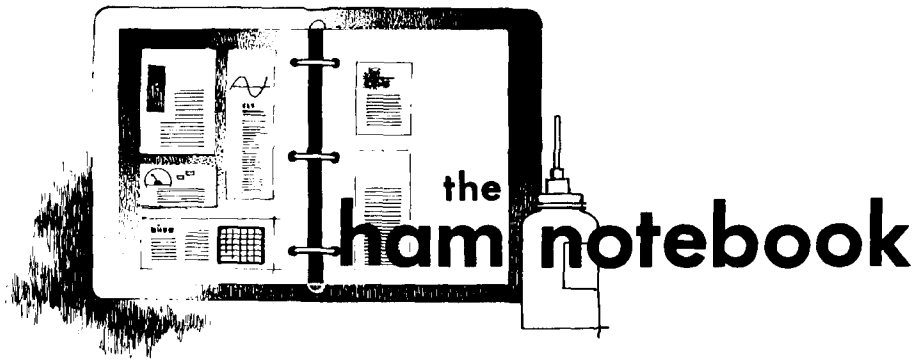
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high voltage fuses in linear amplifiers

The addition of two short pieces of wire to many linear amplifiers will protect expensive components from damage in case of an arc-over or short in the high voltage circuit. Most high-voltage power supplies are fused in the primary circuit only, and a failure in the amplifier can destroy the rectifier string, grid, and plate current meters long before the primary fuse can open.

The partial circuit shown in **fig. 1** is a typical grounded-grid amplifier with the plate meter in the negative lead of the power supply. Note that the negative side of the power supply is

not directly grounded. The ground path goes through both the grid and plate meters. If capacitor C2 shorts out, the short circuit current will go through both meters and, if the filter capacitor, C1, is large, this current can have an instantaneous peak of hundreds of amperes. Such a failure will surely destroy the meters and will very likely destroy the rectifier string. The meter coils will be vaporized and, if the meters are sealed, the glass faceplates may blow out.

The solution is to add high voltage fuses F1 and F2. Each fuse consists of a short piece of no. 40 AWG (0.08mm) copper wire. This wire has a fusing current of 1.75 amperes, high enough that it should never

open up under normal circumstances, but low enough that it will blow in a hurry should a short or arc-over occur. Low voltage glass fuses must not be substituted in this application; they will explode when the internal element vaporizes. They'll also take longer to open up fully as the vaporized element will sustain an arc until the glass breaks and allows it to dissipate. This delay, while probably no more than a few milliseconds, may be long enough to damage the meters.

If resistors R1 and R2 are not present, they should also be added. Their purpose is to keep the negative lead of the power supply from going to a high negative potential with respect to ground should either of the meters or F2 open. They have no effect on normal circuit operation since they are in parallel with the meters, whose resistance is a fraction of an ohm.

In my homebrew 4-1000A linear, these fuse wires have blown twice due to arc-overs in the amplifier. On both occasions, they prevented damage to the power supply and meters, responding fast enough that the primary fuses did not blow at all. In seven years of heavy use, they have never failed during normal operation. I call that cheap insurance.

John Becker, K9MM

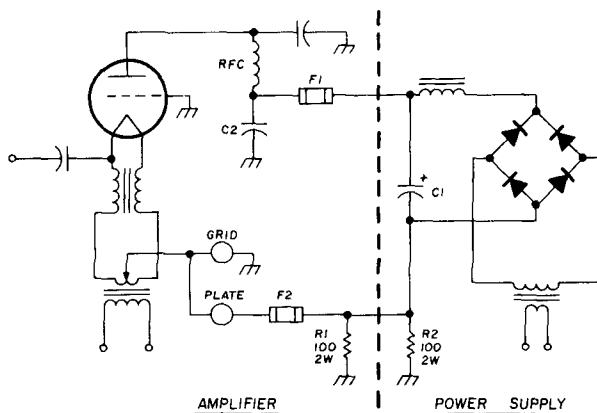


fig. 1. High-voltage fuse circuit for linear amplifier power supplies. Care should be exercised during the installation of the two fuses, F1 and F2. F1 is in the actual high voltage path.

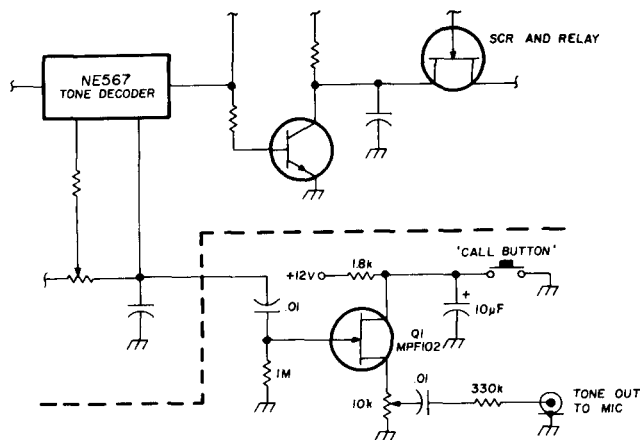


fig. 2. Schematic diagram of the source follower connecting the NE567 decoder and the transmitter. The tone is available, for the transmitter, when the CALL button is pushed.

dual-function integrated circuit

The article "private-call system for vhf fm," *ham radio*, September, 1977, required a separate tone oscillator be used by the initiating station. In reality, the NE567 tone decoder is actually already oscillating at the required frequency. Fig. 2 shows a method of using this IC for both originating the tone, and decoding it upon reception.

