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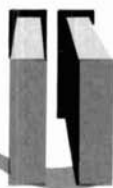
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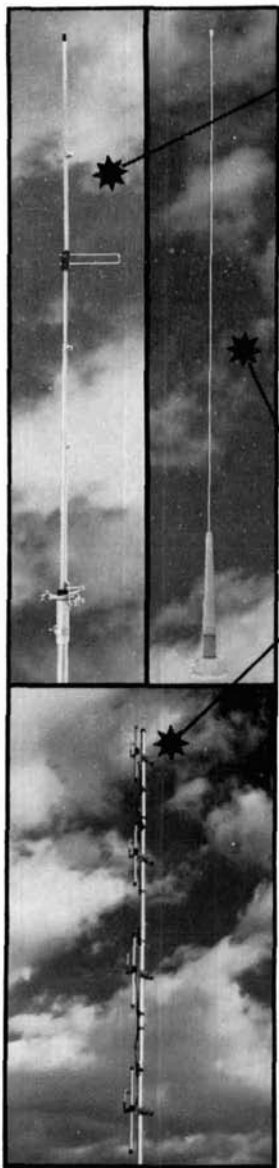
electronic keyer

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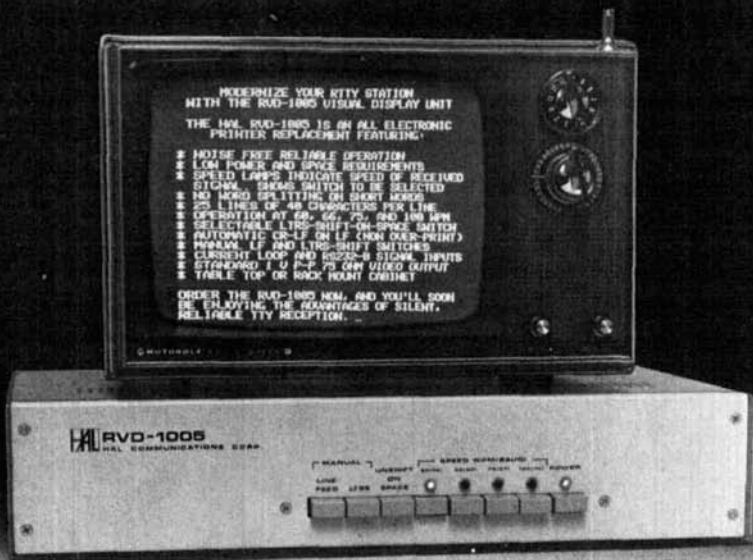
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April, 1975
volume 8, number 4

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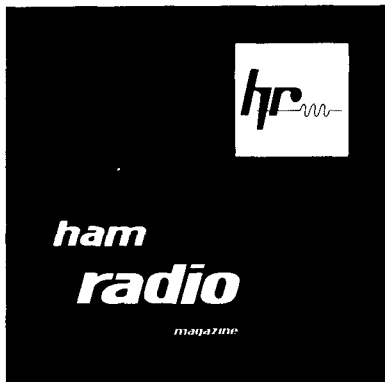
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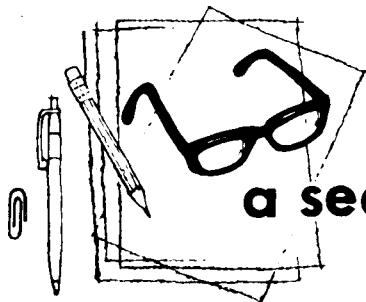
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a second look

by jim
fisk

During the past few months more and more vhf stations have been getting set up for serious moonbounce work. A good share of this activity has been in progress for months, but the February EME tests by WA6LET with SRI's big 150-foot dish in Palo Alto was the stimulant many vhf operators needed to become fully operational. Some operators, particularly in the northern areas, were hampered by the winter weather, but the seed has been planted so the number of vhf EME stations will probably double by the end of the summer.

Five years ago there were but a handful of successful EME stations — only the most serious and persevering workers ever made the grade. Those who did make it developed the techniques that are used today. And today equipment is available so that just about anyone who is interested can hear his own echoes off the lunar surface.

Right after World War II, when the first radio signals were bounced off the moon by engineers at Fort Monmouth, it was not nearly so simple: 8 kW input on 111 MHz, a tremendous billboard array of 64 phased dipoles and an incredibly complex receiver with a 50-Hz passband. Even then, the moon echoes were weak, and success unpredictable.

With this kind of background, the prospects of amateur communications via the moon were pretty remote, but W3KGP and W4AO launched Project Moonbeam in the late 1940s with a goal of hearing their own two-meter echoes from the moon. They finally heard them in 1953.

Until parametric amplifiers became activity was confined to two meters. EME attempts on 432 MHz were impractical because of the power limitation that was in effect, so the next logical step was 1296 MHz. In 1960 W1BU

came on the air with a kilowatt 1296-MHz station that could bounce signals off the moon with some degree of reliability. Sam Harris, W1FZJ, extended the challenge and Hank Brown, W6HB, picked it up. Shortly thereafter the first *two-way* amateur contact via the moon's surface was history.

Shortly thereafter the first *two-way* amateur contact via the moon's surface was history.

Since 1960 progress has been steady if slow. Although all the vhf bands from 50 to 2300 MHz have been used for two-way EME contacts, at the present time most of the activity is found on 144 and 432 MHz. The handful of stations on two-meter EME a few years ago has grown to nearly 100, and there are about 35 stations operational on 432 MHz. Nor is activity confined to the United States. There are successful EME stations on every continent except Asia, and serious interest has been expressed by several Japanese amateurs, so Asia may be represented in the near future, possibly this summer.

EME is still a very sophisticated method of vhf communications, but with the easy availability of low-noise, solid-state converters and high-efficiency kilowatt amplifiers, it is within the grasp of any serious vhf'er. Most of the successful moonbouncers live on city-size lots, so space is not a problem, and many use all commercial equipment, so gear is no problem. Most vhf stations capable of long-distance tropo, scatter or auroral communications, in fact, are also capable of two-way EME. It has taken twenty-five years, but the 1950's pie-in-the-sky dream of reliable two-way EME communications has become a reality.

Jim Fisk, W1DTY
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DECISION ON 220-MHz CLASS-E CB is likely to be put off again. We had hoped for delay until after both amateur and CB restructuring go into effect and their results have a chance to be felt -- we'll probably get a delay that will permit the FCC to study all three dockets together. Any delay we do get will give us a chance to further strengthen our position, increasing the likelihood of preserving 220-225 intact for the amateur service.

WA6LET MOONBOUNCE TESTS brought many vhf stations out of the woodwork, made 53 two-way EME contacts on 144 MHz with 36 different stations during eleven hours of operating time. Score: 14 states and 7 countries on two meters. Most of the QSOs were on CW though five stations made it on ssb. In addition to stations in Germany, France, Sweden and the Netherlands, they worked ZE1DX in Rhodesia for the first U.S.-to-Africa EME contact.

432 MHz EME Test was shortened to two hours due to problems with antenna relays and an arcing HV power supply, but the WA6LET operators still managed to work ten stations via the moon on 432 including PAØSSB and ZE5JJ. K2UYH was the only station to make it both ways on ssb. More 432-MHz EME tests are expected later this year from WA6LET.

AUTOMATIC TRANSMITTER IDENTIFICATION SYSTEM is proposed by the FCC in Docket 20351 released February 13. ATIS requirement would apply to all transmitters licensed in Safety and Special Services produced after one year after adoption and operating between 25 and 960 MHz, with the exception of amateur radio.

ATIS Signal would be ASCII and give the station call sign at the beginning, end and every 30 seconds during a transmission. First or second class license would be required to program the encoder.

AMSAT REQUESTS U.S. DXER HELP in setting up interested DX stations for OSCAR 7, Mode B (432 MHz in, 144 MHz out) operation. AMSAT has acquired a number of uhf transmitters and has them tuned up and ready to go on 432 MHz, will provide one on a loan basis to anyone willing to take the responsibility of getting it into the hands of an overseas amateur interested enough to put it on the air. The DX station would still have to supply his own antenna and two-meter receiver.

OSCAR 7's Mode B has already seen some exciting DX -- both ZL and UAØ have been reported worked from the midwest, and WB6NMT reports hearing over-the-horizon signals several times.

Reports Of "Tandem Satellite" QSOs using both O6 and O7 as their orbits coincided have been received by AMSAT, but more inputs are needed from those who have made contacts using the two transponders in tandem. Send QSO data to AMSAT, Box 27, Washington, D.C. 20044.

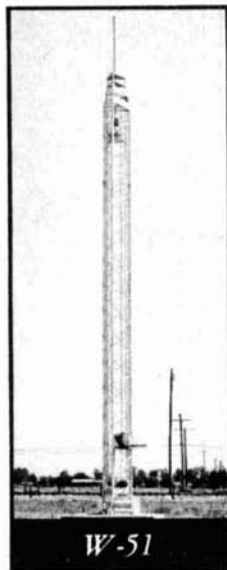
ANOTHER 10-METER BEACON will help 10-meter DXers through the predicted flat conditions ahead. The NZART has put ZL2MHF on 28170 kHz with 90 watts input and a half-wave vertical. ID is in CW every 10 seconds or so. Reports of reception are requested to NZART Upper Hutt Branch 63 Inc., Box 40212, Upper Hutt, New Zealand.

Proposed San Diego Beacon, widely reported elsewhere as on the air last fall, has still not been okayed by FCC. FCC request for further data back in October still has not been answered.

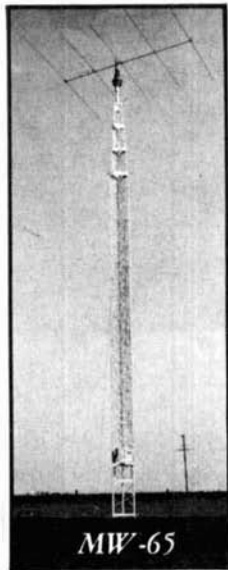
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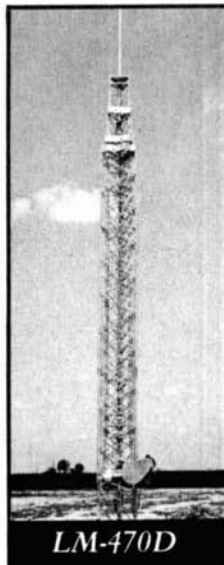
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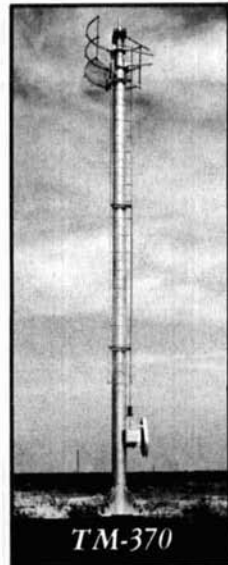
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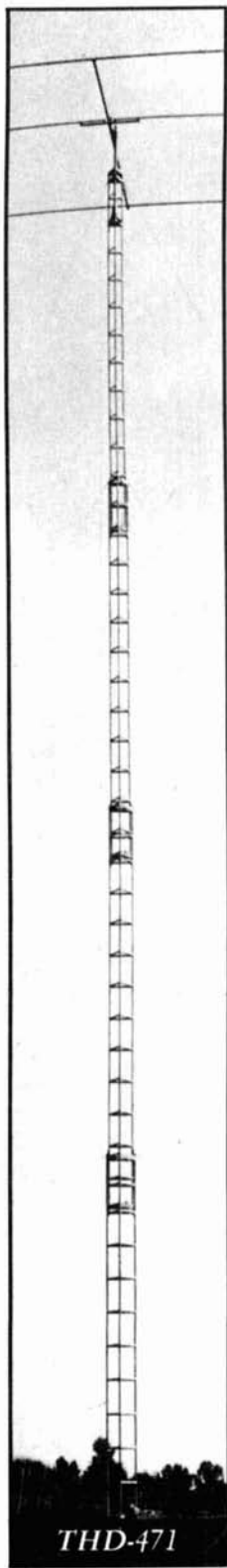
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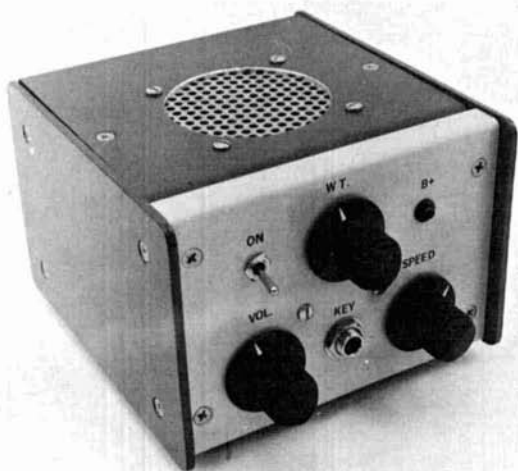
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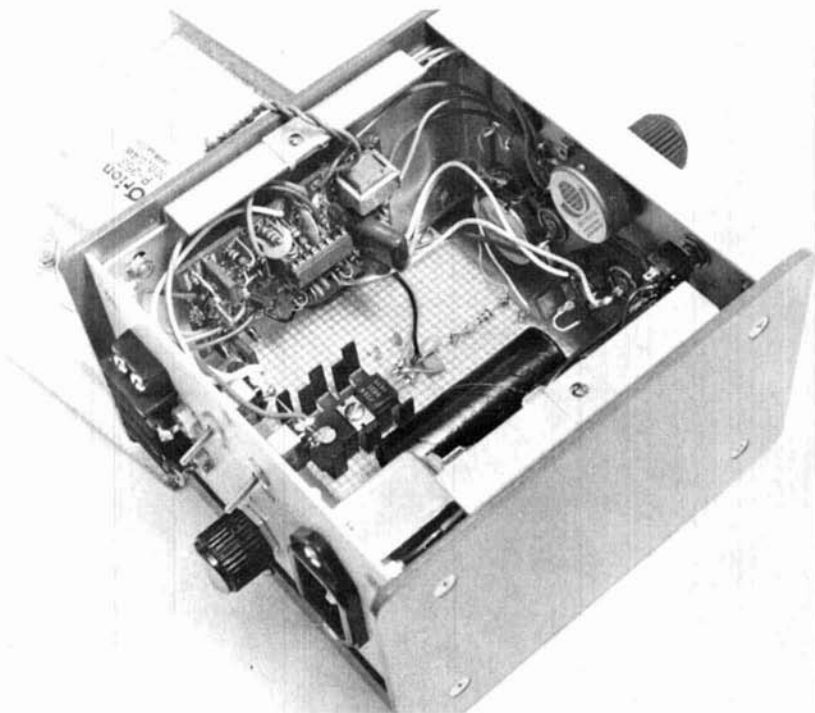
Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California

Any who has built one of the many versions of IC keyers described in the amateur radio magazines probably has been less than satisfied with its performance in the shack. Many circuits use RTL as the basic logic form, but RTL has inherently poor noise immunity and requires large supply currents. Some newer designs use TTL — ICs which have improved noise immunity, but are still in the current-guzzler category. In virtually all the circuits that I have built or even thought about building (including a couple of my own design) there have been "glitches" in the designs. These have shown up in actual use by operators, and are subtle problems that can be difficult for a designer to anticipate.^{1,2}

The Curtis 8043 is a new IC developed by an engineer who has been in the electronic keyer business for years, and who has also been employed by one of the largest IC manufacturers in the United States. Such a unique combination of experience should make the Curtis 8043 a nearly ideal IC around which to build your own keyer, and this has been proven by the keyer described here. First of all, the 8043 device uses

P355 cabinet with all input and output leads decoupled to prevent rf from getting into the circuitry. The 8043 is presumably comparatively rf-proof, but I have found that it's much easier to install RFI-proofing as I build, rather

Curtis is shown in **fig. 1**; asterisk-marked items are builder-supplied. The power-supply circuit is shown in **fig. 2**. Note that there are three jumper wires on the Curtis board; two are associated with Q1 and one with the emitters of Q3 and



Construction of the electronic keyer. The 8043 IC is installed on the small printed-circuit board on the side wall of the enclosure. The regulated power supply is built on the perf board at the bottom of the chassis.

than to try to add it after the fact. The ac line is decoupled using a Corcom 6EF1 line filter; the keyer contacts and transmitter keying line are decoupled with 1000-pF feedthrough or standoff capacitors and ferrite beads. These RFI measures are not absolutely necessary, but probably contribute to the lack of trouble from rf that the unit enjoys.

circuit

The keyer circuit as furnished by

Q4, transistors used as switches to replace the relay commonly found in electronic keyers for keying the transmitter. The jumpers determine which transistor (Q3 or Q4) is used.

In this service keying transistors are better than relays because they have no contact bounce and no contacts to wear. However, the transistor switches do not have the complete dc isolation of a pair of relay contacts, so an npn device must be used for keying trans-

mitters having a positive key-up voltage (such as cathode keying) and a pnp device must be used for keying transmitters having a negative key-up voltage. Both Q3 and Q4 will stand up to 300 volts in the open-key condition.

As can be seen in the photographs of the IC keyer, the *speed*, *volume* and *weight* controls are mounted on the front panel with the ac switch, LED pilot light and key jack. The *pitch* control is mounted on the rear apron of the

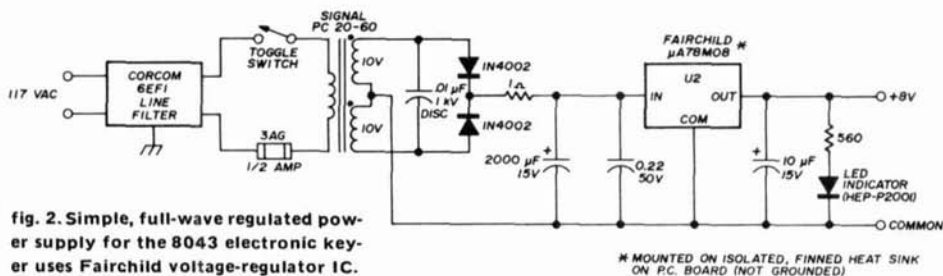


fig. 2. Simple, full-wave regulated power supply for the 8043 electronic keyer uses Fairchild voltage-regulator IC.

In addition to the line switch, there are two other miniature toggle switches, one labeled *tune*, the other *self-test*. The *tune* switch turns the transmitter on for tuning up, and the *self-test* feature allows you to operate the keyer (*listening* to its output) without keying the transmitter. If straight-key operation is desired, the key may be connected (with a jack) in parallel with the *tune* switch.

keyer along with the *tune* and *self-test* switches.

The 1-meg pot referred to in fig. 1 as *dot-space symmetry adjust* is a small board-mounted trimpot. This *symmetry* pot is the only adjustment necessary after the keyer has been built. Starting with the *symmetry* pot at the middle of its adjustment, and with the *weight* control at minimum and the keyer paddle on dots, the *symmetry* control is adjusted until the dot and space periods are equal, as displayed on an oscilloscope.