Cal Sondgeroth, W9ZTK

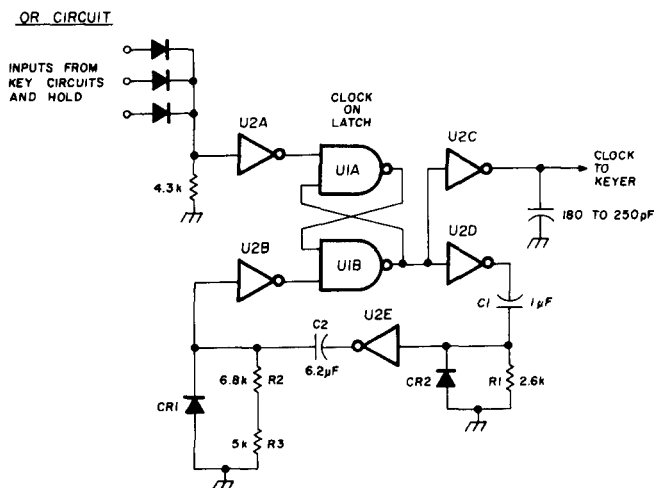


fig. 3. Schematic diagram of the keyer oscillator. U1 is an SN7400, while U2 is an SN74L04. The diodes on the input of U2A form an OR gate that controls the oscillator. These inputs can be used to keep the oscillator running, providing the self-completing feature. The time constant, as determined by C1 and R1, is 4 mS; this is the width of the clock pulse. The values for C2, R2, and R3 give a pulse repetition time of 50 to 95 mS, which equates to approximately 12 to 24 words per minute. For higher speeds, C2 and R2 can be reduced.

capacitor on the output is necessary to prevent noise spikes from falsely triggering the keyer circuits.

J. T. Miller, WB6VZW

wire-wound potentiometer repair

Exact replacement units for those expensive wire-wound pots are often difficult to find. This factor makes repair of the defective control attractive.

The winding is repaired by bridging the opening with a small strip of thin-

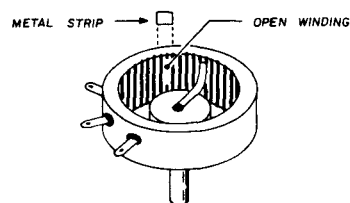


fig. 4. The open winding of a wire-wound potentiometer can be repaired by inserting a small metal strip between the winding and the outer insulation.

sheet metal as shown in fig. 4. One possible material is the metal from a tin can. Cut the strip slightly shorter than the element width, and wider than the break in the winding. Curve the metal strip to conform with the shape of the resistance element.

With power off, remove the rear cover to expose the wire resistance element. The opening in the winding is often evident by discoloration from overheating. Otherwise, it may be located by activating the equipment and adjusting the control knob to the setting where abnormal noise or other faulty performance occurs. The defect is now located directly under the movable wiper contact. Near the break, gently pry the resistance element away from the outer insulation using a thin screwdriver or knife point. This will permit starting the bridging strip into the opening. Now, press the strip behind the resistance element so that it does not interfere with free operation of the slider, or cover replacement.

Gene Brizendine, W4ATE



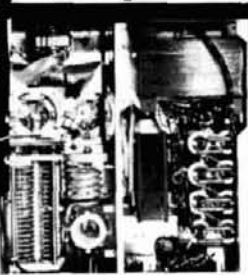
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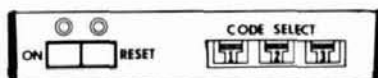
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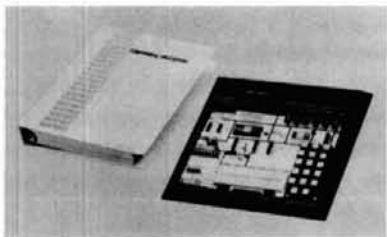
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Heath Self-Instruction Microprocessor course



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EE-3401 microprocessor course and ET-3400 microprocessor trainer, mail-order priced at \$89.95 and \$189.95 respectively, write for a free catalog to Heath Company, Department 350-460, Benton Harbor, Michigan 49022.

multi-family high-speed logic probe



Continental Specialties Corporation has announced the addition of the multi-family, high-speed *Logic Probe 3* to its expanding, popular line of prototyping and test equipment. A more advanced version of the company's popular *Logic Probe 1*, the LP-3 provides precise indications of static and dynamic circuit conditions.

Drawing power from the circuit under test, the LP-3 offers pulse detection, pulse stretching and latching (memory) capabilities, for an instant readout of logic levels; positive and negative transitions or pulse symmetry. The LP-3 also provides positive indication of one-shot and low-rep-rate events via its latching function. Even though the LP-3 is highly accurate (within 0.15V or less), it requires no calibration or other adjustments.

Incorporating special circuitry designed for use with high-speed logic, the LP-3 allows detection of pulses as short as 10 nanoseconds. In the PULSE mode, a built-in pulse stretcher causes the PULSE LED to blink at a 10 Hz rate, indicating pulse transitions. Two level-detector LED's, HI (logic 1) and LO (logic 0) indicate log-

ic levels present in the circuit. At high frequencies, the LP-3 will also indicate whether signals are symmetrical. Pulse trains with duty cycles of less than 30 per cent will activate the LO LED and PULSE LED, while duty cycles of 70 per cent or more will activate the HI LED, in addition to the PULSE LED. The LP-3's high input impedance (500k), which is constant in all logic states, prevents circuit loading problems in both TTL/DTL and CMOS ranges.

The LP-3 also features a plug-in connector system for additional versatility. The plastic housing is equipped with phone jack connector, which accepts the unit's standard 36 inch (91cm) clip-leads. The front of the LP-3 accommodates plug-in, interchangeable tips, including a straight needle probe, alligator clip, quick-attaching insulated clip, as well as a ground lead. At \$69.95, it also represents a major cost breakthrough in high-speed digital testing. The versatile LP-3 is a compact, 5.8 x 1.0 x 0.7 inch (147x25x18mm), portable, easy-to-use instrument for all types of design, testing, and troubleshooting applications.

For more information, contact Continental Specialties Corporation, 44 Kendall Street, Box 1942, New Haven, Connecticut 06509.

quartz technology manual



A new Quartz Technology Manual has just been released by Sentry Manufacturing Company, a leader in quartz-crystal frequency control for quality communications and time-base standards. The 150-page man-

R-X Noise Bridge

ual takes the mystery out of understanding and using quartz crystals. It also includes the most complete military-specification summary ever published, to simplify military-equipment design.

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morse to video converter



DGM Electronics has recently introduced the MVD-1000 Morse Video Display. The MVD-1000 connects between your station receiver and video monitor, or slightly modified television receiver to provide a complete 6-60 wpm morse code receiving system. The Morse code audio signal from the speaker terminals or phone jack is converted to a standard composite video signal which can be used to drive a video monitor or the video amplifier of a television receiver.