Rear view of the electronic keyer showing the tune and test switches, pitch control, and terminal strip for transmitter connections. Plug on the right is for the ac line. Monitor speaker is on top.

summary

The keyer built around the Curtis 8043 semi-kit has proven to be everything it claimed to be, satisfying all comers at a local amateur radio club. No signs of RFI have been evident at the club station nor at other 1000-watt amateur stations. If you've been thinking about building a keyer, the 8043 is clearly the way to go!

references

1. Howard Berlin, K3NEZ, "Memory Keyer," (letter), *ham radio*, December, 1974, page 58.
2. F. E. Hinkle, "Heath HW-7 Modifications," *ham radio*, December, 1974, page 60.

ham radio



microstripline preamplifiers

for 1296 MHz

Complete design and
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for low noise,
solid-state preamplifiers
for the amateur
23cm band

In a previous article I described a low-cost transceive converter for the amateur 1296-MHz band.¹ Calculated parameters and preliminary measurements indicated a receive system noise figure of 7.5 dB. As is often the case, these original claims have proved overly optimistic. Subsequent measurements with an argon-discharge noise source and an automatic noise-figure meter showed the true ssb noise figure to be more on the order of 9.5 to 10 dB. Neither improvements in the homebrew balanced mixer nor the use of high-grade commercial mixers significantly altered this figure.

It became apparent that optimum performance would necessitate the use of one or more low-noise preamplifier stages preceding the diode balanced

mixer. Several popular amplifier circuits^{2,3,4} were tried and the results were entirely satisfactory. However, I wanted to depart from the conventional technology (pi-network input and output with slab inductors) used in these designs. Since I had obtained considerable success with microstriplines in a family of transmit linear amplifiers, I decided to build a family of receive preamps using the same basic design techniques. The resulting circuits, presented here, represent a reasonable tradeoff between cost, performance, simplicity and reproducibility.

system considerations

It was assumed that these amplifiers would precede a receive converter with a ssb noise figure on the order of 10 dB. The Simple Sideband System¹ meets this criteria, as do most properly adjusted trough-line converters,⁵ and at least one popular commercial unit.*

A workable rule of thumb is the principle that, if the gain preceding a receive converter is at least 10 dB greater than the converter's noise figure, then the noise figure of the total system will, for all intents and purposes, equal the noise figure of the preamplifier. Thus, 20 dB of preamplifier gain preceding a 10-dB NF converter will have the effect of masking the converter's noise.

A number of readily available microwave transistors are capable of 2- to 3-dB noise figures in the 1296-MHz

*Spectrum International MM_C 1296, \$85.95 from Spectrum International, Box 1084, Concord, Massachusetts 01742.

Paul Shuch, WA6UAM, 14908 Sandy Lane, San Jose, California 95124

amateur band. These devices offer conservatively rated power gains on the order of 10 dB per stage so if two such stages of preamplification precede the 10-dB NF converter, an overall noise figure of 2 to 3 dB will result.

When receive preamplifiers are connected in cascade, the first stage should be adjusted for optimum noise figure; subsequent stages are adjusted for optimum power gain. In practice the only difference lies in the input matching circuit. When a power-match is desired, the input circuit presents a complex conjugate match to the transistor's input impedance. In the case of a noise-match, a predetermined mismatch is introduced into the input circuit to minimize the stage's noise figure. In both cases the transistor's collector should look into a complex conjugate match.

microstripline considerations

The exact details of a microstripline design vary with the material used for the substrate. Such dielectrics as Teflon, Rexolite, Duroid, and glass offer superior performance at microwave frequencies. The material most readily available to the experimenter, however, is fiberglass-epoxy printed-circuit board. After having designed and built more than a dozen preamplifiers of the type described here, on a variety of substrates, I can state conclusively that the degradation in performance resulting from building on lowly glass-epoxy board is beyond the measurement capabilities of the average experimenter. Thus the amplifiers presented here were designed to be etched onto 1/16-inch (1.5mm) thick G-10 PC board, double-clad with 1-ounce copper.

It is possible to design an amplifier using microstriplines that requires no external tuning — all the resistive and reactive matching elements are provided by the microstriplines. Such designs

have been published previously, and if properly built, will provide excellent performance without the need for tuning adjustments. However, amplifiers without tuning adjustments are practically an affront to the amateur spirit. Though considered frivolous by some, my amplifiers include trimmer capacitors. In addition to giving the dyed-in-the-wool experimenter something to tweak, these adjustable components provide some degree of compensation for slight variations between transistors, as well as variations from one PC board to the next.

preamplifier transistors

The second-stage amplifier is built around a Hewlett-Packard 35826E, a low-cost version of the well known HP-21 family. An acceptable substitute is the VO21 manufactured by the Nippon Electric Company, a second-source device which performs identically in the circuit. At the time of this writing, both transistors sell for \$17.50 each in single quantities.*

The device used in the optimum-noise-matched first stage depends on the needs and budget of the individual builder. The design presented here uses a Hewlett-Packard 35866E option 100, one of the least expensive members of the low-noise HP-22 family. Priced at \$45, it provides an overall system noise figure of 2 dB, challenging the best parametric amplifiers of yesteryear. Unless you anticipate EME or long-haul troposcatter communications, you will probably find it more cost-effective to use the lower priced HP-21 or VO21 in

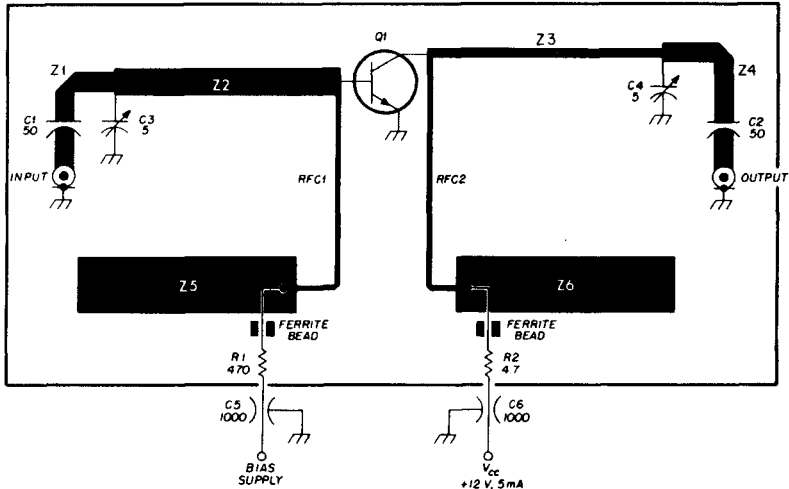
*For the address of your nearest Hewlett-Packard distributor, look in the *Yellow Pages* or write to HPA Division, 540 Page Mill Road, Palo Alto, California 94304. Nippon Electric Company semiconductors are distributed in this country by California Eastern Labs, Inc., 1 Edwards Court, Burlingame, California 94010.

the first stage as well. These devices, when tuned for optimum noise figure in the circuit shown in fig. 1, will yield an overall system noise figure on the order of 3 dB.

The cost factor can be easily analyzed in terms of dollar expenditure per dB improvement in signal-to-noise ratio.

but at an added cost of \$27.50. The overall cost-effectiveness of the two-stage preamplifier now becomes \$7.81 per dB — not an unreasonable figure when maximum performance is required.

Since it is common practice for microwave semiconductor manufac-



C1,C2 50 pF chip capacitor (ATC 100 or equivalent)

C3,C4 1 to 5 pF precision piston trimmer (Johannson JMC 4642)

C5,C6 1000 pF feedthrough capacitor

Q1 Hewlett-Packard 35866E, option 100 preferred (HP 35826E or NEC VO21 acceptable)

R1 470 ohm, 1/4 watt carbon composition

R2 4.7 ohm, 1/4 watt, carbon composition

RFC1 100 ohm, quarter-wavelength microstripline, 0.02" (0.5mm) wide, 1.25 inch (32mm) long

Z1,Z4 50 ohm microstripline, 0.1" (2.5mm) wide, any convenient length

Z2 41.7 ohm, quarter-wavelength microstripline, 0.14" (3.5mm) wide, 1.18" (30mm) long

Z3 75 ohm, quarter-wavelength microstripline, 0.04" (1mm) wide, 1.22" (31mm) long

Z5,Z6 Rf short. 25 ohm, quarter-wavelength open-circuited microstripline, 0.30" (7.5mm) wide, (32mm) long

fig. 1. Circuit for the noise-matched 1296-MHz preamplifier stage. Printed-circuit layout is shown in fig. 2.

For example, two stages of HP-21 or VO21 at \$17.50 per transistor will improve a 10-dB NF converter by 7 dB at a device cost of \$5.00 per dB improvement. With the higher performance HP-22 in the front end, an additional dB of sensitivity can be achieved,

turers to produce a family of transistors by mounting the same chip in a variety of packages, it is often possible to substitute a different transistor of the same family without significantly influencing amplifier performance.

The part numbering system used at

Hewlett-Packard provides some insight into the interchangeability of their devices. A system of five digits and a one-letter suffix is used, the first three digits being 358. The fourth digit position indicates the device family, with the number 6 designating HP-22 type devices (low noise microwave transis-

packaged chips carry the suffix A. The HP 35821E, for example, is a general-purpose (HP-21) microwave transistor mounted in a 200-mil (5mm) round strip package in the common-emitter configuration.

Obviously, differently packaged versions of the same semiconductor chip

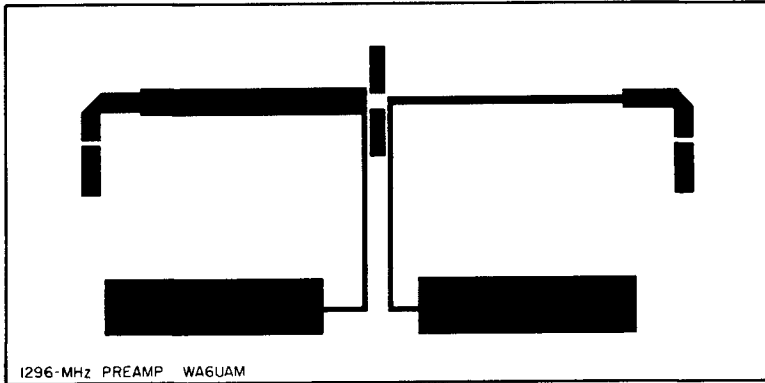


fig. 2. Full-size printed-circuit layout for the noise-matched 1296-MHz preamplifier shown in fig. 1.

tors), the number 2 referring to HP-21 type transistors (for general-purpose microwave applications), and the numbers 3 and 5 indicating the linear power transistors of the HP-11 and HP-12 families.

The fifth digit of the Hewlett-Packard part number indicates the type of package in which the semiconductor chip is mounted. Among the strip-packs, a number 2 indicates a 70-mil (1.8mm) diameter; 1, a 200-mil (5mm) diameter; and 6, a 100-mil (2.5mm) square package. The number 4 refers to a metal TO-72 package, and 7 is a coaxial package. Grounded-stud packages are designated by a 4, and 3 indicates a grounded-bar configuration. Unpackaged chips are coded zero. The letter suffix indicates whether the device is mounted in common-emitter (E) or common-base (B) configuration. Un-

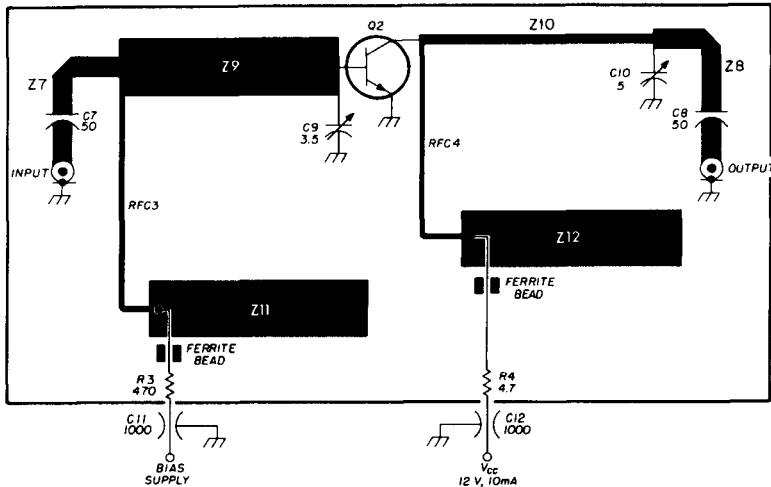
will differ from one to another in terms of their complex input and output impedances. However, at low frequencies (and for microwave devices 1.3 GHz can be considered low), the contribution of package parasitics to the overall input (S_{11}) and output (S_{22}) S-parameter values can be considered small in relation to the characteristics of the chip itself.* These amplifiers, though designed around the characteristics of the 100-mil (2.5mm) square package, function well with like transistors in the 70-mil (1.8mm) and 200-mil

*The S-parameters or scattering parameters are one of several methods used for describing the input-output characteristics of transistors. The basic difference between S-parameters and other systems are that the S-parameters are measured with transmission-line inputs and outputs; this greatly simplifies measurements at uhf and microwave.

(5mm) round packages with only minor degradation in input and output vswr.

An analogous situation exists with Nippon Electric Company transistors, although NEC lacks the elaborate numbering system which identifies the HP

with this newer transistor in the existing boards, performance was wholly satisfactory. As a matter of fact, the VO21 in the 100-mil (2.5mm) square package exhibited 0.2 dB lower noise figure than the original device.



- C7,C8 50 pF chip capacitor (ATC 100 or equivalent)
- C9 0.35 to 3.5 pF precision piston trimmer (Johannson JMC 5801)
- C10 1 to 5 pF precision piston trimmer (Johannson JMC 4642)
- C11, C12 1000 pF feedthrough capacitor
- Q2 Hewlett-Packard 35826E or NEC VO21
- R3 470 ohm, 1/4 watt carbon composition
- R4 4.7 ohm, 1/4 watt carbon composition

- RFC3 100 ohm, quarter-wavelength micro-RFC4 stripline, 0.02" (0.5mm) wide, 1.25" (32mm) long
- Z7,Z8 50 ohm microstripline, 0.10" (2.5mm) wide, any convenient length
- Z9 25 ohm, quarter-wavelength microstripline, 0.30" (7.5mm) wide, 1.14" (29mm) long
- Z10 75 ohm, quarter-wavelength microstripline, 0.04" (1mm) wide, 1.23" (31mm) long
- Z11, Rf short. 25 ohm, quarter-wavelength
- Z12 open-circuited microstripline, 0.30" (7.5mm) wide, 1.14" (29mm) long

fig. 3. Circuit for gain-matched 1296-MHz amplifier stage. Printed-circuit layout for this amplifier is shown in fig. 4.

devices. The NEC VO21 was initially manufactured in a 150-mil (3.8mm) round package, and it was the scattering parameters for that device which were used in the design of these amplifier boards. Recently NEC introduced a VO21 chip mounted in a 100-mil (2.5mm) square package. Although perfect matching could not be achieved

When ordering the NEC VO21 transistor, specify the 320 package if a 150-mil (3.8mm) round package is desired. For the 100-mil (2.5mm) square, request package ML-3. There is no price difference.

construction

Figs. 1 and 3 are functional sche-

matics of the amplifier stages. Microstriplines opposite a groundplane (the unetched side of the double-clad printed-circuit board) comprise all matching transformers, rf chokes and rf bypasses. Figs. 2 and 4 are full-size printed-circuit layouts for the noise-matched and power-matched stages, respectively. Note that dimensions are applicable only to 1/16-inch (1.5mm) G-10 double-sided glass-epoxy printed-circuit board.

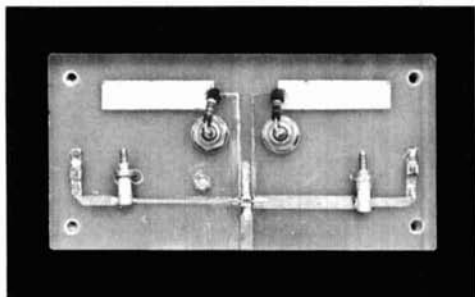
Mounting of the transistors is detailed in fig. 5. A hole is drilled in the PC board to allow direct strapping of the emitter traces to the groundplane. It is desirable to cover this hole on the groundplane side after the transistor is installed, to improve shielding and furnish physical protection for the transistor.

A word of caution is in order regarding mounting the transistors to the boards. There is a difference in lead layout between the HP and NEC transistors which can cause considerable confusion. The emitter leads of both devices are the wider pair of leads. One of the two remaining leads is tapered 45° at the end. The slashed lead is the *base* of the NEC devices, but the *collector* of the HP transistors. This small difference cost me two expensive transistor failures while developing the prototype amplifiers for this article.

The grounded (adjusting screw) end of the trimmer capacitors must be connected directly through the board to the groundplane to minimize tuning-tool interactions when making adjustments. Although concentric-ring piston trimmers (Johannson, JFD, etcetera) are ideal for this purpose, performance of the lower cost ceramic piston trimmers used in uhf TV tuners is entirely satisfactory.

The dc blocks at the input (C1 and C7) and output (C2 and C8) should ideally be ceramic chip capacitors. If

these are not available, modified miniature disc ceramic capacitors are usable. An Xacto knife is used to scrape the insulation off of the sides of the capacitor, the leads are removed and, using the lowest possible soldering heat, the PC



Microstripline side of the noise-matched 1296-MHz preamplifier. The low-noise H-P 35866E transistor is at the top center of the board. The gain-matched stage is similar except for the width of the input quarter-wave transformer and placement of the input trimmer capacitor.

traces are bridge-soldered directly to the capacitor plates.

The prototype preamplifiers shown in the photographs use miniature SMA coaxial connectors, but the less expensive JCM connectors made by E. F. Johnson are satisfactory substitutes. TNC connectors may be used, too, but they are much larger. BNC connectors should be avoided as they will slightly degrade the noise-figure performance of the amplifier stages. Fig. 6 shows a method of modifying flange-type bulk-head connectors for use as microstripline launchers.

All power-supply leads for the collector supply and bias current are isolated from the rf circuitry by rf chokes, ferrite beads and bypass and feed-through capacitors. Therefore, the necessary bias circuitry can be installed on the groundplane side of the circuit board.

bias circuits

With the transistors specified, optimum noise figure occurs at a collector current of approximately 5 mA. A higher collector current will improve stage gain. In the conventional configuration of a low-noise first stage and higher-gain second stage, it is desirable to bias the two stages for collector currents of 5 and 10 mA, respectively.

Virtually any bias circuit which will maintain the desired collector current is acceptable. Many of the simpler resistive bias circuits should be avoided due to their low stability factor (that is, high dependence of collector current on transistor dc current gain) and the resulting danger of thermal runaway. To quote a useful Hewlett-Packard applications note, "Often the least considered factor in microwave transistor circuit design is the bias network. Considerable effort is spent in measuring S-parameters, calculating gain, and optimizing bandwidth and noise figure, while the same resistor topology is used to bias the transistor. Since the cost per dB of microwave gain or noise figure is so high, the circuit designer cannot afford to sacrifice rf performance by inattention to dc bias considerations."⁶

An active bias circuit (variable constant-current source) is desirable in

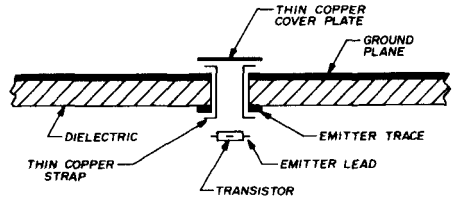


fig. 5. Method for mounting the transistors to the printed-circuit boards. Note thin copper strap used to connect the emitter circuit trace to the ground plane. Solder emitter leads as close as possible to the transistor package and use minimum soldering heat to avoid transistor damage.

that it affords a degree of protection for the transistor while permitting ready collector current adjustment for optimizing stage gain and/or noise figure. One such circuit, shown in fig. 7, furnishes a variable collector current of 2 to 12 mA, more or less independent of the dc current gain of the transistor being biased.

For initial amplifier tuneup, adjust the trimpot in the base of the bias transistor to produce 5.5 mA of total current in the first amplifier stage, and 10.5 mA in the second. (This accounts for a quiescent bias circuit current of 0.5 mA). Upon completion of all rf tuning, the trimpots may be adjusted to optimize the overall system noise figure.

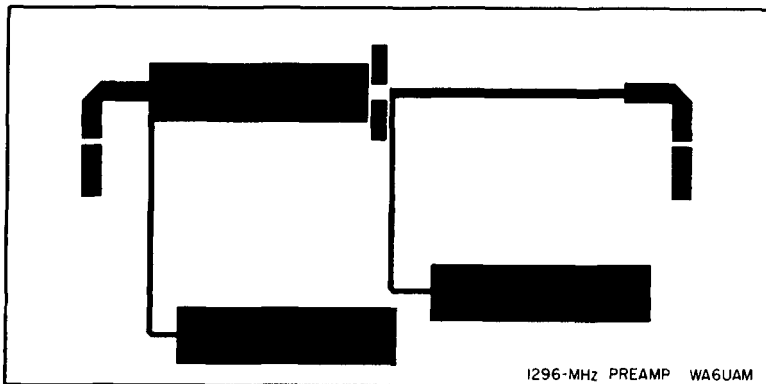


fig. 4. Full-size printed-circuit layout for the 1296-MHz gain-matched amplifier shown in fig. 3.

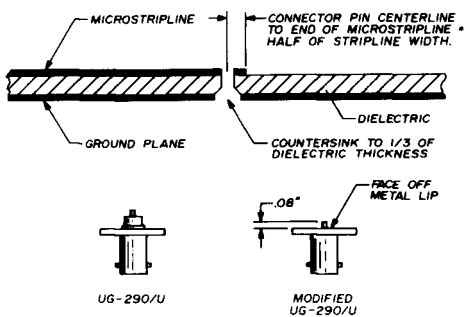


fig. 6. Modifying coaxial bulkhead connectors for use as microstripline launchers.

alignment

The two preamplifier stages are built as separate subassemblies and connected together with a short length of coaxial cable. The reasons for this modular approach are twofold. Most obvious is the fact that many operators may wish to add a single stage of preamplification initially, expanding their system later as needs and budget dictate. In this case it

is recommended that the power-gain matched stage (which uses the less expensive transistor) be built first.

Even if two stages of preamplification are built, tuneup and matching can be most readily accomplished if the stages are in separate modules. The power-matched stage is connected to the receiving converter first, its input terminated in 50 ohms, and the trimmer capacitors adjusted for maximum received signal from a beacon source.^{7,8,9} Once a power match is achieved in the second stage (both input and output), the first stage is connected and, using a weak-signal source, its *output circuit only* is adjusted for maximum received signal. A power match now exists between the two amplifiers as the output of the first stage and the input to the second stage are each matched to 50 ohms, and a 50-ohm coaxial cable connects the two.

The tuning process is completed by obtaining a proper noise match into the first stage. This is most readily accomplished by using an argon-discharge type noise source.^{10,11} Unfortunately, few experimenters have access to such equipment, except perhaps at regional uhf conferences.* A semiconductor diode noise source is an acceptable alternative.^{12,13} A number of articles have described the process of tuning an amplifier for minimum noise figure.¹⁴⁻¹⁸

When adjusting the input circuit of the first stage, an important consideration is the interactive nature of the input matching and bias current adjustments. Since the transistor's optimum source reflection coefficient, Γ_o (see page 24), varies with collector current,

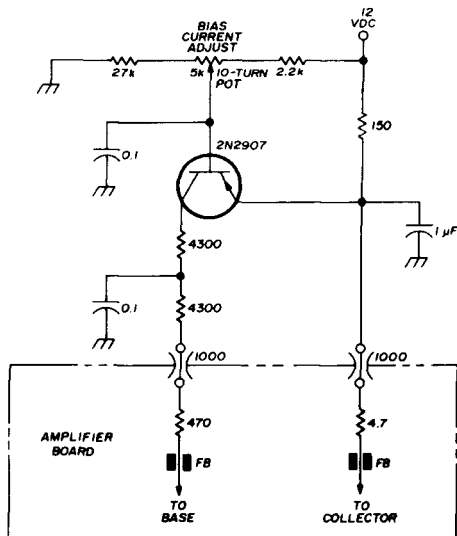


fig. 7. Active bias circuit provides adjustable collector currents from 2 to 12 mA. Components below dotted line are part of the amplifier circuit (see figs. 1 and 3).

*One of the more popular attractions at the annual West Coast VHF Conference is a receiver noise-figure competition, during which participants may optimize the performance of their receiving equipment with an automatic noise measuring system. A gas-discharge noise source is usually available for noise-figure measurements on both 1296 and 2304 MHz.

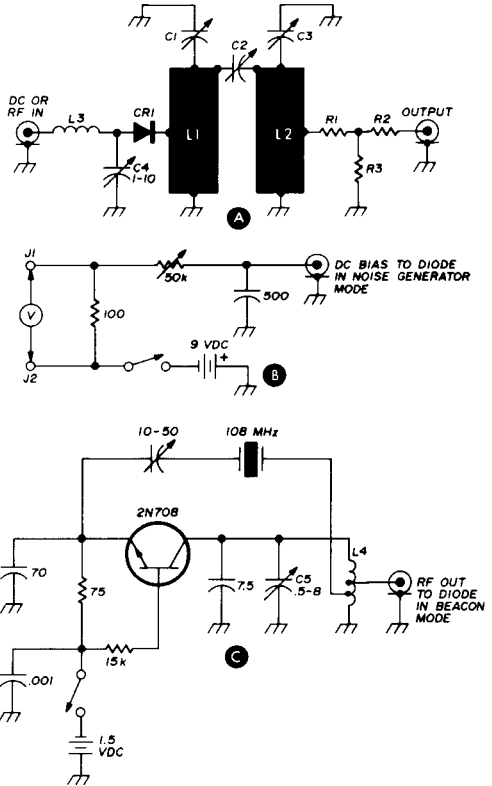
the ultimate in performance is achieved by alternate adjustments to first the input trimmer capacitor, then the bias pot. These adjustments are repeated until no further improvement in noise figure can be achieved.

The tuning elements of these amplifiers are inherently broadband. Without adequate selectivity at the input, bipolar transistors can exhibit disastrous overload characteristics. The cross-modulation and intermodulation effects of operating these preamplifiers in a high rf-density environment (i.e., virtually any populated area of the world today) can completely nullify any system noise-figure improvement. As a precaution it is good practice to provide input selectivity external to the preamplifier in the form of a high-Q filter in series with the antenna input. Single-pole coaxial or trough-line resonators, as well as some multi-pole microstripline filters, can provide adequate selectivity against out-of-band signals with a minimum of insertion loss (which would add to the amplifiers's noise figure in establishing receiver sensitivity).

When an input filter is used it is important to first optimize the preamplifier's noise figure as discussed above, then adjust the filter for minimum insertion loss at the operating frequency in a 50-ohm system. After mating the two, no further adjustments should be made to either the filter or the preamplifier unless a precision automatic noise-figure meter is available.

Multipurpose uhf tuning instrument

It is interesting to note the similarities between the diode noise sources and rf beacons commonly used for adjusting uhf receiving preamplifiers. Both instruments normally have an output circuit consisting of a microwave diode feeding a tuned circuit, with an output port matched to 50 ohms. The two pieces of test equipment differ in that the diode is fed with direct current



- C1, C2 0.5 to 2 pF piston trimmer capacitor
 - C3 (low-cost ceramic type acceptable)
 - CR1 microwave diode (1N25 or equivalent)
 - L1, L2 grounded microstripline, 0.3" (7.5mm) wide, 0.9" (23mm) long, tapped 0.2" (5mm) up from grounded end
 - L3 6 turns no. 20, 0.1" (2.5mm) diameter, 0.5" (13mm) long
 - L4 5 turns no. 20, 0.25" (6.5mm) diameter, 0.5" (13mm) long, tapped at 1 and 2½ turns from grounded end
 - R1, R2 10 ohm, ¼ watt carbon composition
 - R3 110 ohm, ¼ watt carbon composition
- fig. 8. Combination weak-signal source (C) and diode noise generator (B). Microstriplines L1 and L2 are etched on 1/16" (1.5mm) double-clad, fiberglass-epoxy circuit board (A).**

when used as a noise generator, versus rf when used as a weak signal source. The diode functions as a white-noise generator in the former application as opposed to harmonic generation in the latter.

There is no reason why the two

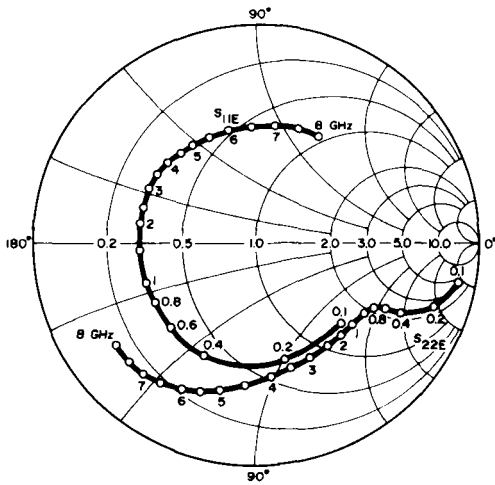


fig. 9. Input reflection coefficient, S_{11} , and output reflection coefficient, S_{22} , vs frequency for the Hewlett-Packard 35866E transistor ($V_{CE} = 10$ volts, $I_C = 10$ mA).

functions cannot be combined into a single instrument. Fig. 8 represents an attempt to do so. The circuitry associated with the diode and its two-pole resonator is similar to the scheme I used in the local-oscillator section of the 1296-MHz transceiver converter with filtering consisting of microstripline inductors etched on G-10 glass epoxy PC board. The three resistors at the output port form a 3-dB T-attenuator which assures a reasonable approximation of a 50-ohm output impedance. (In the weak-signal-source mode, additional pads may be placed between the output of the generator and the input to the receiver.)

The diode's input circuit is an adaptation of the familiar L-match used in the multiplier circuits of trough-line converters. The crystal-controlled 108-MHz oscillator which feeds the diode in the signal-generator mode is borrowed from a popular signal source design⁷ while an existing noise-generator circuit¹⁷ provided a workable diode biasing scheme. In short, the design presented in fig. 8 is an amalgamation

of various pieces of uhf test equipment into a single package.

Tuneup of the multipurpose instrument consists of merely connecting it in the rf beacon mode and tuning capacitors C1, C2, C3, C4 and C5 for maximum 1296-MHz signal into your receiver.

circuit design

For those readers who care to follow the calculations involved, the remainder of this article documents the procedure used to design the microstripline matching circuits of these 1296-MHz amplifiers. It should be pointed out that there are at least as many different methods for designing microstripline amplifiers as there are microwave engineers, and no one technique is necessarily any better or more workable than the others. The method shown here represents nothing more sacred than my own personal preference. It should be pointed out, however, that many of the more elegant amplifier designs used in the microwave industry are so complex as to be solvable only with the aid of a large digital computer. The designs shown here, though somewhat less precise, can be calculated by the average experimenter using only a slide rule.

Let us first consider a method for obtaining a complex conjugate impedance match to the amplifier transistors. Toward the end of this article I will discuss the special case of precisely mismatching the input to the first stage to obtain optimum noise figure.

Nearly all microwave semiconductor manufacturers publish Smith charts depicting the complex input and output impedances (S_{11} and S_{22} , respectively) as a function of frequency. Figs. 9, 10 and 11 show such data for the HP 35866E, HP 35826E and NEC VO21, respectively. In addition to these charts, most manufacturers furnish tabulated data listing the input and output impedances at various frequencies and differ-

ing bias conditions. It is important to note that these impedances vary significantly with changes in the dc operation of the transistor.

When tabular data is furnished, complex impedances are generally shown in polar form, i.e., magnitude and angle. This polar form may be converted to the more familiar rectangular notation ($A \pm jB$) on a Smith chart as indicated in fig. 12. Note that the magnitude is a decimal indication of the distance along a radius of the Smith chart, from zero (center) to 1 (circumference). The angle

output matching

It was desired to develop a single output circuit which would approximate a match to 50 ohms for any of the three referenced transistors, under either of the two indicated bias conditions. Assume for a moment that the reactive component of each transistor's parallel-equivalent complex output impedance is cancelled by a reactance of like magnitude and opposite sign shunting the collector (this will be accomplished shortly). Under these conditions

table 1. Typical S-parameters at 1.3 GHz (given in polar form, rectangular form and as parallel shunt equivalent, respectively).

	HP 35826E $I_c = 5 \text{ mA}$	HP 35826E $I_c = 10 \text{ mA}$	HP 35866E $I_c = 5 \text{ mA}$	NEC V021 $I_c = 10 \text{ mA}$
Input reflection coefficient, S_{11}	—	$0.61 \angle 178^\circ$ $12.5 + j0.5$ $12.77 \parallel +j325.6$	—	$0.626 \angle 171^\circ$ $11.75 + j4.0$ $13.11 \parallel -j38.5$
Output reflection coefficient, S_{22}	$0.57 \angle -41^\circ$ $75 - j80.0$ $160 \parallel -j150$	$-0.51 \angle -40^\circ$ $76 - j67.5$ $136 \parallel -j153$	$0.61 \angle -37^\circ$ $80 - j90.0$ $181 \parallel -j161$	$0.266 \angle -65^\circ$ $55 - j27.5$ $69 \parallel -j138$

listed represents the direction of that radius. (For further material on the use of Smith charts see reference 19.)

Table 1 lists complex impedances, in both polar and rectangular form, for the transistors used in these amplifiers under the applicable dc bias conditions, at 1296 MHz. It is useful to convert the complex series impedances to their shunt equivalent circuit values. The applicable formulas are

$$R_p = R_s + \frac{X_s^2}{R_s}$$

$$X_p = \frac{R_s \times R_p}{X_s}$$

where R_s and X_s represent the resistive and reactive components, respectively, of the complex series impedance, and R_p and X_p represent the components of the parallel equivalent circuit. The parallel equivalents are included in table 1.

the impedance to be matched to 50 ohms is a real value of magnitude R_p . For the conditions being considered, the value of R_p varies between 68.75 and 181.0 ohms. A compromise output circuit should match to the geometric mean of these outside values

$$R_p (\text{mean}) = \sqrt{R_p (\text{max}) \times R_p (\text{min})}$$

$$= 111.6 \text{ ohms}$$

A 111.6-ohm nonreactive source may be matched to a 50-ohm nonreactive load through a quarter-wave matching transformer of characteristic impedance

$$Z_o = \sqrt{Z_{in} \times Z_{out}} = \sqrt{111.6 \times 50}$$

$$= 74.7 \text{ ohms}$$

Still assuming no reactive component, the actual amplifier output impedance resulting from the use of each transistor

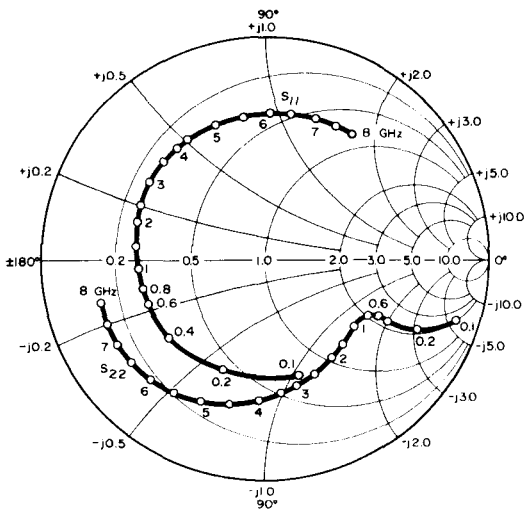


fig. 10. Typical input reflection coefficient, S_{11} , and output reflection coefficient, S_{22} , vs frequency for the H-P 35826E transistor ($V_{CE} = 15$ volts, $I_C = 15$ mA).

in this circuit would be that transistor's equivalent parallel output resistance transformed through a 75-ohm quarter-wave section. These values, along with

the resulting output vswr, are listed in table 2. Note that, even though restricted to a single output circuit for various combinations of device and collector current, this results in an acceptably low vswr well below 2:1.

Now, what about the reactive component of the transistor's output impedance so blithely ignored up to this point? Table 1 reveals the various values of parallel equivalent reactance, X_p , to be a shunt capacitive reactance ($-j$) in all cases, varying between 137.5 and 161.0 ohms. Obviously these capacitive reactances could be cancelled out by a variable inductor of like reactance range connected in shunt with the transistor's collector. However, from a practical standpoint, a variable capacitor is a more desirable tuning element than a variable inductor.

table 2. Actual output impedance and vswr of the Hewlett-Packard and NEC transistors in the circuits of figs. 1 and 3.

device	I_C	$Z_{out} = 75^2/R_p$	vswr
HP 35826E	5 mA	35.16 ohms	1.42:1
HP 35826E	10 mA	41.36 ohms	1.21:1
HP 35866E	5 mA	31.08 ohms	1.61:1
NEC V021	10 mA	81.82 ohms	1.64:1

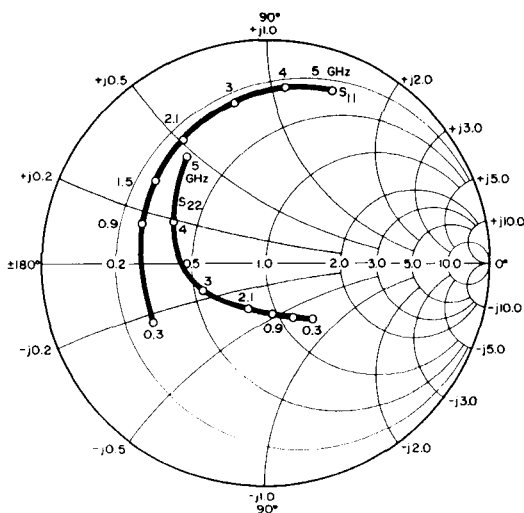


fig. 11. Typical input reflection coefficient, S_{11} , and output reflection coefficient, S_{22} , vs frequency for the NEC V021 transistor ($V_{CE} = 10$ volts, $I_C = 10$ mA).