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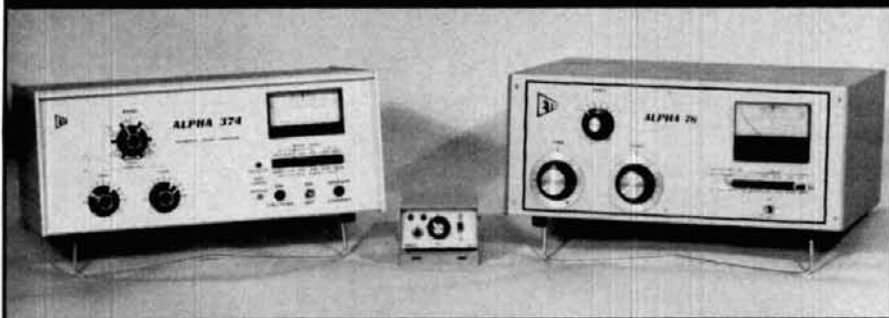


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of displays. When the screen is full, automatic scrolling shifts all lines up one line causing the top line to be stored on the second page, allowing new characters to be written on the bottom line. The *recall* feature allows you to look at the second page while still copying incoming code on the first page. The unit will copy letters, numbers, punctuations, and five special characters for a total of 51 decoded Morse characters.

Automatic and manual speed tracking are provided to accommodate wide variations in the Morse code signal. A 4-pole active filter provides a 200-Hz bandwidth for excellent QRM rejection. A practice input is also provided, for a straight key or keyer output, to allow the user to practice sending while displaying what is sent on the display. This feature is very helpful in learning the Morse code and increasing your sending and receiving speed.

The unit is housed in an attractive 3 x 10 x 10 inch (7.6x25x25cm) case with a gray wrinkle finish cover and a brushed aluminum front panel. The MVD-1000 Morse Video Display is priced at only \$350.00 postpaid, from DGM Electronics, 787 Briar Lane, Beloit, Wisconsin 53511.

electronic equipment and kits catalog

A new 36-page factory mail-order catalog of electronic equipment and kits is now available from Caringella Electronics, Incorporated. Many hard-to-find products, all manufactured by Caringella Electronics, are listed in the catalog. Among the many unique items found in the catalog are a WWV standard-time receiver, panel-mount digital clocks with and without BCD outputs and day-of-year calendar, giant digital wall clock, audio compressors, and electronic security alarms.

More than just an ordinary catalog, the format has been designed to take the guesswork out of buying! Each



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product is completely described with all technical specifications, a photograph, application information, schematic diagram, and a complete description of how each unit works.

The new free 36-page factory mail-order catalog may be obtained from the factory by writing to Caringella Electronics, Incorporated, Post Office Box 727, Upland, California 91786.

Kenwood TS-520S Transceiver



Trio-Kenwood has recently introduced the new TS-520S hf transceiver. It retains the field-proven characteristics of the popular TS-520, and incorporates many of the ideas, comments, and suggestions for improvement from amateurs worldwide.

The first major difference to be seen in the TS-520S is its full-band coverage, including 160 meters. Of course, transverters for 6 and 2 meters can also be used with the TS-520S. Provisions for attaching a digital display (model DG-5) have been included on the back panel. The DG-5 contains 6 digits which display your operating frequency while you transmit and receive.

Other new features of the TS-520S include an rf attenuator; new, improved speech processor; vernier tuning for final plate control; and a new monoscale analog dial. The TS-520S is completely self-contained with a built-in ac power supply. The addition of the DS-1A dc to dc converter (optional) permits mobile operation of the TS-520S. The transceiver also has two convenient RCA-type phono jacks on the rear panel,

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labeled *Phone Patch In* and *Phone Patch Out*, for quick, easy, hookup of almost any phone patch. For CW buffs, the CW-520 500-Hz filter can be easily installed and will provide improved selectivity on CW.

In addition to the DG-5 and CW-520, other accessories such as an external vfo (VFO-520S) and external speaker (SP-520) are also available to make a complete station which any amateur would be proud to own. These and other Kenwood Products are available from authorized Trio-Kenwood dealers throughout the United States. For a list of authorized dealers and more information on the TS-520S, write Trio-Kenwood Communications, Inc., 1111 West Walnut Street, Compton, California 90220.

The Low and Medium Frequency Radio Scrapbook

The frequencies below the 160-meter amateur band were the cradle of amateur radio until hams were exiled to the "useless" short waves. A number of experimenters are probing the world between 10 kHz and 1600 kHz. Here, through a little-known provision of FCC regulations, experimenters are permitted to operate unlicensed transmitters. Operation is presently limited to a power of only one watt and antennas smaller than ten feet (3 meters) but that is adequate for radio communications out to several hundred miles when conditions are good.

The transmitting restrictions only increase the challenge and fascination of these all-but-ignored frequencies. Low-frequency experimenters have rediscovered the thrill of the earliest days of our hobby when transoceanic DX was a pipe dream, and real-life DX records were well under 100 miles.

Ken Cornell, W2IMB, has put a time machine between two covers. His book describes loose couplers, honeycomb coils, absorption wave-

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meters, and classic loop designs — all of which were mainstays of the wireless pioneers and still serve effectively today. On the other hand, Cornell includes IC and modern filter technologies in the endless quest for transmitter efficiency and the conquest of the plague of all low-frequencies — man-made noise.

The book includes dozens of simple circuits and diagrams for electronic and mechanical station components — most using readily available parts. Indeed, some experimenters might label W2IMB's vacuum tube circuits as technological dinosaurs, but they are time-tested, flexible, forgiving, amateur favorites. An objective look also reveals that efficient low-frequency vacuum tubes are plentiful and inexpensive. In an age of black boxes and IC chips, who can deny the nitty-gritty, hands-on learning opportunities of basic, discrete-component construction? Cornell does include many solid-state circuits, but the emphasis is clearly on proven tube techniques.

This is an informal experimenter's scrapbook with a unique format. It is pre-punched so you can conveniently keep it in a standard three-ring binder along with your own collection of notes, catalogs, and article clippings. The book has no fancy printing or polished professional prose; but it is a straightforward collection of one man's views, tips, experiences, and suggestions gleaned from years of actual on-the-air experimentation.

Of particular interest are the author's chapters on converting low-cost military surplus equipment; construction of receiver converters for the most popular 160- to 190-kHz band, excerpts from the relevant FCC regulations, and perhaps most useful of all, comprehensive details on the design and winding of low-frequency coils.

Soft cover, 110 pages, \$6.95 postpaid from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048. Order catalog number HR-LF.






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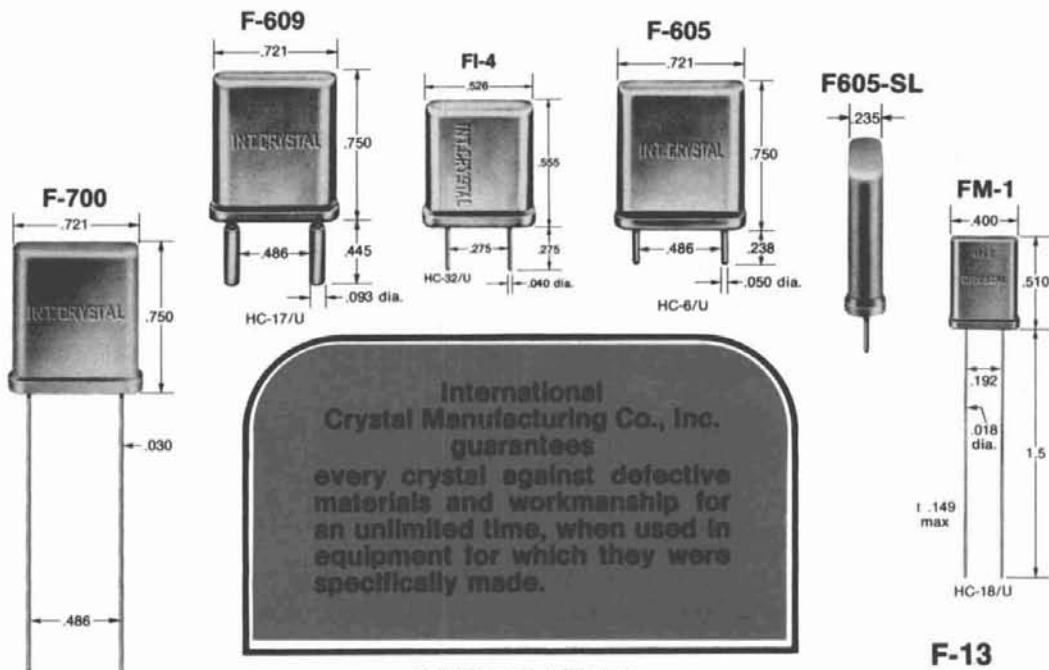
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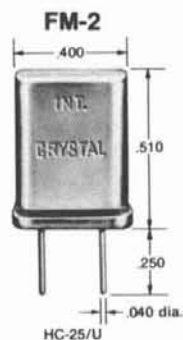
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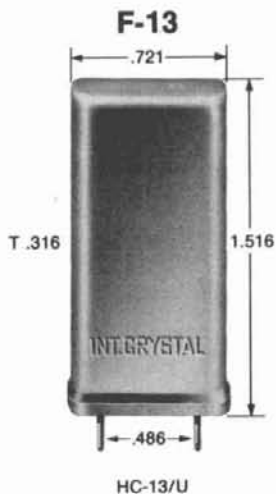
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273 Crystal for 28.5-29 MHz	\$ 5.00
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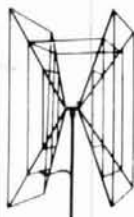
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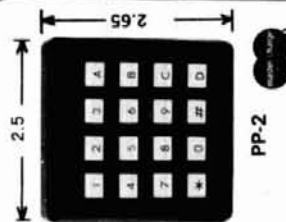
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Receiver gain	30 dB typ
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Prime Power	12 V D.C.

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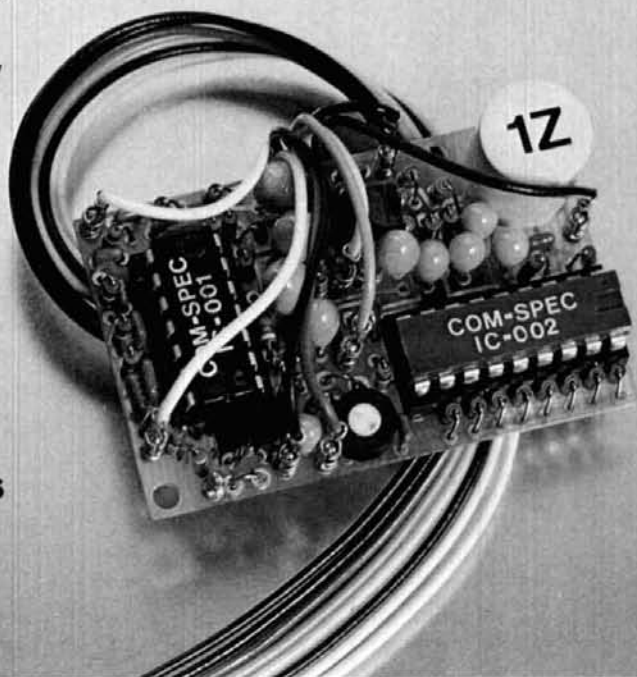
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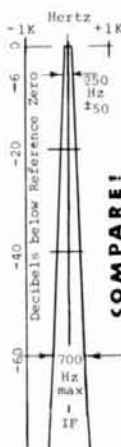
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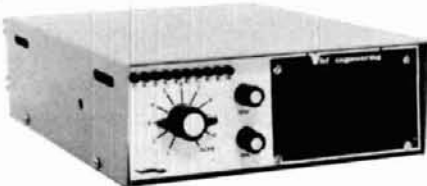