The desired inductance may be realized by connecting a shunt capacitance to the collector through a quarter-wave transformer. As luck would have it, a quarter-wave transformer already exists at the collector circuit — the 75-ohm section used to match the transistor's parallel equivalent resistance to 50 ohms!

What value of capacitive reactance must be connected to the load end of the 75-ohm quarter-wave transformer? It must match the inductive reactance, X_{out} , resulting from transforming the transistor's shunt capacitive component, X_p , through a 75-ohm quarter-wave section. The relationship is

$$X_{out} = -\frac{Z_0^2}{X_{in}}$$

which results in a range of values for X_{out} of between $+j34.9$ and $+j40.9$ ohms. The required values for the trimmer capacitor, from $C = 1/2\pi f X_c$, is 3 to 3.5 pF at 1296 MHz. Allowing for variations between devices and imperfections in the micro-striplines, a 1 to 5 pF trimmer will assure cancellation of any of the transistors' shunt reactive

amounts to transforming 50 ohms to the required equivalent parallel resistance, R_p . The geometric mean of the R_p values for the two transistors is

$$R_p (\text{mean}) = \sqrt{12.77 \times 13.11} = 12.94 \text{ ohms}$$

The required quarter-wave matching section has a desired characteristic impedance of

$$Z_o = \sqrt{12.94 \times 50} = 25.4 \text{ ohms}$$

Fig. 3 reflects these values. The resulting amplifier input impedance, as shown in table 3, yields an input vswr of better than 1.1:1.

noise-matching the first stage input

A parameter frequently specified for low-noise microwave transistors is the source reflection coefficient for optimum noise figure, $\Gamma_s \text{ opt}$, or simply Γ_o , optimum source reflection coefficient. Either designation is an indication of the *source* impedance seen by the transistor input terminals when matched for optimum noise figure. The reference plane is at the edge of the transistor package, perpendicular to the input lead. The optimum source reflection coefficient, Γ_o , is often plotted as a function of frequency on a Smith chart, as indicated in fig. 13. It is important to realize that the impedances read from the Γ_o Smith chart are those looking back toward the source, *not* the transistor's input impedance.

If the input impedance to the amplifier (typically 50 ohms non-reactive) can be transformed to appear at the transistor input as Γ_o , optimum noise

table 3. Actual input impedance and input vswr of Hewlett-Packard and NEC transistors in the circuit of fig. 3.

device	$Z_{in} = 25^2/R_p$	vswr
HP 35826E	48.9 ohms	1.02:1
NEC V021	47.7 ohms	1.05:1



Ground-plane side of the preamplifier stage. Note that component placement in the dc bias circuitry is not critical but the feed-through capacitors make convenient stand-offs. The transistor mounting hole is covered with copper foil to promote good rf shielding.

components. The resulting compromise output circuit is used in the circuits of figs. 1 and 3.

input matching

This discussion applies only to the second stage as a complex conjugate match into the first stage is not desired. From table 1, the input shunt reactance values for the two different transistors are found to be $+j38.5$ and $+j325.6$ ohms. As these values are positive (inductive), a shunt variable capacitor directly at the base will provide reactive matching. The capacitance range required is 0.38 to 3.2 pF. To provide sufficient tuning range, a 0.35 to 3.5 pF capacitor was chosen.

With the shunt reactive component thus disposed of, input matching now

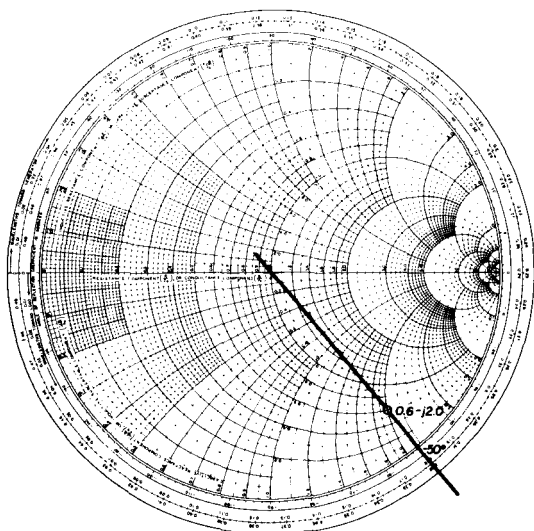


fig. 12. Using the Smith chart to convert impedance in the polar form to rectangular form ($A \pm jB$). The magnitude and angle $0.8 \angle -50^\circ$ lies on a radius passing through the -50° point on the outer circumference, 80% of the linear distance from the center to the edge of the chart, at $0.6 - j2.0$ on the normalized Smith chart shown here ($30 - j100$ ohms in a 50-ohm system).

figure will result. Fig. 13 shows that, for an HP 35866E option 100 device operated with $I_C = 5$ mA, Γ_o at 1.3 GHz is equal to $40 + j25$ ohms. This translates to a shunt equivalent circuit of 34.8 ohms in parallel with $+j55.6$ ohms. Disregarding the inductive reactance shunting the base for a moment, the required real component can be readily realized by transforming the 50-ohm input impedance through a quarter-wave section with a characteristic impedance

$$Z_o = \sqrt{50 \times 34.8} = 41.7 \text{ ohms}$$

As before, the desired inductive reactance shunting the transistor is achieved by adding a shunt capacitive reactance a quarter wavelength away from the transistor. The required capacitive reactance is

$$X_C = \frac{Z_o^2}{X\Gamma_o} = 31.3 \text{ ohms}$$

which at 1.3 GHz would represent a capacitance of 3.9 pF. A 1 to 5 pF trimmer is used for tuning the input to the first stage for optimum noise figure as shown in fig. 1.

modified interstage circuit

A number of the active 1296 operators who reviewed preliminary copies of this manuscript expressed an interest in a modified interstage design which would allow both stages of preamplification to be combined on a single amplifier board. Although I personally prefer separate modules, I concede that such a design would be of some value. A two-stage 1296-MHz preamplifier design by W6KQG²⁰ transformed the output impedance of one HP-21 into the complex conjugate of the second HP-21's input impedance through a single quarter-wave microstripline. A similar approach for these amplifiers, with the added provision of reactive tuning, is shown in fig. 14.

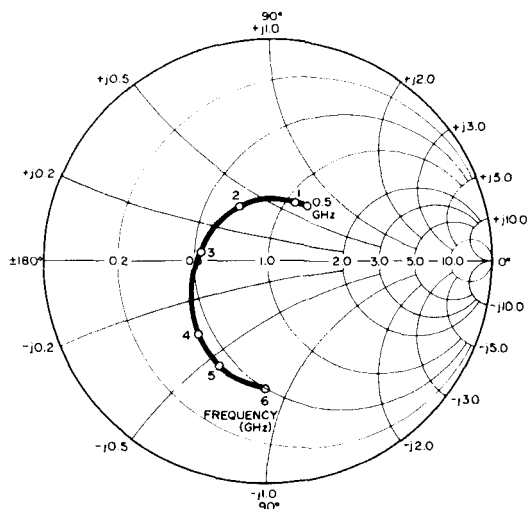


fig. 13. Typical optimum source reflection coefficient, Γ_o vs frequency for the Hewlett-Packard 35866E transistor ($V_{CE} = 10$ volts, $I_C = 5$ mA).

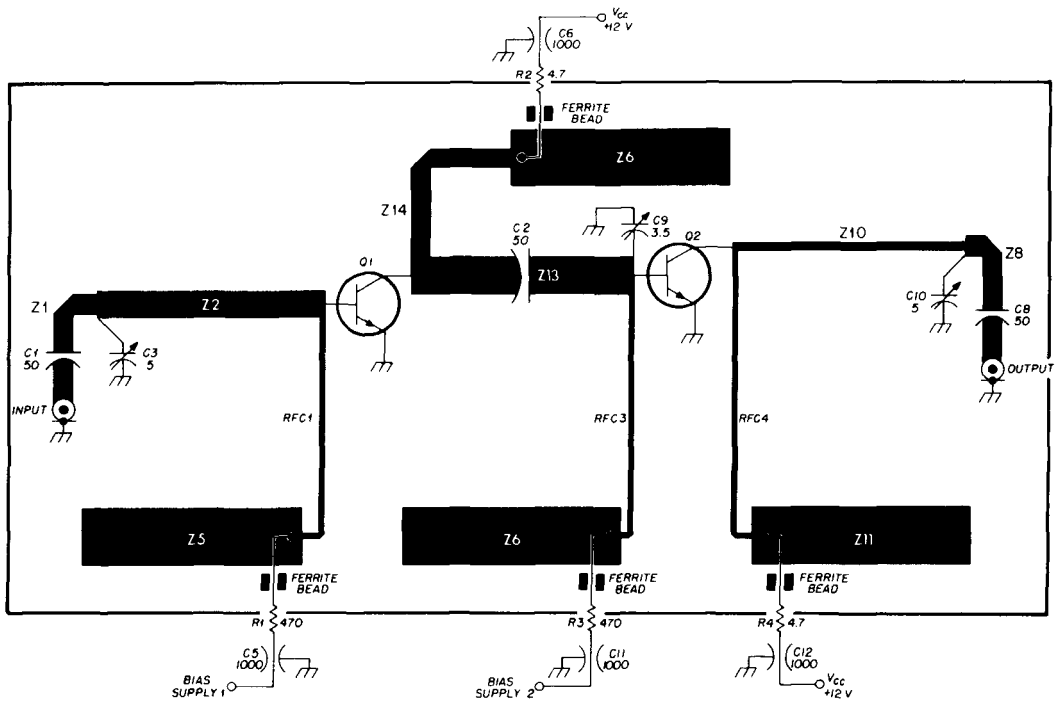


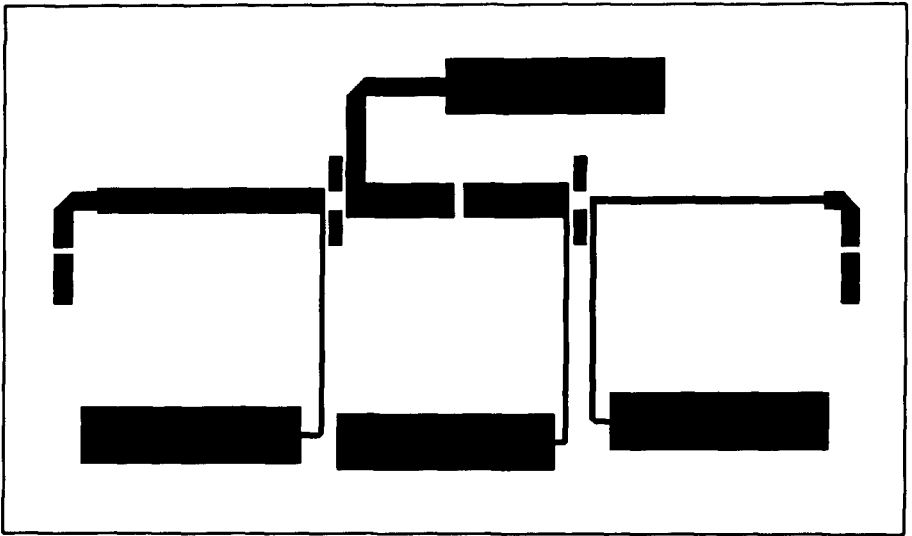
fig. 14. Modified interstage circuit for building the two 1296-MHz amplifier stages in figs. 1 and 3 on one circuit board. Circuit-board layout is shown in fig. 15.

It should be pointed out that, with two stages cascaded in this manner, it is impossible to measure the output vswr of the first stage or the input vswr of the second. Therefore, proper matching can be approximated only by tuning the amplifier for maximum gain. This is complicated by the fact that interstage matching is influenced by the output reflection coefficient, S_{22} , of transistor Q1 (which varies with I_{C1}), the input reflection coefficient, S_{11} , of Q2 (which varies with I_{C2}), and the complex impedance of the interstage network (which is controlled by capacitor C9). In practice C9 should be adjusted for maximum amplifier gain with I_{C1} set at 5 mA and I_{C2} at 10 mA. Further adjustments should then be made while monitoring total system noise figure with an automatic noise meter.

A full-size printed-circuit layout for the unified two-stage amplifier board is provided in fig. 15. Preliminary tests show the total gain of this amplifier to be within 0.5 dB of the two-module arrangement. With HP-22 transistors at Q1, both amplifier configurations yielded noise figures of 2.3 dB, measured on a Hewlett-Packard 340B Automatic Noise Meter with an AIL 7010 argon-discharge noise source.

acknowledgements

I wish to express my appreciation to George Bowden and Len Lea, both of Hewlett-Packard, as well as to Jerry Arden and Jerry Swan of California Eastern Labs, for furnishing technical assistance, device scattering parameters and sample transistors for building these amplifiers.



Z13 37 ohm, quarter wavelength microstripline, 0.18" (4.5mm) wide, 1.16" (29.5mm) long, gap in center for installing C2

Z14 150 ohm shunt inductive reactance consisting of 50 ohm, 0.2-wavelength microstripline, 0.10" (2.5mm) wide, 0.96" (24.5mm) long

fig. 15. Full-size printed-circuit layout for the two-stage 1296-MHz amplifier. For identification of microstriplines Z13 and Z14, see fig. 14.

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ham radio

digital touch-tone encoder

for vhf fm

Touch-tone encoder
design is simplified
through the use
of a new cmos
tone-encoder IC —
the Motorola MC14410

A new cmos integrated circuit introduced by Motorola, the MC14410, allows the design of a very compact, accurate, low power Touch-Tone[†] encoder system. The IC provides the full 2-of-8 encoding from a basic 1.0-MHz crystal oscillator. All that is required in addition to the MC14410 IC is a 2-of-7 or 2-of-8 key pad switch matrix such as a Chromerics ER-21623 or a ER-21611. Some of the outstanding features provided by the MC14410 are low power

(typically 7.5 milliwatts at 5 volts during standby), full 4x4 frequency matrix (16 characters), multiple key lockout, output frequency accuracy of $\pm 0.2\%$, built-in internal crystal oscillator and output available for transmitter PTT.

encoder operation

Operation is most easily followed by referring to **fig. 1**. The encoder consists of a common 1-MHz crystal oscillator, two separate frequency divider chains with decoders and a digital sine-wave generator for each of the dividers. The inputs from the key pad originate from a 4x4 switch matrix which generates a four-row (R1 to R4) and four-column (C1 to C4) input signal according to **table 2**. The row and column inputs to the MC14410 are true when they are at a logic low level. For example, when a number 4 is pressed, R2 and C1 go to a logical zero level or ground. These inputs program the low group and high group counters to provide frequencies

* Manager, Digital Communications Technology Development, Computer Applications Engineering, Motorola Semiconductor Products, Inc., Phoenix, Arizona.

[†]Touch-Tone is a trademark of American Telephone and Telegraph.

Jon DeLaune, W7FBB*

of 770 Hz and 1209 Hz, respectively. The row inputs produce the low group frequencies of 697, 770, 852 and 941 Hz, while the column inputs produce the high group frequencies of 1209, 1336, 1477 and 1633 Hz. The row and column inputs must each provide a valid

logic exclusive NOR of the low and high group frequency dividers. The logic function is of no consequence for this particular application, but the pulsed output which is produced is important for providing a transmitter push-to-talk (PTT) signal.

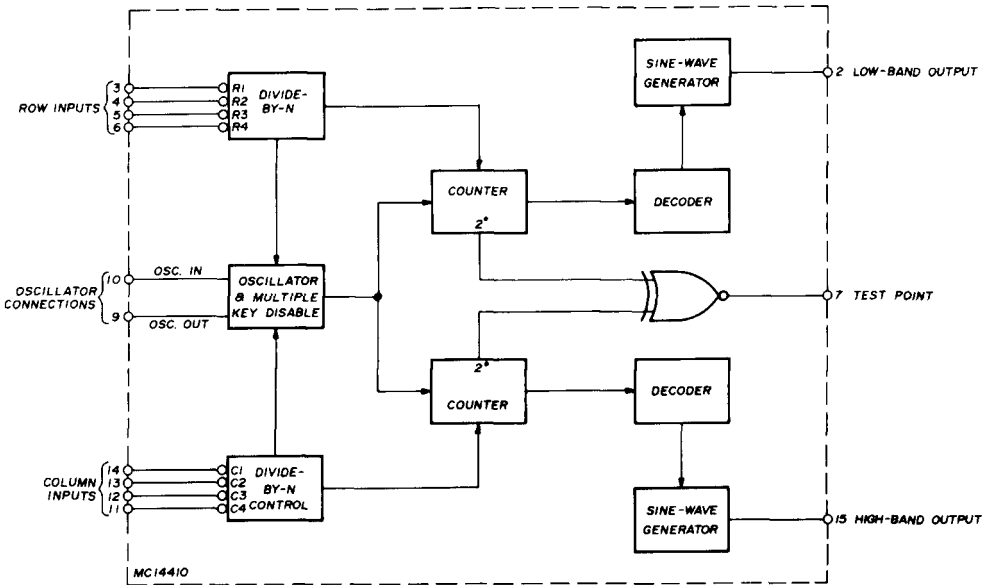


fig. 1. Block diagram of the CMOS tone encoder IC, the Motorola MC14410.

1-of-4 input condition, as shown in table 1, before a two-tone output will occur.

The low and high group frequency dividers provide inputs to two eight-level digital-to-analog converters which synthesize the two sine-wave group tones. The outputs of these generators (low group at pin 2 and high group at pin 15), when summed together, provide the resultant two-tone output signal. Summing can be accomplished by a resistor summer with capacitive filtering required to smooth out the digital steps in the digital sine wave. This rolloff is effected by following the resistor summer with a 0.01 μ F capacitor to ground.

Another output of interest is the one labeled as a test point. This output is a

The 1.0-MHz oscillator is provided internally in the MC14410; the only external parts being the crystal and a 15-megohm bias resistor. The crystal should be a parallel resonant type, maximum $R_{series} = 540$ ohms, C_o maximum = 7 pF, test level = 1 mW, and fre-

table 1. Input to output conditions. Row and column inputs must each provide a valid 1-of-4 input before a two-tone output will occur.

row lines depressed	column lines depressed	tone group outputs
none	none	off
none	one	off
one	none	off
one	one	both on
two or more	one	high only
one	two or more	low only
two or more	two or more	off

quency = 1.0 MHz \pm 0.1% with 13-pF load capacitance.

tone encoder

Now that we have a general understanding of what the MC14410 tone encoder can provide, let's look at the

will drive loads down to 600 ohms with a level of 0.4 volt rms. This should be more than adequate for all 600-ohm radio systems such as produced by Motorola and others. The MPS-A17 amplifier provides a voltage gain of about five. If a level of only 0.1 volt rms

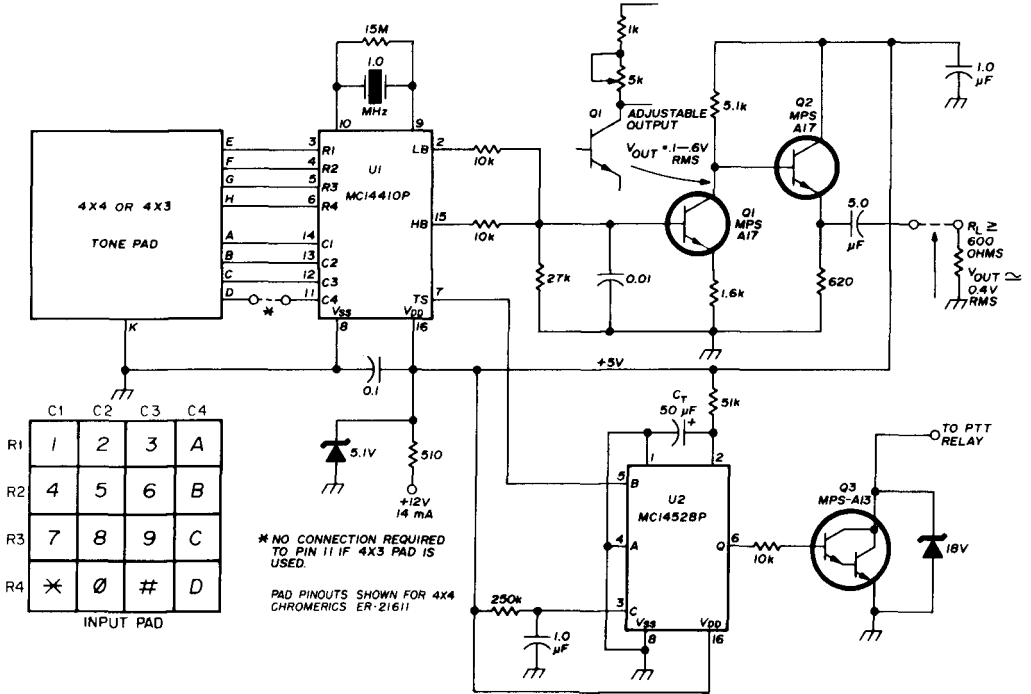


fig. 2. Digital tone encoder using the MC14410. With $C_T = 50 \mu\text{F}$, PTT hold time is 1 second (see text). Key pad matrix is Chromerics ER-21623 or ER-21611*.

amateur radio application shown in fig. 2. Included in the circuit are the Chromerics pad, the MC14410 IC encoder, U1, the resistor summer and rolloff filter capacitor, and a tone-amplifier/emitter-follower line driver, Q1/Q2. Shown below the amplifier is the PTT one-shot timer using a MC14528, U2. Although a 12-volt supply voltage is shown, any voltage of 5 volts or more may be used as long as a zener diode is used to supply 5 volts to the ICs.

The two-tone output level from Q2

is required for proper deviation, then the amplifier may be omitted and the 10k summing resistors may be followed by the MPS-A17 emitter-follower output stage.

The tone output to the transmitter should be ac coupled through a 5- μF capacitor. The output may also be coupled through an attenuating resistor

*For address of your local Chromerics sales representative, write to Chromerics, Inc., 77 Grand Dragon Court, Woburn, Massachusetts 01801.

to provide both the required deviation level and some impedance isolation to the normal microphone input, i.e., minimum loading of normal voice audio. The 5100-ohm MPS-A17 collector resistor could be a 5000-ohm potentiometer in series with a 1000-ohm fixed resistor. This would provide an adjustable output level between 0.1 and 0.6 volt rms.

If the transmitter you plan to use with the encoder has a high input impedance audio circuit, a large isolation resistor is required between the output of the tone encoder and the audio input to the transmitter. This resistor may require values between 50k and 300k for proper transmitter deviation.

The two digital group sine waves and the final summed output are shown in fig. 3. The upper waveform has a frequency of 697 Hz and a peak-to-peak amplitude of 0.7 volt. The center waveform has a frequency of 1336 Hz with an amplitude of 0.7 volt peak-to-peak. The bottom waveform is the two-tone

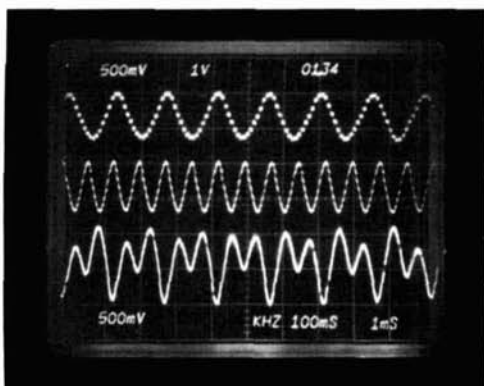


fig. 3. Output waveforms of the digital tone encoder. Upper waveform is low-group output at 697 Hz, center waveform is high-group output at 1336 Hz, and bottom waveform is the two-tone output.

Tone dial and key the transmitter. The MC14528 retriggerable one-shot IC has a 1.0-second hold time after tone release, i.e., the operator has at least one second of interdigit time available with-

table 2. Tone encoder truth table.

character input	row				column				high-group tone (Hz)	low-group tone (Hz)
	R1	R2	R3	R4	C1	C2	C3	C4		
1	0	1	1	1	0	1	1	1	1209	697
2	0	1	1	1	1	0	1	1	1336	697
3	0	1	1	1	1	1	0	1	1477	697
A	0	1	1	1	1	1	1	0	1633	697
4	1	0	1	1	0	1	1	1	1209	770
5	1	0	1	1	1	0	1	1	1336	770
6	1	0	1	1	1	1	0	1	1477	770
B	1	0	1	1	1	1	1	0	1633	770
7	1	1	0	1	0	1	1	1	1209	852
8	1	1	0	1	1	0	1	1	1336	852
9	1	1	0	1	1	1	0	1	1477	852
C	1	1	0	1	1	1	1	0	1633	852
*	1	1	1	0	0	1	1	1	1209	941
Ø	1	1	1	0	1	0	1	1	1336	941
#	1	1	1	0	1	1	0	1	1477	941
D	1	1	1	0	1	1	1	0	1633	941

output that is applied to the audio input of the transmitter; this signal has an amplitude of about 2.2 volts peak-to-peak.

The PTT switching circuitry allows the operator to simultaneously Touch-

out transmitter dropout. This time may be adjusted to suit the operator by changing the value of the 50- μ F capacitor, C_T , or the 51k resistor, R_T , or both.

ham radio

direct-reading capacitance meter

Design and construction
of a simple
direct-reading
capacitance meter
that measures
1 pF to 1 uF
in five ranges

Inflation and scarcity of parts have driven many hard-core homebrewers like myself to the surplus markets and swapfests where you can still find quality components at bargain prices. I have picked up a number of variable capacitors from these sources at very reasonable prices. Most of the time, however, the only way I had of determining the capacitance range was an eyeball estimate, and sometimes that proves very unreliable. I started to

appreciate the idea of owning a direct-reading capacitance meter.

A brief search through back issues of the amateur magazines netted a couple of circuits, but they didn't appeal to me. It seemed there ought to be a simpler, more direct way to do it, and I began to see what I could cook up.

theory

Since the readout is a dc meter, it is desirable that there be a linear relationship between the capacitance-to-be-measured and the dc output of the measuring circuit. It so happens that the average dc value of a pulse waveform has a direct linear relationship to the duty cycle of the waveform. This is illustrated in **fig. 1**.

Duty cycle is defined here as the ratio of pulse width (in units of time) to the time between the beginning of each pulse. In **fig. 1**, T1 is the pulse width, and T2 is the period (reciprocal of frequency) of the waveform. The pulses in **fig. 1** have a peak amplitude of 10 volts; therefore, when the duty cycle is 0.2, the dc value of the waveform is 0.2 times 10 volts, or 2 volts. In like manner, when the duty cycle is 0.5, the dc value of the waveform is 5 volts, and when the duty cycle is 0.8, the dc value is 8 volts. A general expression for the

Courtney Hall, WA5SNZ, La Verdura Drive, Dallas, Texas 75240

dc value of the waveform may be written as

$$\text{dc value} = \frac{T_1}{T_2}(E_p) \quad (1)$$

where E_p is the peak amplitude of the pulse in volts. If T_2 and E_p can be held constant, then the dc value will be directly proportional to T_1 , the pulse width. T_2 can be held constant by making the pulse train have a constant frequency, and E_p can be made constant by regulating the power supply voltage to the pulse-forming circuit.

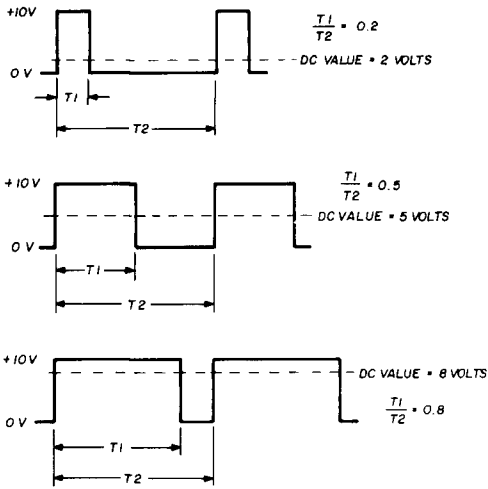


fig. 1. Average dc value of pulse waveform varies linearly with duty cycle.

This constant-voltage requirement makes an ac-operated power supply preferable to battery operation.

Now, if the width of the pulse (T_1) can be made directly proportional to the value of the capacitor-under-test, the problem is solved. Fortunately, a monostable multivibrator (one-shot) can be designed so that its pulse width is directly proportional to its timing capacitor, and this completes the idea for the system.

Fig. 2 shows a block diagram of how such a system may be built. The trigger

source is simply a free-running pulse generator which has a constant frequency and produces a narrow negative output pulse. Each time a trigger pulse occurs, the one-shot multivibrator initiates an output pulse whose width is determined by the capacitor-under-test. The larger the capacitor, the wider the pulse. Since a dc meter reads the average value of the pulse waveform, the meter may be calibrated directly to read capacitance. Care must be taken, however, that the pulse width does not exceed the time between trigger pulses. Also, the frequency of the trigger source must be high enough to prevent jitter of the meter needle.

circuit

A complete schematic of the capacitance meter is shown in fig. 3. The trigger source consists of a programmable unijunction transistor, Q1, an inverter-amplifier, Q2, and their associated components. There is nothing particularly critical about this part of the circuit, but the output trigger pulse at the collector of Q2 should have a frequency pretty close to 500 Hz, and the pulse amplitude should be about 12 volts (the power supply voltage). The pulse width of the trigger is about one microsecond in this circuit. Any trigger circuit which will provide this output may be used.

A Signetics NE555V timer IC is used as the one-shot multivibrator. According to the data sheet, the output pulse width of this device is given by

$$PW = 1.1RC \quad (2)$$

where PW is pulse width (the same as T_1 in eq. 1), R is the timing resistor (selected by the range switch) and C is the capacitance-under-test. Substituting into eq. 1,

$$\text{dc value} = \frac{1.1 RCE_p}{T_2} \quad (3)$$

With the range switch in any one position, all of the terms on the right-hand

side of eq. 3 are constant except C, the unknown capacitor; therefore, the dc value of the output voltage has a direct linear relationship to the capacitance being measured.

The resistors associated with the range switch (1k through 10 meg) should have a tolerance of 5 percent or

is no capacitor connected to the test terminals. Without the zero adjustment pot to buck out the voltage from this pulse, the meter will read 25 percent of full scale when the range switch is in the 100 pF position with no capacitor connected to the test terminals.

The zero adjustment circuit must have a relatively low resistance so that variations in its setting will have no appreciable effect on calibration. Since the zero setting circuit consists of a 470-ohm resistor and a 100-ohm pot in series across the 12-volt supply, it draws about 20 mA. This is another reason battery operation was ruled out.

The simple zener-regulated power supply provides 12 volts at up to 50 mA when plugged into a 117-Vac line (see fig. 4). No power switch or pilot light was included because the instrument is plugged into the ac line only when it is in use.

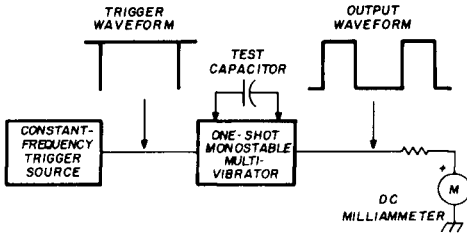


fig. 2. Block diagram of the direct-reading capacitance meter.

less. A 10k trimpot in series with the meter is used to make a one-time calibration; once adjusted, it needs no further attention.

It was necessary to include a front panel zero-adjustment pot for the lower capacitance ranges. This is because the input capacitance of pins 6 and 7 of the NE555V and stray capacitance of circuit wiring is about 25 pF. This produces an output pulse even when there

construction

A 5x7x3-inch (12.7x17.8x7.6cm) aluminum box houses the capacitance meter very nicely. Most of the circuitry is mounted on a 3½x3-inch (8.9x7.6cm) piece of perf board which is mounted directly to the meter terminals. The only important point in layout is to try to minimize the stray capacitance asso-

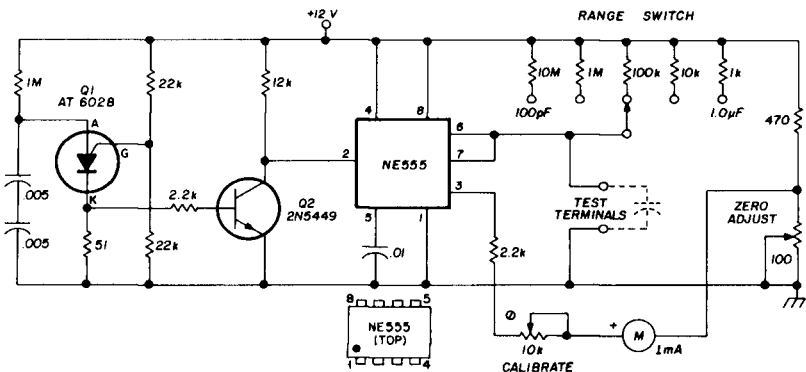


fig. 3. Simple direct-reading capacitance meter covers zero to 1 μ F in five ranges, and uses easily available components.

ciated with the wiring to the range switch and test terminals. I used a Radio Shack two-terminal pushbutton strip (274-315) for the test terminals. The circuit ground should be connected to the aluminum box.

The meter is a Radio Shack 0-1 milliammeter (22-052) which has a scale

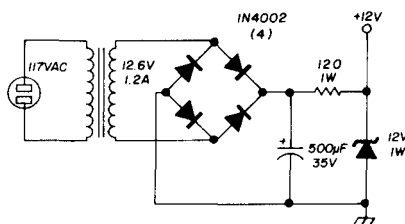


fig. 4. Simple power supply for the direct-reading capacitance meter.

length of about two inches. When the range switch is in the 100 pF position, the meter reads zero to 100 pF with minor scale divisions corresponding to 2 pF. It is not difficult to resolve a change in capacitance of 1 pF on this range.

Nearly all of the parts are available from Radio Shack. A notable exception is the 100-ohm zero adjustment pot, but any 100-ohm pot should work okay, and several are listed in the Allied Electronics catalog.

calibration and operation

Plug the ac power cord into the power line and set the range switch to 100 pF. With no capacitor connected to the test terminals, adjust the zero knob for a meter reading of zero. Now connect a 100 pF, 5 percent mica capacitor to the test terminals, and adjust the 10k calibration trimpot for a full-scale meter reading of 1.0 mA. This completes the calibration.

Each time the range switch is set to a new position, it will be necessary to readjust the zero knob *before* the

capacitor-under-test is connected to the test terminals. This is a bit inconvenient, but I judged it a reasonable tradeoff for the simplicity of the circuit. Most ohmmeters require this type of operation.

If desired, a two-pole range switch may be used, and the extra pole used to select one of five preset 100-ohm trim pots instead of the one 100-ohm pot I used. The five trim pots would then be adjusted for a zero meter reading on each range with no capacitor connected to the test terminals. This would simplify operation by eliminating the front panel zero control but would increase complexity and cost.

Since ac is not applied to the capacitor-under-test, polarized capacitors may be tested, but care should be taken that the negative lead of the capacitor is connected to the grounded test terminal.

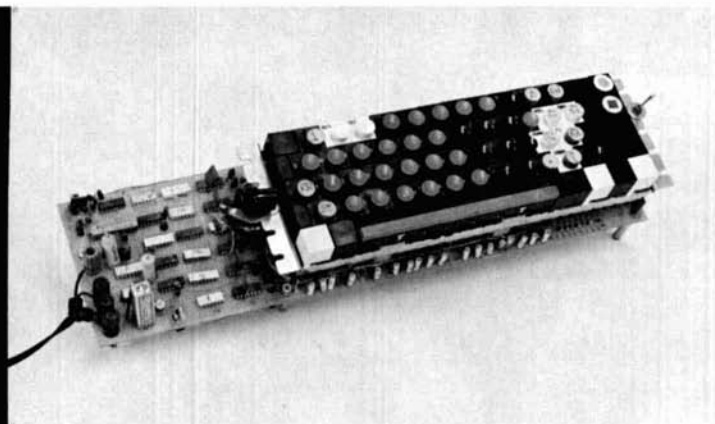
table 1. Capacitance range of the direct-reading capacitance meter.

range switch resistor	capacitance range
10 meg	0 to 100 pF
1 meg	0 to 1000 pF
100,000	0 to 0.01 μ F
10,000	0 to 0.1 μ F
1,000	0 to 1.0 μ F

Any attempts to modify the circuit so that it will measure capacitors larger than 1 μ F should be done by lowering the trigger frequency. Resistors in the range switch circuit should not have a value of less than 1000 ohms. Table 1 shows the capacitance range that may be measured with each range switch resistor.

Since the dc voltage across the capacitor-under-test reaches a value approximately two-thirds that of the power supply voltage, any capacitor to be measured should have a voltage rating of at least 8 volts.

ham radio



keyboard morse code generator

How to build a
keyboard
morse generator
for effortless,
high-speed CW

Talk about one-way people! Back in 1971, Dick Vogler, K7KFA, and I developed an Automatic Fist Follower (AFF)¹ that was capable of translating incoming Morse code and recording it with a strip printer. The machine immediately increased my capability for copying Morse code by several factors, but left my fist unaffected at about 30 wpm. This incongruity proved to be embarrassing, as a fast-fisted friend in Honolulu may well recall.