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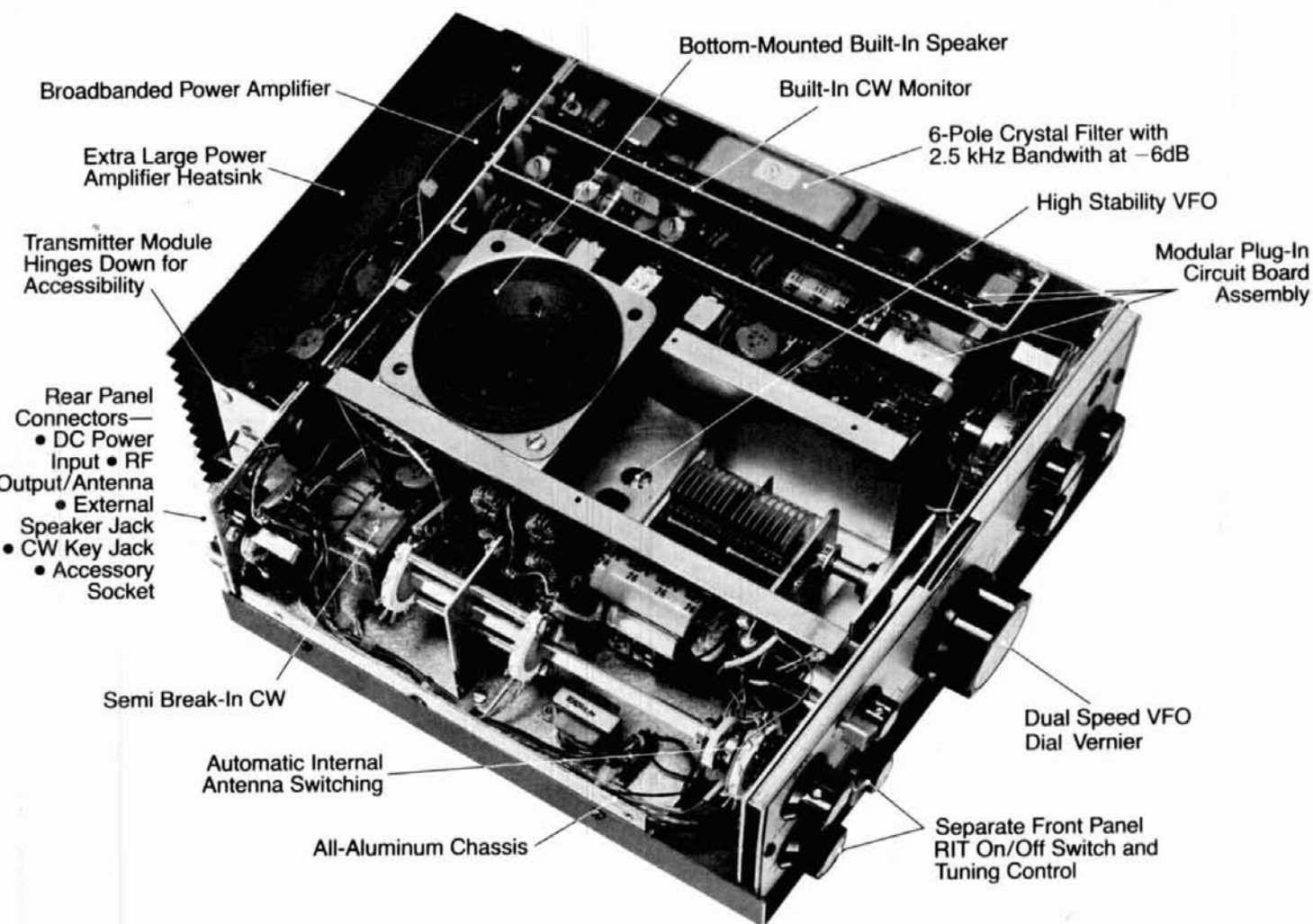
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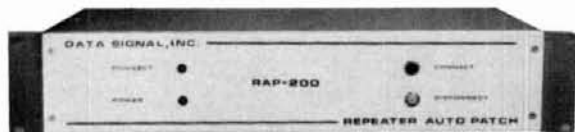
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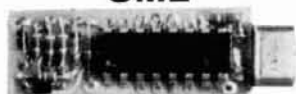
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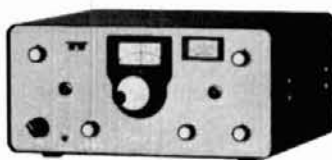


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Insertion VSWR	1.1 to 1.0 max.	
Accuracy	± 8% of Full Scale	
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*May be used with 50-52 ohm cable.

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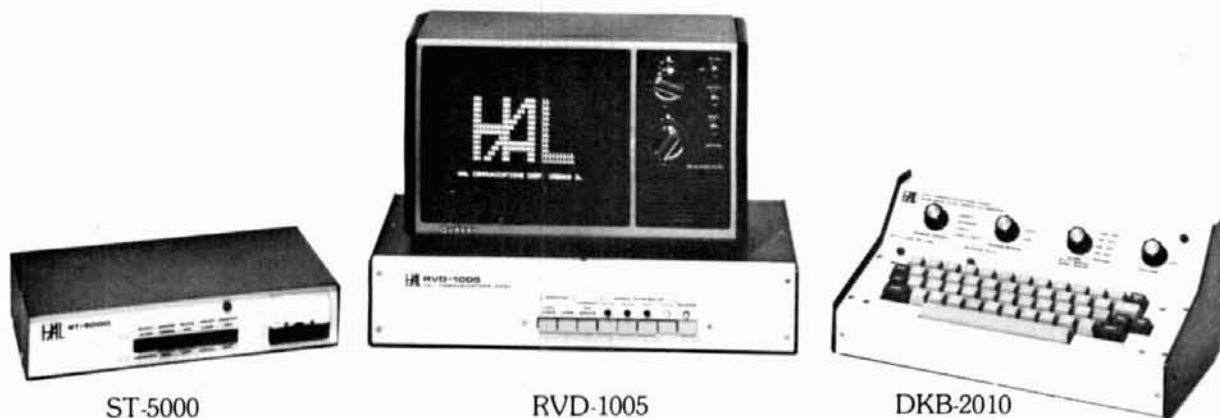
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DKB-2010

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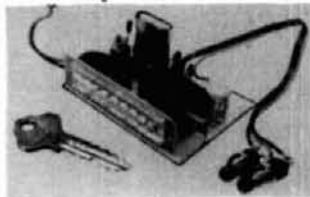
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21.50

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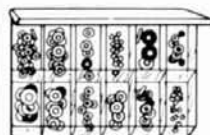
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flea market

Coming Events

CUYAHOGA FALLS AMATEUR RADIO CLUB — 24th Annual Electronic Equipment Auction and Flea Market, 9AM to 4PM, Sunday, February 26th at North High School, Akron, Ohio. Tickets \$1.50 advance, \$2.50 at door. Bring own tables, some available for \$1 each. Refreshments, prizes — grand prize Triton IV. Over 32,000 square feet for buyers and sellers, easy access on Tallmadge Avenue off-ramp, North Expressway (Route 8); connected to major interstates and Ohio Turnpike. Check in on 146.52 and 223.5 simplex, or 146.04/.64 and 147.84/.24 repeater. Details from CFARC, P.O. Box 6, Cuyahoga Falls, Ohio 44222.