I had completed development of the AFF and had tested its receipt of taped code at up to 300 wpm. All I had to do to copy code was tune my receiver to

provide an 800 Hz beat note or so from a signal, lock onto the sender's fist by a quick setting of the adjustable rate control, and then set the machine to automatic copy. From then on the machine would automatically track the incoming transmission, creating a printed tape similar to that of a stock market ticker.

I was confident the machine could copy anyone on the air and must admit I entered the first live demonstration with the suppressed glee of a drag-strip driver who is carefully concealing a Corvette engine inside a stock Volkswagen chassis.

I methodically monitored the airways, awaiting suitable game. Eventually I found two operators chewing the rag in the 55 wpm range and decided to toss out the bait. The conversation I broke into was between a KH6 in Honolulu and a W6 in California.

Once I got the go-ahead I settled into an easy 20-wpm transmission, complimenting them on their fists. I then casually asked them how fast they could send.

"How fast do you want to receive?" inquired the KH6.

With appropriate confidence I again asked, "How fast can you send?"

Obviously intending a quick brush-off,

Clarence Gonzales, W7CUU, Phoenix, Arizona 85029

the KH6 switched to his keyboard and sent me a burst at about 75 wpm.

The AFF read him letter perfect, encouraging me to further bait him with, "Quite good, but please QRQ!"

I knew his second burst at about 85 wpm would be approaching his keyboard limit, so figured another QRQ would be the *coupe de grace*. The fellow was well prepared, however, shifting into third gear and sending me the next transmission in excess of 100 wpm. This caught me quite by surprise and though

At this point I disclosed my 'Vette engine and he admitted to the use of an automatic McElroy keyer. We then did a little more testing during which I successfully received him at 150 wpm, his limit since he had a dirty keyer head. Later that month, after he had cleaned the head, I copied him at 200 wpm and knew I had a going machine.

Most people knew I had to be pulling a fast one, but only inbound, mind you. There were suspicions that I was taping the incoming CW, but even though my

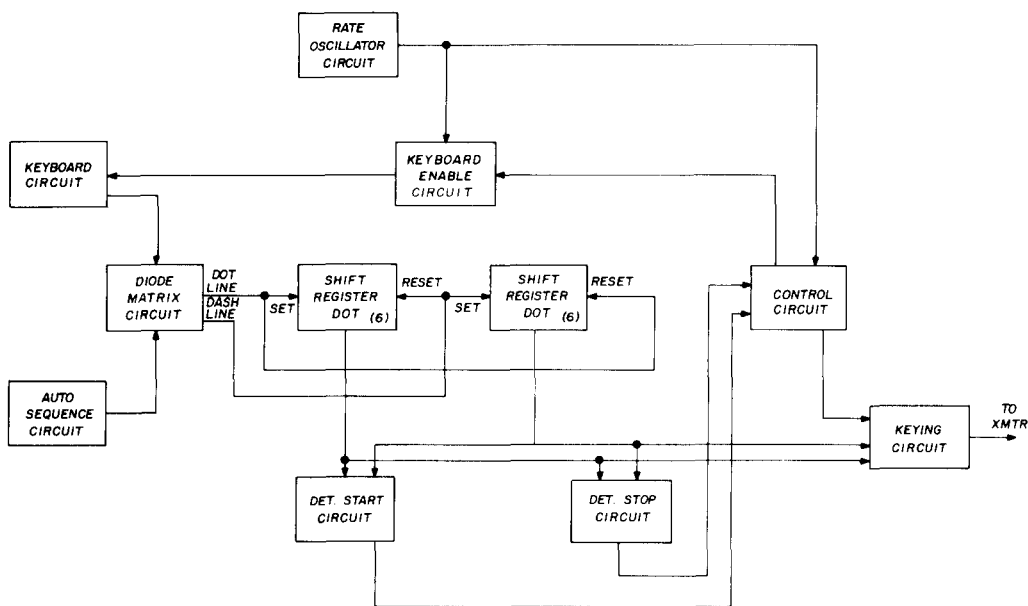


fig. 1. Block diagram of the keyboard Morse code generator. Detailed block diagram of element storage and space generation is shown in fig. 3.

my machine was experiencing no difficulty, I began to have some doubts. Neither of us knew what was up the other's sleeve, but he must have known something was phony in Phoenix when I was receiving 100 wpm easily while struggling to transmit 30 wpm.

"Hey," I admonished, "You're not sending by keyboard or keyer, KH6."

Came the equally perplexed reply, "Well you're not receiving by ear, either, buddy!"

responses were relatively slow, they weren't delayed. "Very suspicious," they said.

At this point I began an avid search for a CW keyboard. I was determined to increase my sending speed, if only to keep my AFF from developing egomania.

CW keyboards

Let's face it though, current commercial CW keyboards are not a good answer in view of most amateurs' already over-

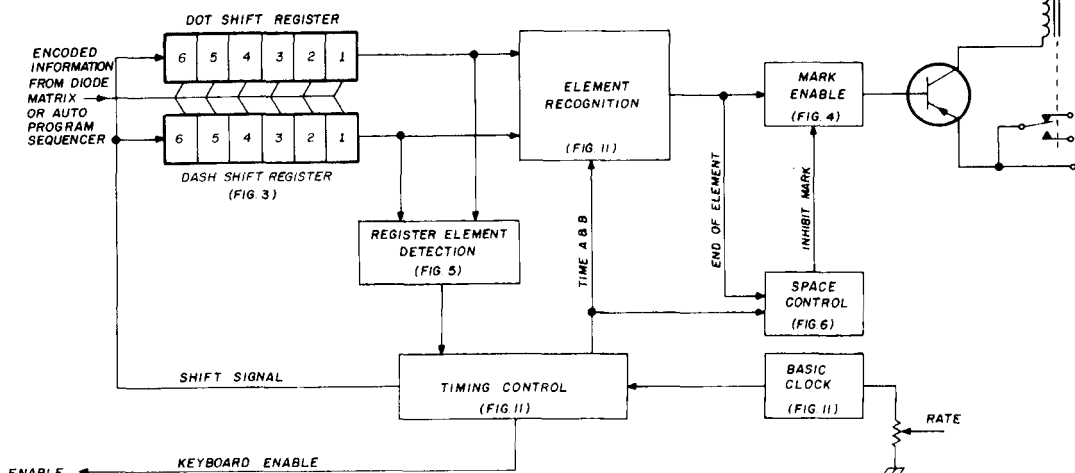


fig. 3. Block diagram of the encoding, element storage and code-generation portions of the keyboard CW generator.

extended hobby budgets. I had to find a solution which would satisfy my inherent stinginess as well as my compelling urge to "kluge," so I dug into my past issues of the various ham radio magazines and consulted my spare parts catalogs. I was convinced that I could build a better (and cheaper) CW keyboard.

I discovered during my research that code typewriters have been in existence for more than a decade.* My first reference was an article by Paul Horowitz, W2QYW describing a keyboard design,¹ in which he noted earlier attempts by others,^{2,3} but he was probably the first to recognize the advantage of adapting surplus circuitry from the computer market for this purpose. This market has expanded greatly since then and today offers not only rugged, low-power integrated circuits, but also low-cost computer-related keyboards ideally suited for CW keyboard sending purposes.

*The idea has undoubtedly been around a very long time, and attempts were made to put it into practice even before integrated circuits made it practical. Such a device, designed by an engineer/ham and using only relays, was demonstrated in a relay manufacturer's booth at the National Electronics Conference in Chicago in the early 1950s. editor

Another amateur-designed automatic Morse code generator, affectionately dubbed the "Button Box" by its designer, W5VFZ, is no doubt familiar to many hams.⁵ W5VFZ, while apologizing in his article for his lack of crafting expertise, nevertheless constructed an ingenious device from plywood, micro-switches and discrete circuitry. He also relied extensively upon computer surplus, adapting computer plug-in printed circuit boards containing NAND logic. His article included an "Introduction to Logic" that was quite informative for those not having the benefit of education in this area. Most of the basic concepts he covered are still applicable today, even for the newer ICs. Another logic discussion, broader in nature and more up-to-date, appeared in a recent issue of *ham radio*.⁶

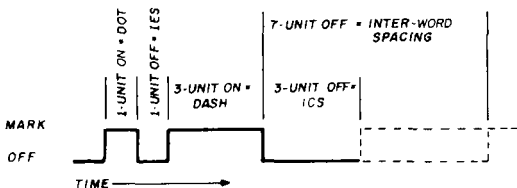


fig. 2. Morse Code element and spacing relationships. IES indicates inter-element space; ICS indicates inter-character space.

mark-state of three unit lengths. The *off* condition is used to specify separation of code elements, as well as separation of characters and words within a message. Specifically, as shown in fig. 2, a single unit space is used to separate elements within a character or *pro-sign*. Character separation is established by an off-state of

The following description explains the timing control necessary for varying unit timing and describes the overall encoding and generation operations involved in transforming the key selections into automatic code transmissions. This description is supported by the basic block diagram, fig. 1, and other supporting

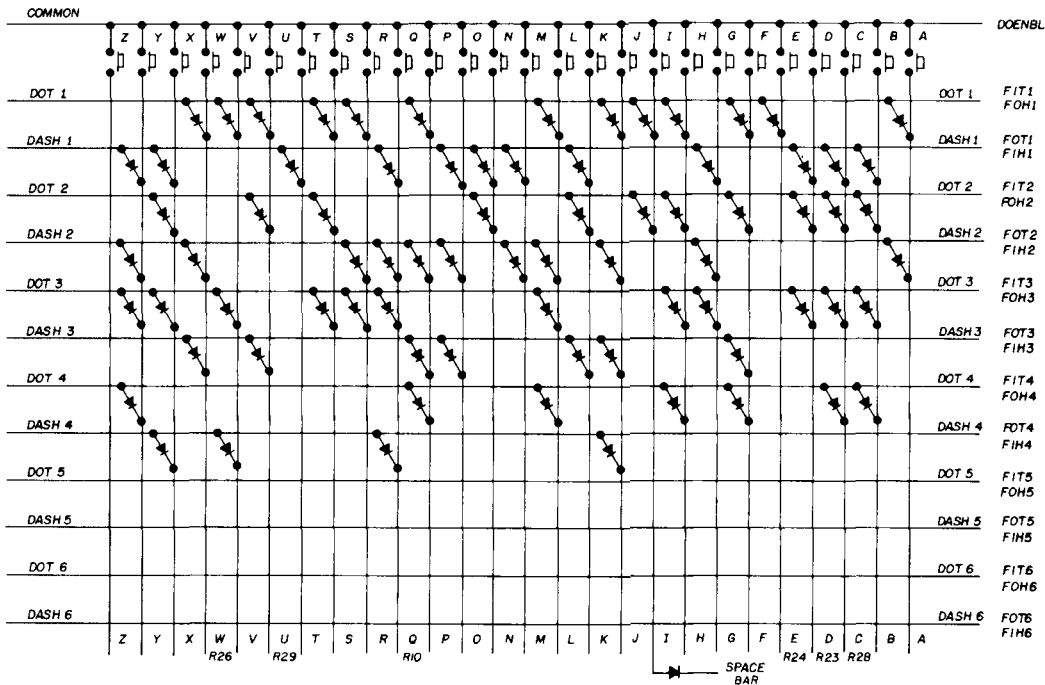


fig. 5. Diode matrix for the 26 letters of the alphabet. Diodes are 1N914 or equivalent. Connections at right side go to shift registers shown in fig. 7.

three units, and a seven-unit off-state indicates a separation between words.

A single time unit does not specify any finite time period; instead it provides a time base upon which code elements and spaces can be distinguished. Single time units vary according to sending speed, but the relationships of elements and spaces to the unit length remain constant. Since I wanted to build a code generator with a flexible transmission rate, it was necessary to build in the capability of varying the basic unit timing while retaining the code and space relationships.

diagrams referenced through the remaining text.

As indicated in fig. 1, the code generator's operation can be considered in two basic parts. One part is the *character encoding*, which responds to a key selection by encoding it into the selected character's dot and dash elements. This is accomplished through a diode matrix that transfers the encoded element status to a storage area. Once stored, the elements are acted upon by the second part, the *code-generation* circuitry.

The code-generation circuitry individ-

ually examines the code elements in storage and then enables the keying control circuit until the mark-state has represented the detected element. Following each element transmission the circuit automatically disables the keying control for one time period to separate the completed element from the next. After

number of code element combinations. Each position has a pair of flip-flops associated with it. During encoding the lowest-order flip-flop pair is loaded with the first element to be transmitted.

If the encoded element for any specific code position is a dot, the dot flip-flop at that position of the register is

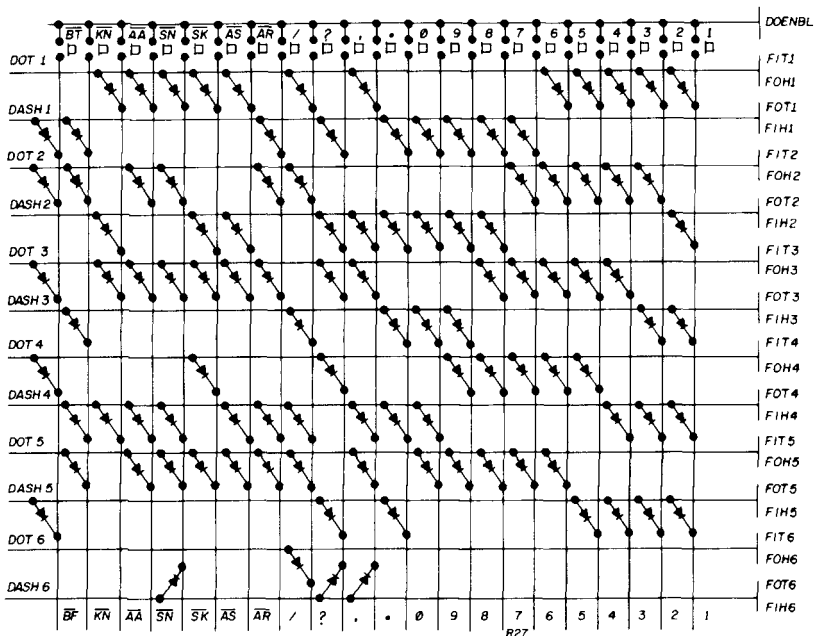


fig. 6. Diode matrix for numerals and special characters. Diodes are 1N914 or equivalent.

all elements for the selected key have been examined and generated, the circuit again disables the keying control for three time units before allowing the next key selection to be processed. This provides the necessary inter-character spacing.

character encoding

As shown in fig. 1, two shift registers perform the holding function for those elements encoded by the diode matrix when a key has been pressed. This pair of registers — one for dot elements and one for dash elements — can be considered as a functional unit representing six element positions. Six positions are necessary to fully represent the largest possible

set during the key depression. The dash flip-flop at the same position sets (and the dot flip-flop resets) if the element is a dash. Codes of less than six elements leave unused higher-order positions of the registers unfilled following the encoding. An unfilled position is indicated by both dot and dash flip-flops reset.

Note, for example, the shaded positions in the shift registers shown in fig. 4. These positions represent the flip-flops that would be set following the encoding of the letter A. Beginning at the lower-order position (one), the register could then be interpreted as containing the code for an A; a dot followed by a dash. The necessary inter-element spacing is not held

by the registers, but is supplied by the circuit during the actual code transmission.

Fig. 4 provides a partial illustration of the diode matrix and dot/dash shift registers used in the encoding operation.

causes the F1H1 and F1T2 flip-flops to be dc cleared. Since position one of the registers is the first examined during code transmission, this correctly represents the code for an A – with the exception of

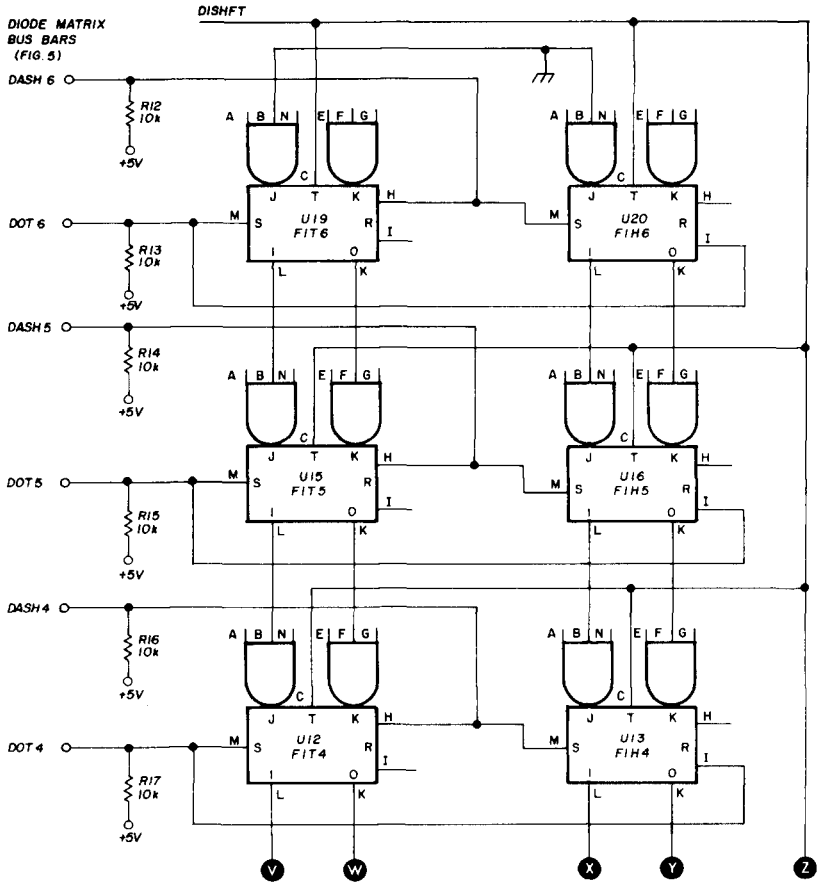


fig. 7. Dot-dash shift register receives inputs from diode matrices shown in figs. 5 and 6. All ICs are Sylvania SF200.

Only three of the six (total) flip-flop pairs and two of the possible key encodings are shown. Note that selection of an A permits the clock-driven enable (*downbl*) signal to assume a low level during each of the generator's clock pulses. This action causes the dc setting of flip-flops F1T1 (*dot register, position one*) and F1H2 (*dash register, position two*); it also

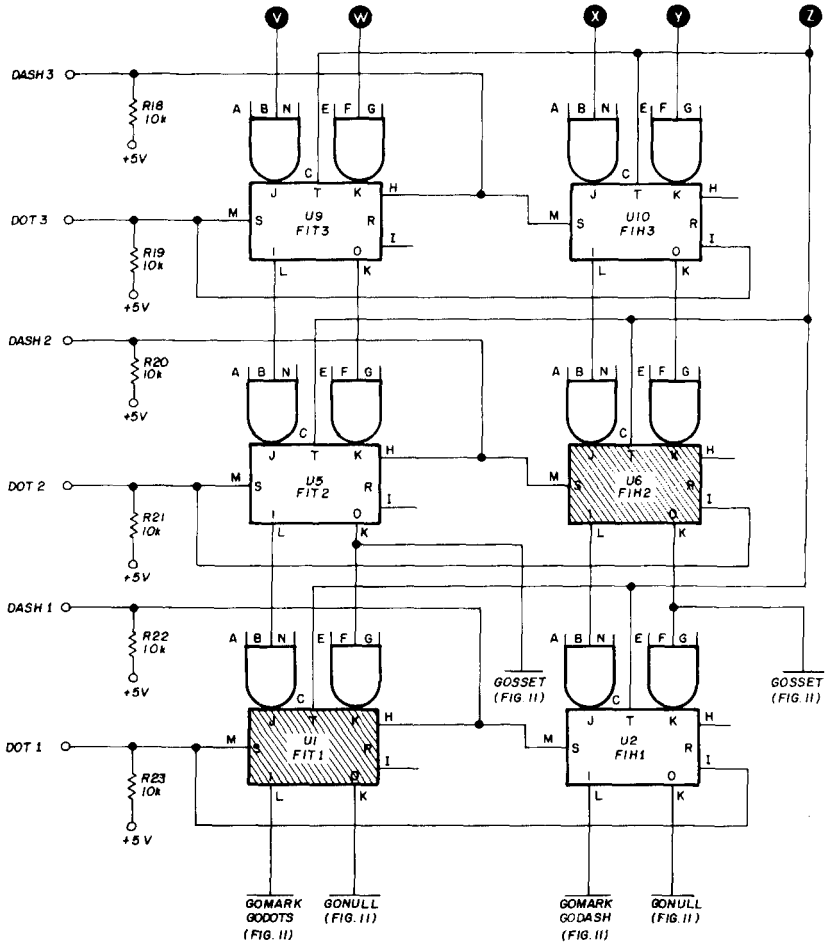
the inter-element spacing, which is provided by the code generation circuit.

code generation

Following code storage the elements are sequentially shifted through the registers from the higher-order toward the lower-order positions. The lowest-order position (*one*) is examined after each

shift and the mark circuitry within the generator is enabled for the time period necessary to reflect the detected dot or dash element. Figs. 7 and 11 depict part of the logic necessary to shift elements from

F1TIMA to the *ac set* input to F1TIMB, which forms a two-stage binary counter. This permits the flip-flops to count elapsed clock periods fed into F1TIMA and thus measures time durations when



the register and to determine the correct mark period for the elements as they are examined at position one.

The timing control of the mark period is established by flip-flops F1TIMA (time A) and F1TIMB (time B), shown in fig. 11, which are enabled for counting during each code transmission sequence. Counting is achieved by the cross-coupled connection from the 1 output of

enabling the mark circuit. The mark circuit is enabled for one clock period to represent a dot element and for three clock periods to represent a dash element. During a code character generation the mark circuit is disabled for one clock period following the mark period and the counter is readied (F1TIMA and F1TIMB cleared) in preparation for the next element.

The count versus TIMA/TIMB (time A/time B) table in fig. 11 indicates the state of flip-flops F1TIMA and F1TIMB when permitted to count for up to three clock periods. These states are monitored by gate GOMARK (*mark*) and result in the enabling of either AND gate input when counting is initialized. If counting begins with position one of the shift

takes place on the clock pulse when F1TIMA is set and F1TIMB is clear, which would be the first clock pulse. Shifting for a dash takes place on the third clock pulse when both F1TIMA and F1TIMB are set.

spacing control

Inter-element and inter-character spaces are automatically inserted in the code transmission by the disabling of the generator's mark circuitry. This disabling occurs as the circuit detects end-of-element and end-of-character conditions. An end-of-character condition occurs when the mark circuitry has been enabled for the required time to represent the detected element in position one of the registers. At this time a pulse is enabled to shift the next element to position one and to clear the time counter. One time period then elapses between the clock pulse that clears the time counter and the next one that resumes the counting, thus providing a disable for the mark circuit long enough to represent the space.

Inter-character spacing is provided by a three-unit disable of the mark circuit. One of the time units is generated in the normal manner of a space following an end of element; however, detection of the end-of-character condition sets a flip-flop that causes the time period to be extended an additional two time units. This flip-flop, shown in fig. 10, is set when the shift pulse occurs to transfer the last element from the register and clears when the time counter detects a count of two. As indicated by the FOT2 (dot element position 2) and FOH2 (dash element position 1) inputs to the *set* NAND gate in fig. 10, the last element detection actually detects when position two of the register is empty. This condition occurs when the one, which still has an element present. The flip-flop remains set, disabling the inputs to GOMARK and, consequently, the mark circuit, until the NAND gate generating the flip-flop-clear signal is enabled. It is enabled by a clock pulse during a count of two in the time counter when the registers are empty of elements.

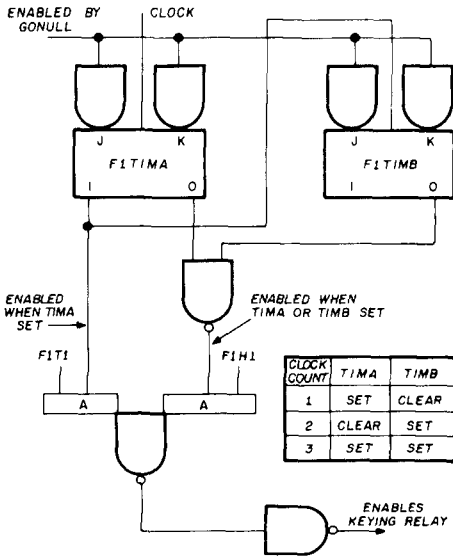


fig. 8. Mark enable logic. See fig. 11 for complete logic diagram.

registers containing a dot, the left AND gate is enabled only during the first count. If counting began with a dash in position one, the right gate would be enabled for all three counts. Thus GOMARK is enabled for one clock period if position one contains a dot and for three clock periods if position one contains a dash. GOMARK, in turn, enables the keying-control circuit, establishing a mark state.

Once a code element has been generated it is necessary to provide the necessary inter-element spacing and to shift the register contents so that the next element is available at position one. Fig. 9 shows the simple three-gate arrangement necessary to determine when it is time for shifting. If the element is a dot, the shift

The time counter is permitted to count for the three-unit period following a register empty period, but is then disabled. At this time the keyboard is again enabled to accept a new character entry.

automatic program sequencer

The automatic program sequencer provides the capability for generating a pre-

count, thus enabling the next sequential output character.

The binary counter only progresses through ten counts since that is the limit of the decoder portion of the logic pair. At the end of ten counts the logic automatically steps to the next logic pair. This action continues until all logic pairs have been accessed, completing the mes-

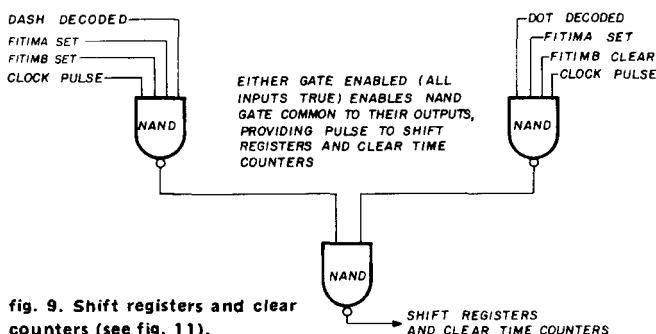


fig. 9. Shift registers and clear counters (see fig. 11).

determined message in response to a key selection. The unit that I built incorporated my call sign and CQ, but any message that suits your own needs can be substituted. The message content is determined by the wiring interconnection between the sequencer's decoder output and the generator's diode matrix encoder.

The sequencer logic is constructed in 10-character groups, with each group sequenced under the control of a pair of logic functions. The logic function pair consists of a four-bit binary counter, to sequence the character transmission, and a 1-of-10 decoder to enable the pre-determined character-select lines for each step in the count.

Each decoder output is connected to the appropriate character-select line in the diode matrix. As the count progresses in the binary counter the individual decoder outputs are sequentially enabled. During the time they are enabled they select a character line into the diode matrix, which encodes the character with dot/dash elements in the manner described previously for the manual operation. Upon completion of a character, the binary counter is incremented to the next

sage. The message can be prepared by depressing the sequencer key again, while holding the key down will provide continuous message transmission.

construction notes

The entire generator was mounted on a single 4½x17-inch (11.4x43.2cm) pre-punched Vector board. Approximately two-thirds of the board was used for the diode matrix layout and the few IC chips of the sequencer. The keyboard was mounted on the breadboard at four corners of its base and covers the matrix area, with the base approximately a half-inch (13-mm) above it. The other five inches (12.7 cm) of breadboard holds the remaining circuitry including the keying relay and connection jacks for the monitor and power supply.

I used a single 6-volt lantern battery to power the generator, although any supply capable of providing 800 milliamps at 6 volts would be adequate. The battery was convenient (and inexpensive), and provided me with approximately a month's service if I remembered to disconnect it when I wasn't using the equipment. There was one lapse in my thoughtfulness

when I inadvertently hooked up a 48-volt supply to the jacks. This converted my integrated circuits to disintegrated circuits and leads me to make two suggestions: First, provide a 6-volt zener at the input to serve as a shunt limiter; and second, use IC sockets if you can locate a reasonably inexpensive source.

The keyboard, which cost only \$14.00,* suited my application perfectly. As you can see from the photograph, it is of modular construction, with plug-in keys and filler spaces that can be positioned along five-row segments. Using the

hole-per-inch breadboard since their pins are spaced at tenth-inch (2.5-mm) intervals. I interconnected the ICs and discrete circuitry with 28-gauge wire, except for the power connections to the ICs which were made with 24-gauge wire. As you will note from the photograph, I stood the resistors, capacitors and diodes on end. This conserved some space and simplified their connections.

conclusion

In my previous article on the automatic fist follower, I mentioned a note

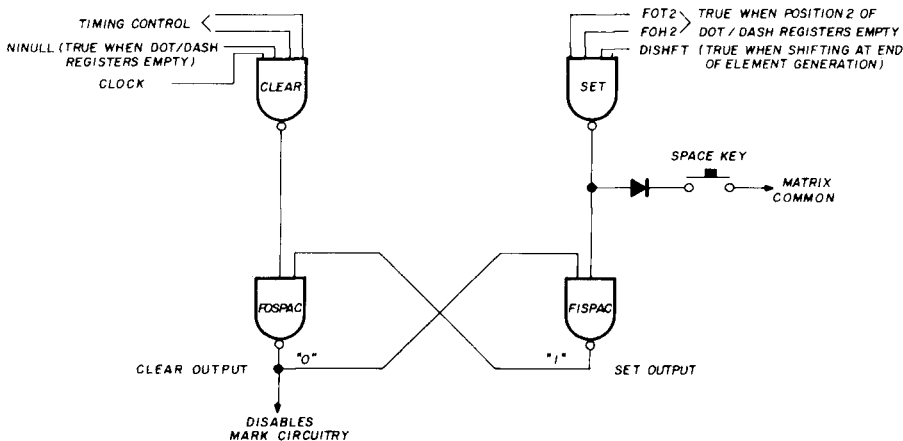


fig. 10. Space control logic (see fig. 11).

plug-in keys and appropriate spaces you can arrange virtually any keyboard pattern you desire. I opted for the standard typewriter keyboard layout, complete with space bar, but with my special-purpose pro-sign keys added. I had to place my own labels over the keys assigned the pro-sign functions, since they are not usually-found options.

The ICs plugged nicely into the ten-

that I had placed over the machine and which I later used on a post card depicting it. The note rather facetiously questioned; "Who needs to know CW?" I now have an answer — my code generator! For all practical purposes (except license exams), I could forget Morse code and not suffer.

After several months of development and testing of the generator I discovered an interesting sidelight to the machine — I was, as expected, sending faster with the generator, but I was also receiving faster by ear. This may have been due to my intense involvement with Morse code during this period, but it led me to consider the potential use of a keyboard with a monitor for Morse code instruction.

*Tri-Tek, Inc., Post Office Box 14206, Dept. H. Phoenix, Arizona 85063. Packets containing schematics, logic drawings, parts lists, theory of operation are available from VMG Electronics, 2138 W. Sunnyside Avenue, Phoenix, Arizona 85029. A similar packet is also available for the Automatic Fist Follower.

approximately 20 wpm. After the design and test period I found my speed had increased to approximately 30 wpm for both send and receive, with a capability of about 50 wpm in the burst mode. I also found when returning to my bug that

The question is intriguing, but probably somewhat academic in view of my daughter's reaction to the entire affair. She observed my painstakingly contrived equipment that converted a touch of the finger to a coded message received by an

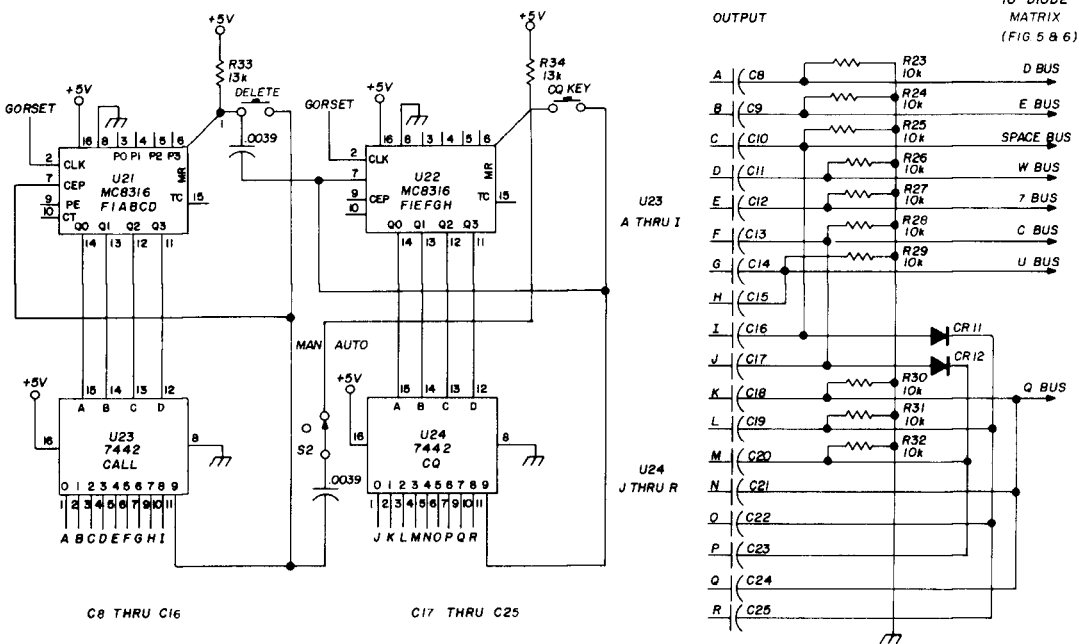


fig. 12. Automatic program sequencer for the keyboard CW generator. All capacitors are 0.0039 μ F, 100-volts. Diodes are 1N914 or equivalent.

I had straightened out some of the undesirable "swing" from my fist.

I then decided to use the generator to teach Morse code to my daughter, who was completely inexperienced in both code and typing. In a short time she was capable of sending and receiving all letters of the alphabet at a rate of 5 wpm, a rate required to pass the Novice license exam. This leads to an interesting question: Would a person knowledgeable of the rules for a Novice license qualify if he appeared with and operated this equipment at the required rate? The rule specifically states that the person shall receive by ear, but an ear can be defined as a sense of hearing. Is the AFF not hearing the code when it translates it?

electronic marvel translating it into printed words. Her response? "Daddy, if you just want to see what you're typing, why didn't you just buy a typewriter?"

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hex inverter

VXO circuit

Basic data for
a variable-frequency
crystal oscillator
plus buffer
covering 2-20 MHz —
all on one
IC chip

Recent experiments with TTL hex inverter ICs in variable-frequency crystal oscillator circuits (VXOs) have been quite favorable, even with simple circuits. The following text presents crystal data, circuit design using the Signetics N7404A (although its equivalent may be substituted), and performance data on VXOs covering 2-20 MHz.

Simplicity, TTL compatibility, and modern design are emphasized.

device data

The following information is presented courtesy of Signetics, Inc., for the benefit of readers who may not have access to data sheets for these devices. Each inverter in the 54/74 TTL series (plastic dual-inline package) has the absolute maximum ratings (in free air temperature) shown in table 1. Some of the features of this logic family are presented below.

Logic definition. Logic is defined in terms of standard positive logic, in which low voltage is logical zero and high voltage is logical 1. Each input requires current into the input at a logical 1 voltage level. This current is 40 μ A maximum for each emitter input.

table 1. Device absolute maximum ratings.

Supply voltage, V_{CC} *	7V
Input voltage, V_{in} *	5.5V
Operating free-air temperature range:	
Series 74 circuits	0°C to 70°C
Storage temperature	-65°C to 150°C

*Voltage values are with respect to network ground terminal.

Bill King, W2LTJ, 5 Midwood Drive, Floham Park, New Jersey 07932

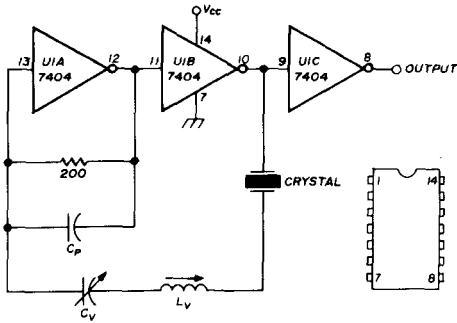


fig. 1. Hex inverter Vxo using the Signetics N7404A hex inverter IC.

The switching characteristics of this IC family are typically 8 and 12 ns to logical zero and logical 1 voltage level, respectively.

Clamping diodes. All devices in the Signetics 54/74 logic family incorporate input diodes. Each clamping diode will limit negative excursions at the input to 1.5 V maximum below ground, even if -12 mA of current is drawn.

vxo circuit description

Three inverters are used per oscillator circuit (fig. 1). Two form the oscillator and the third is the output buffer. Positive 5 Vdc at 50 mA is connected to pin 14; negative is ground (pin 7).

Three inverters are connected in series: pin 12 joins pin 11, pin 10 connects to pin 9, and pin 8 is the output connection. A 200-ohm (linearizing) resistor is connected between pin 13 and the junction of pins 11 and 12. A frequency-limiting capacitor, C_p , shunts this resistor. The crystal is connected in series with an inductor and a variable capacitor between pin 13 and the junction of pins 9 and 10. Starting with pin 13, you should have first a variable capacitor, C_v , an inductor, L_v , the crystal, and finally a connection to pin 9.