COLUMBUS AMATEUR RADIO CLUB Annual Hamfest April 8-9, 1978, Columbus Municipal Auditorium Fairgrounds. Spacious, air-conditioned exhibit area, prizes, flea market, Saturday night banquet, FCC exams, and a luncheon at the Hamfest site. Contact Eddie Kosobucki, K4JNL, 5525 Perry Ave., Columbus, GA 31904.

MANSFIELD MID-WINTER HAMFEST — Auction — February 5, 1978 Richland County Fairgrounds, Mansfield, Ohio. Prizes, Flea Market, Auction. Large heated buildings. Doors open 8AM. Talk-in 146.34/146.94. Tickets \$1.50 in advance, \$2.00 at the door. Contact Harry Friezhen, K8HF (K8JPF) 120 Homewood, Mansfield, Ohio 44906 or phone 419-529-2801 or 419-524-1441.

MICHIGAN — Livonia Amateur Radio Club's 8th annual Swap 'n Shop, Sunday, February 26, 1978, 8:00 A.M. to 4:00 P.M., at the Stevenson High School, Livonia, Michigan. Plenty of tables, door prizes, refreshments, and free parking available. Talk-in on 146.52 Simplex. Write Neil Coffin WA8GWL, P.O. Box 2111, Livonia, Michigan 48150.

17th ANNUAL MICHIGAN CROSSROADS HAMFEST Saturday 3/4/78 8:00 opening Marshall High School, Exit 110 from I-94 near I-69. Over \$300 in door prizes. Check in 146.07/67 146.52 for lucky QSL card. Donation \$1.50 advance, \$2.00 at door. Table donation 50¢ each foot. Contact K8UCQ, Goodrich, 110 Perrett, Marshall, MI 49068, (616) 781-3554.

ANNUAL DAVENPORT RADIO AMATEUR CLUB HAMFEST, Sunday Feb. 26, 1978 at the Masonic Temple in Davenport, Iowa. Admission: \$2.00 advance, \$2.50 at door. Talk-in on 28/88 and 52 simplex. Refreshments and tables available. For info and tickets send S.A.S.E. to John Birmingham, WB0QCC, 2022 Brown, Davenport, Iowa 52804.

STERLING-ROCK FALLS AMATEUR RADIO Society Hamfest March 5, 1978, Sterling High School Field House, 1608 4th Avenue, Sterling, Illinois. Indoor flea market restricted to radio and electronic items only. Tables obtained at door, or bring your own. (\$3.00 for 1/2 table, \$6.00 for full table). Free parking available, including campers and trailers. Admission: \$1.50 advance, \$2.00 after Feb. 15th, 1978 or at the door. Write — Don Van Sant, WA9PBS, 1104 5th Avenue, Rock Falls, IL 61071. Make checks payable to Sterling-Rock Falls Amateur Radio Society. Talk in 146.94 simplex.

NEW HAMPSHIRE QSO PARTY — 2000Z February 10 to 0500Z February 11, and 1400Z February 11 to 0200Z February 12. Stations may be worked once per band per mode. New Hampshire stations send RS(T) and county; other send RS(T), ARRL section or country. Suggested frequencies: CW 1810, 3555, 7055, 14055, 21055, 28130 kHz; Phone 1820, 3935, 3975, 7235, 14280, 21380, 28575 kHz; Novice 3730, 7130, 21130, 28130 kHz; and VHF 50.115, 145.015 MHz (simplex only). SASE to Concord Brasspounders, Inc., 9 Via Tranquilla, Concord, New Hampshire 03301. Mailing deadline March 15, 1978.

MECKLENBURG AMATEUR RADIO SOCIETY — 1978 ARRL-sanctioned Hamfest, April 1st & 2nd, 1978, Charlotte Civic Center; plenty of parking available. Details from W4BFB, 2425 Park Road, Room 023, Charlotte, NC 28203.

Stolen Equipment

REGENCY HR2B Registration Number 2200-363 Engraved on the left front side and the speaker terminal strip is replaced with a mini-plug. Transceiver is bracketed to an AR-2 Regency Amplifier. If located notify the Sandusky Police Dept., Sandusky, OH 44870 or Call Earl Carrier, K8WCP collect 1-419-625-1817.

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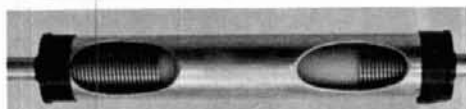
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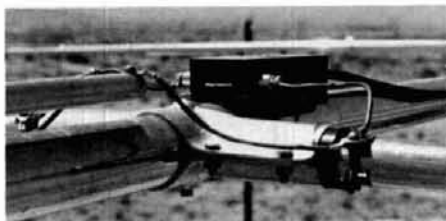
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Longest Element 26' 7"
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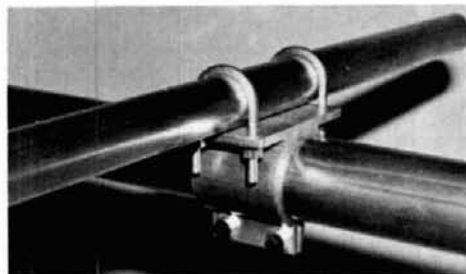


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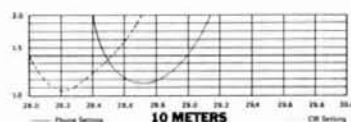
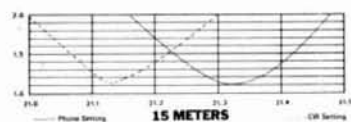
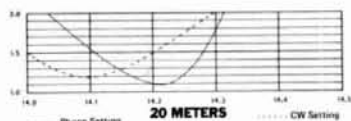


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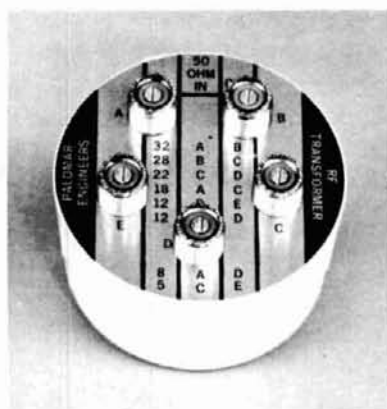
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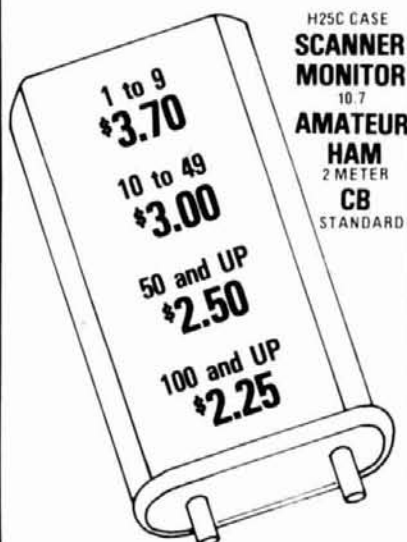
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	100/6 90	7408	10/1 70	3401 8	2/1 90
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1N4007	20/1 90	7425	10/2 50	3401 18	2/1 90
	100/8 50	7430	10/1 50	3401 24	2/1 90
2N3055	2/1 65	7432	10/2 30	NE555V	10/3 40
2N3904	8/1 00	7438	10/2 00	565CN	5/4 75
2N3906	8/1 00	7441	10/7 00	566	1 75
CD	Quantity:	7442	10/3 70	567	1 25
CMOS 1C	Price	7447	10/7 90	709 T05	10/2 50
4001	10/1 60	7453	10/1 50	723 DP	10/4 00
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4017	5/3 50	7476	10/2 60	741 MIN DP	10/2 50
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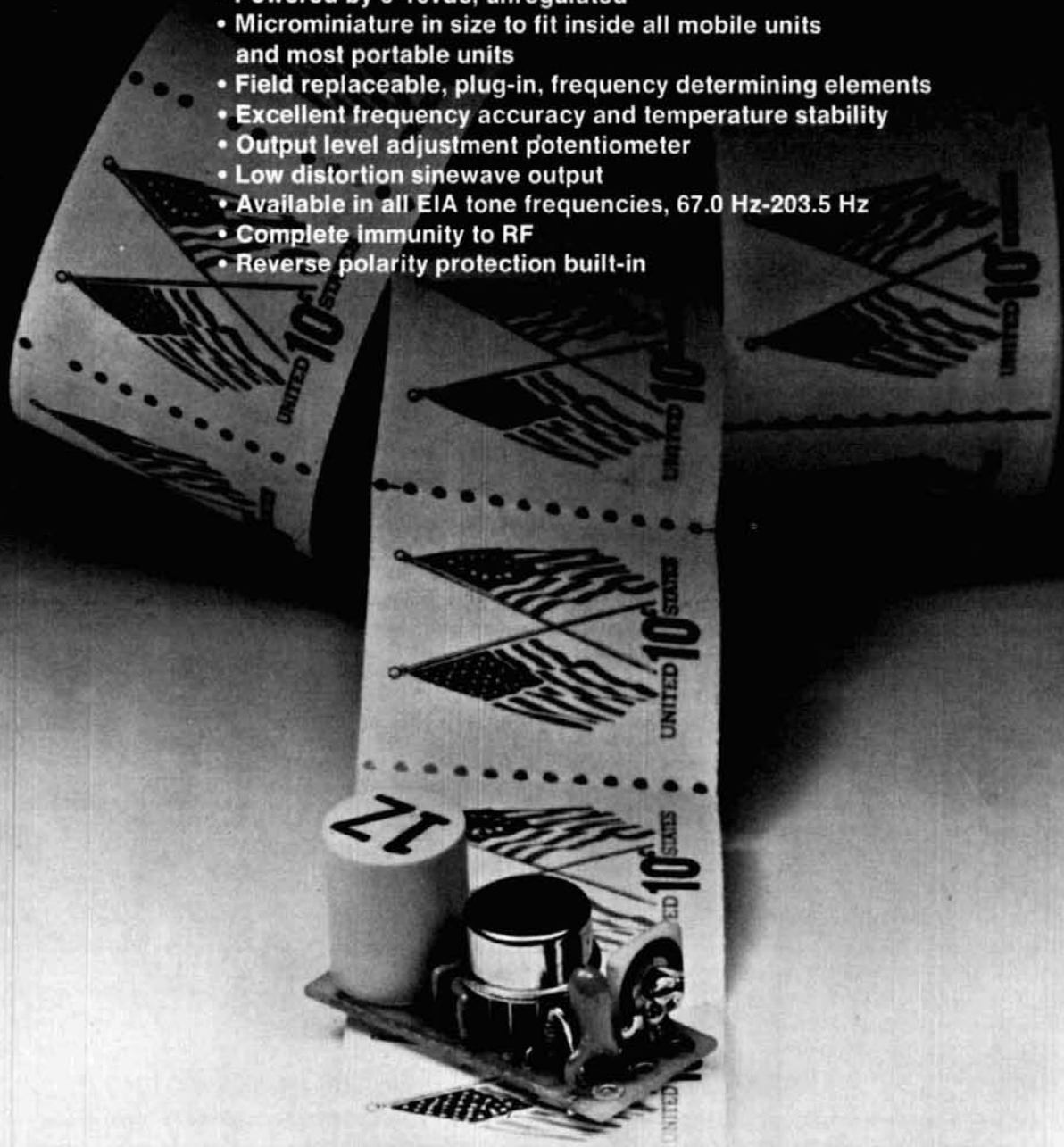
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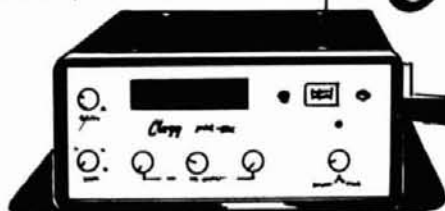


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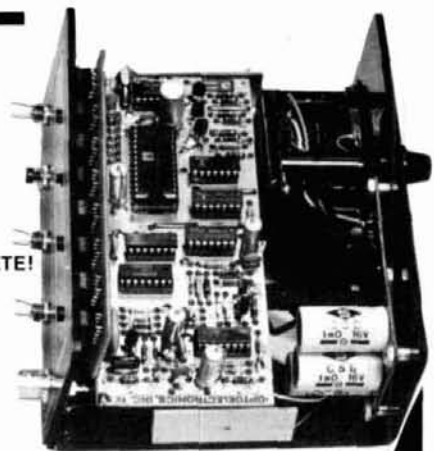
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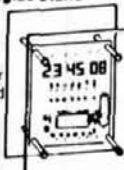
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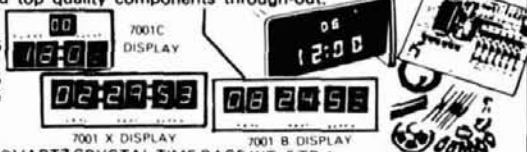
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50 watts	50A	50C	50D	50E	50F	
100 watts	100A	100C	100D	100E	100F	
250 watts	250A	250C	250D	250E	250F	
500 watts	500A	500C	500D	500E	500F	
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1N758A	10v	z	.25
1N759A	12v	z	.25
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4018	1.10
4019	.60
4020	.85
4021	1.35
4022	.95
4023	.25
4024	.75
4025	.35
4026	1.95
4027	.50
4028	.95
4030	.35
4033	1.50
4034	2.45
4035	1.25
4040	1.35
4041	.69
4042	.95
4043	.95
4044	.95
4046	1.75
4049	.70
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7410	.10
7411	.25
7412	.30
7413	.45
7414	1.10
7416	.25
7417	.40
7420	.15
7426	.30
7427	.45
7430	.15
7432	.30
7437	.35
7438	.35
7440	.25
7441	1.15
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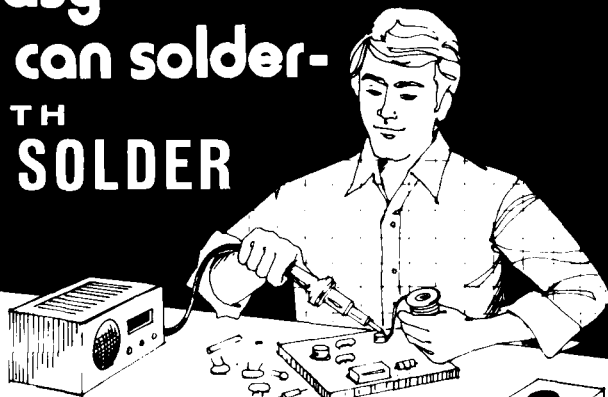
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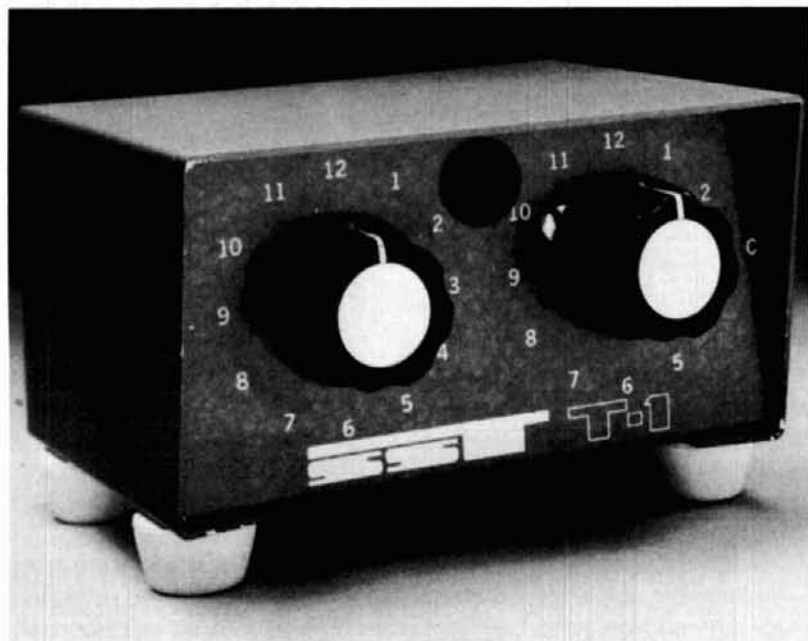
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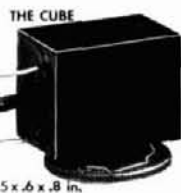


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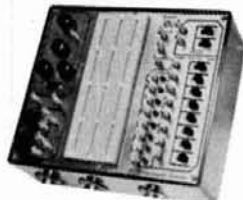
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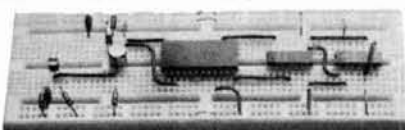
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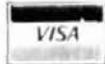
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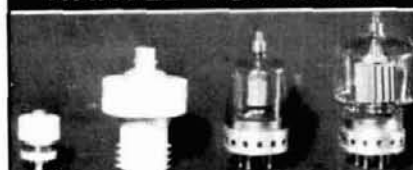
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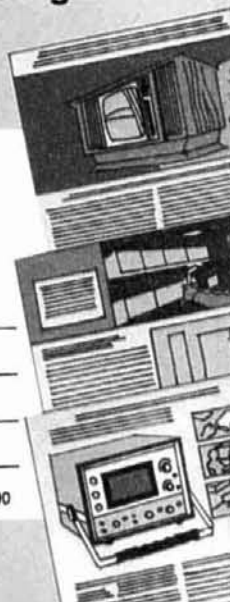
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