The N7404A circuit performs in somewhat the same way as the crystal impedance meter (CIM) described below. The crystal is in series with two inverters, thus the circuit oscillates best when the crystal appears as a pure low resistance. The situation is similar to a dog biting its own tail: the two inverters are the dog (who bites harder the sharper his teeth and the stronger his bite). You guessed it. The crystal is his teeth.

test crystals

Table 2 lists the crystals used in the investigation. The parameters shown were measured on a crystal impedance meter, which is a special circuit that can tune the crystal to its series-resonant

table 2. Crystal impedance meter data on selected test crystals. Crystals E and H are designed for third-overtone operation at 19.825 and 27.255 MHz. Crystal F is for fifth overtone at 33.825 MHz. All remaining crystals are fundamental mode.

source, type	cut	resonances (kHz)		C_o (pF)	R (ohms)	Q ($\times 10^{-3}$)
		series	32 pF anti			
A Military FT-171-B	BT	2219.179	2219.998	19.0	14	136.0
B International Crystal HC6/U	AT	2998.055	2999.159	3.6	22	92.0
C International Crystal HC6/U	AT	4998.433	5000.028	5.3	4	334.0
D Keystone Electronics FT 243	BT	5997.536	5999.495	17.2	6	138.0
E Military CR-9/U	AT	6610.817	6612.788	5.5	40	27.0
F Military CR24/U	AT	6763.373	6764.408	5.5	30	68.0
G International Crystal HC6/U	AT	8996.756	8999.661	5.2	4	184.0
H James Knight Co., channel 23 HC6/U	AT	9088.159	9089.804	3.8	40	33.0
I RCA HC6/U from fm receiver	AT	11647.911	11654.463	10.2	3.5	82.0
J International Crystal HC6/U	AT	14986.986	14991.926	4.8	2	219.0
K International Crystal HC6/U	AT	19982.309	20000.020	6.2	2	59.0

frequency. This is the frequency where the crystal appears to the circuit as a pure resistor of low value. When so adjusted, the crystal is switched out of the circuit and a resistor substituted for it. When the circuit amplitude equals that of the crystal, the resistor is said to be equal to the series resistance of the crystal, since the circuit can't tell if the resistor or the crystal is in its feedback loop.

The antiresonant frequency is determined in the same way, except that a 32-pF capacitor is now in series with the crystal. The value of C_0 in the table is the capacitance between the two pins that includes the holder capacitance as well as the interelectrode capacitance of

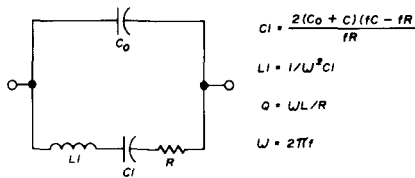


fig. 2. Equivalent circuit of a quartz crystal. The capacitance C_0 is the static capacitance and is primarily a function of the electrodes sandwiching the quartz. C_1 , L_1 and R are the equivalent elements of the motional arm.

table 3. Crystal frequency (kHz) in the circuit of fig. 1 vs shunt capacitance, C_p . L_v is zero and C_v is short-circuited.

C_p (pF)	0.0	15	100	470
A	24399.86	---bad---	2219.07	2218.49
B	8924.61	2999.37	2999.31	2999.10
C	4998.30	4998.27	4997.42	0.00
D	5997.27	5997.12	5995.36	0.00
E	19824.41	6611.33	6609.33	0.00
F	20298.39	6763.76	6761.99	0.00
G	8996.01	8995.74	0.00	0.00
H	9089.71	9089.47	0.00	0.00
I	11644.59	11642.35	0.00	0.00
J	14986.36	14984.60	0.00	0.00
K	19978.93	---bad---	0.00	0.00

Note: Bad means an erratic output; counter readings bore no relationship to the crystal under test.

the crystal. The value of Q was calculated from the above data by:

$$C1 = \frac{2(C_0 + 32)(f_{32} - f_R)}{f_R}$$

$$L1 = \frac{1}{\omega^2 C1}$$

$$Q = \omega L/R$$

$$\omega = 2\pi f$$

Where f is the frequency and the other components are as shown in fig. 2.

table 4. VXO frequency (kHz) as a function of circuit parameters.

C_p (pF)	15	15	15	15	15	15
C_v (pF)	inf	1.0	2.0	7.8	25.5	100
L_v (μ H)	0	0	0	0	0	0
G	8995.74	9011.46	9009.57	9004.17	8999.71	8997.13
H	9089.47	0.00	9099.66	9094.98	9091.70	9090.11
I	11642.35	11667.03	11665.98	11660.70	11652.80	11646.09
J	14984.60	15010.24	15006.96	14998.28	14990.11	14986.25
K	19932.75	20021.41	20012.50	20001.67		

C_p (pF)	15	15	15	15	0.0	0.0
C_v (pF)	7.8	100	7.8	100	25.5	1.0
L_v (μ H)	7	7	13	13	0	0
G	9003.08	8995.33	9001.92	8992.33	8999.93	9011.16
H	9094.12	9089.33	9093.23	9087.11	9091.92	
I	11658.65	11635.35	11654.20	11592.61	11653.78	11666.94
J	14994.34	14973.37			14992.02	15009.81
K	19982.30				19992.39	20021.24

Note: A blank space means a counter output completely unrelated to the crystal — probably due to noise. A similar reaction can be obtained with a small capacitor replacing the crystal. (Counter was capable of counting to 33 MHz.)

equivalent circuit

The equivalent circuit for a crystal (fig. 2) is the parallel combination of C_0 with the series string of R, C1 and L1. From this equivalent circuit it is now apparent that, at series resonance, the positive inductive reactance of L1 is

The N7404A is basically a high-frequency device. To use it for low-frequency service the high-frequency response must be inhibited by bypassing one of the inverters with a low-value capacitance. The effects of this are shown in table 3. Crystal A, you'll note,

table 5. Circuit Q as a function of VXO shifting.

C_v cap (pF)	with C_v and no L_v		with C_v and L_v			
	freq (kHz)	Q ($\times 10^3$)	$L_v = 13 \mu H$		$L_v = 17 \mu H$	
			freq (kHz)	Q ($\times 10^3$)	freq (kHz)	Q ($\times 10^3$)
1.0	9011.22	21.7	9009.56	19.4	9007.72	16.0
2.0	9009.38	28.3	9007.17	20.4	9005.95	27.3
4.0	9007.03	49.1	9005.17	32.0	9003.33	33.5
8.0	9004.39	61.3	9001.55	49.1	8999.79	40.9
16.0	9001.14	81.8	8998.14	73.6	8994.64	54.5
32.0	8999.14	147.0	8995.32	92.0	8991.83	56.6
100.0	8997.56	147.0	8993.43	92.0	8987.69	53.0
∞	8996.76	184.0	8991.54	105.0	8983.92	64.0

Crystal tested: $f_R = 8996.756$; $f_{32} = 8999.661$; $C_0 = 5.2$ pF;
R = 4.0 ohms

table 6. AT-cut crystals are better than BT cut for VXO applications. Here are N7404A VXO data on two-meter crystals.

series elements		frequency shift from base point	
capacitance (pF)	inductance (μH)	AT cut (kHz)	BT cut (kHz)
∞	13	-7.271	-2.774
100.0	13	-5.715	-1.771
∞	7	-5.307	-1.324
100.0	7	-4.010	-0.835
∞	0	-3.456	-0.829
100.0	0	-2.352	-0.524
27.0	0	0.000	0.000 (base point)
7.8	0	4.327	0.575
2.0	0	8.954	0.948

Notes. (-) means below base point. No sign is above base point. At base point both are 8081.675 kHz. Other parameters are:

cut	holder	F32 (kHz)	FR (kHz)	R (ohms)	C_0 (pF)
BT	FT-243	8081.649	8081.038	13	13
AT	HC6/U	8081.703	8079.103	12	4.2

exactly equal and opposite to the negative reactance of C1. The circuit is now only a resistor, R, and a capacitor, C_0 in parallel, since the resistance is very low, the crystal now appears as a low-value resistor (since the few pF in C_0 are swamped).

doesn't behave itself until the inverter is shunted with 100 pF. With zero pF, crystal A took off on its eleventh overtone, which would be fine if we wanted it because it was very stable. Crystals E and F are overtone crystals by design: they are tamed to their

fundamental mode with only 15 pF. The rest of the table is self-explanatory and should be studied before applying the circuit to any particular crystal situation you have in mind.

frequency shift

It makes sense to VXO only the higher-frequency crystals, since they can be moved in meaningful and useful amounts without getting into instability problems. **Table 4** shows the magnitude of frequency shift you can expect from

frequency crystals, of course, can be shifted less.

stability considerations

The reason we use crystals to control frequencies is simply because they are stable, high-Q devices; and when you VXO them you pay for it through the loss in effective Q in the circuit. When the VXO act is carried to extremes, you no longer have a crystal in charge of the frequency, and the circuit promptly becomes an old LC deal with the usual

table 7. Capacitor values calculated from calibration correlation for crystal impedance meter (CIM) and N7404A circuits using CIM experimental data.

experiment capacitor (pF)	predicted capacitor		correlated frequency (kHz)	experiment frequency (kHz)
	CIM (pF)	N7404A (pF)		
1.0	1.40	1.26	9012.052	9011.123
2.0	2.11	2.09	9009.916	9009.718
4.0	3.90	3.80	9007.036	9007.145
7.8	7.43	7.40	9004.005	9004.217
32.0	32.70	33.8	8999.232	8999.181
100.0	93.20	180.20	8997.574	8997.637
∞	-705.00 (0.44 μH)	-178.00 (1.8 μH)	8996.667	8996.531

The other parameters are $C_0 = 5.2$ pF, $R = 7.5$ ohms, and $Q = 34,160$. The correlation coefficient is 0.9996; the equation is:

$$\text{freq (kHz)} = 8996.667 + \frac{95.388}{C_0 + C}$$

various crystals by changing the series reactances. The data for 1 pF are probably not within reach of most users, since a 100 pF variable will probably be used, and this low value is not obtainable. However, the data clearly demonstrate the value of reduced stray capacitance, since most of the easy-to-get shift comes on the low capacitance side. To shift the frequency below the f_R , the series-inductive reactance must predominate.

Consider crystal J. Using an inductor of 7 μH and a 100- to 2-pF variable, the possible frequency shift is about 25 kHz above, and 13 kHz below, series resonance. With care you can VXO a 15-MHz crystal about 50 kHz. Lower-

problems. So consider **table 5** and be your own judge. A circuit Q of 20,000 is good to stay above, but it is not the end of the line. You trade a little, so if you are aware of what you are doing, you can get away with it under the right conditions.

two-meter 8-MHz crystals

There are two kinds of crystals available to amateurs, the BT and AT cut. The old surplus variety (still around) in the FT-243 holders is most likely BT cut. The AT cut crystal is better for VXOing. However it is more expensive and is the crystal most available today from most manufacturers. In many cases crystal suppliers won't even talk to

you about the antique BT cut. **Table 6** shows the amount of shift you can expect with two-meter 8-MHz crystals of both types.

calculating frequency from series capacitance

Here is a way to calculate the frequency of a VXO from the series capacitance. The N7404A IC has almost the property of a series tuned circuit without the knob; maybe you noticed this when scanning the data tables. I decided to see how close it really was and made measurements of crystal-capacitor combinations on both an honest series-resonant circuit and the N7404A circuit. The instruments used were a Radio Frequency Labs Model 459A crystal impedance meter (CIM) and General Radio Model 1612AL capacitance meter.

To get an easy-to-visualize table of comparison, I took the data obtained from both circuits and calculated a capacitor value from the correlation of reciprocal ($C_0 + C$) vs frequency, which is a straight line for all crystals. The intercept is the series-resonant frequency, and the slope is a function of the resonator cut, etc.

The line for any crystal can be determined (as it turned out) by measuring the crystal frequency in the N7404A circuit with two different capacitors in series, preferably one about 5 pF and one not larger than 32 pF. Since each crystal, even with the same cut and frequency, will be an individual with its own slope, don't expect one line to work too well for other crystals. But the same performance can be expected, so it is a useful guide. **Table 7** shows that it is reasonable to predict frequency from a capacitor with the N7404A circuit. Of interest to those of you with a counter: you can measure low-value capacitors with a calibrated crystal (two known points are all you need if you know C_0 ; if not, you need three points).

ham radio

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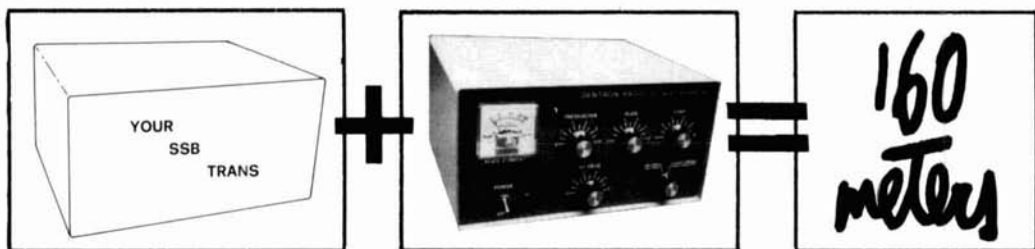


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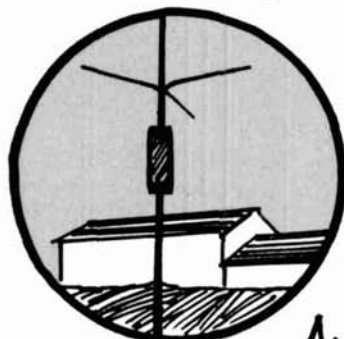


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general purpose wideband rf amplifier

Design and construction
of a two-stage,
small-signal amplifier
that may be used
in the frequency range
from 500 kHz
to 500 MHz

The basic building block of practically all electronic circuits is the amplifier. Each time a circuit is designed and breadboarded the amplifiers in that circuit must be designed, tested, trouble-shot and, sometimes, redesigned. I have often thought that there had to be a better way. For many requirements the answer lies in the circuit described here. It is a standard gain block capable of operating from typical intermediate frequencies to nearly 1 GHz. It is simple, inexpensive and amplifies without oscillating!

The design goals for this gain block were high gain, wide bandwidth, unconditionally stable operation, low noise figure, simple construction and low cost. It is not possible, however, to obtain state-of-the-art performance in all these categories in one amplifier.

What is achieved with the circuit in fig. 1 is a good compromise. The amplifier will perform well as an rf amplifier, i-f amplifier or general purpose preamp.

transistor selection

The selection of transistors is very important in high performance amplifiers. The 2N5179 transistor used in this amplifier provides high gain, large bandwidth and low cost. At 150 MHz a typical device will provide nearly 14 dB gain when inserted in a 50-ohm transmission line with no tuned circuits on the input or output. Stability at this heavy loading is guaranteed. The noise figure for this device, with optimum source resistance, is about 3 dB at 150 MHz. The 2N5179 sells for \$0.77 in small quantities. For further information see reference 1.

circuit

The amplifier circuit is a two-stage, capacitively-coupled, common-emitter cascade. The collector-base feedback resistors are key elements in this circuit. They provide a simple means of bias, reduce gain at low frequencies to stabilize the circuit, and lower both the input and output impedance of the amplifier. A lot of work for two resistors! Capacitor C3 decouples each stage and R5 and C5 decouple the amplifier from the supply line.

The 2N5179 transistors have a rather wide specification in dc current gain (beta). With the simple bias network used in the circuit the bias point may therefore vary from transistor to transistor, and you may find it necessary to adjust R5 so that the voltage drop

Randall Rhea, WB4KSS, 1560 Jennings Way, Norcross, Georgia 30071

across R2 and R4 is 3 to 4 volts, each. This corresponds to about 8 mA collector current. This operating current is a compromise, with higher currents giving increased gain-bandwidth and lower currents providing improved noise figure.

The capacitors in the circuit cause the low-frequency gain to begin dropping off below about 2 MHz. Increasing these capacitors to 0.01 μF will lower the frequency response to about 200 kHz. It might be possible to use the amplifier at even lower frequencies by using larger values of capacitance, but I have not investigated the stability of the amplifier when doing this.

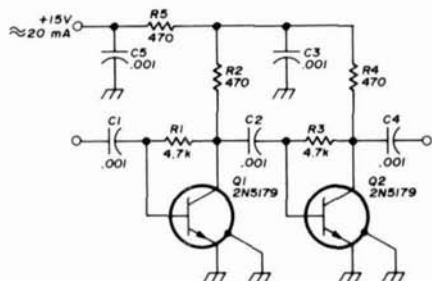


fig. 1. Schematic diagram of the wideband rf and i-f amplifier. To lower the frequency response to 200 kHz, increase the value of all capacitors to 0.01 μF . All resistors are $\frac{1}{4}$ watt, all capacitors are ceramic disc.

construction

Mount the transistors close to the PC board, fig. 2, to keep the emitter leads very short. Be sure to ground the transistor case (fourth lead). This reduces unwanted capacitive coupling between transistor elements. If the PC board isn't used keep all leads as short as possible. In one breadboard I used no terminals, soldering each component to the next with very short leads. The result was a circuit just over one-inch (25mm) square.

A full-size printed-circuit layout is shown in fig. 3. Connectors may be

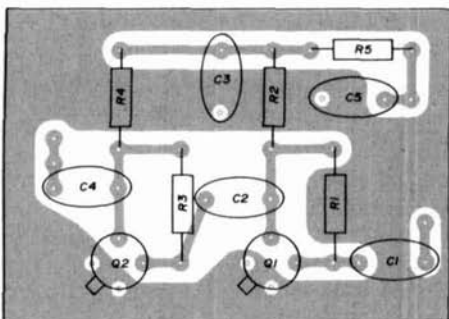


fig. 2. Printed-circuit component layout for the wideband rf amplifier. Full-size printed-circuit layout is provided in fig. 3.

soldered onto the board with short wire jumpers for the center conductor. Extra holes in the board are for these jumpers and for the +15 volt supply line. Coaxial input and output lines may be soldered directly onto the board if desired.

performance

The gain of the amplifier vs frequency is plotted in fig. 4. It is essentially the same for 50 or 75 ohms input and output loading. The measured input impedance is about 65 ohms from 500 kHz to 500 MHz with only a small series reactance. The output impedance is around 100 ohms series resistance with some reactance. The gain of fig. 4 is with 50 or 75 ohms driving and load impedance, there being little difference between the two. The maximum output level will vary some from amplifier to

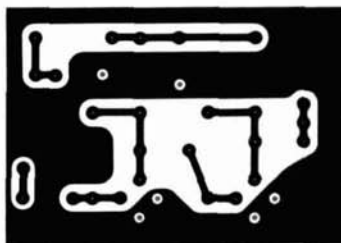


fig. 3. Full-size printed-circuit layout for the wideband amplifier.

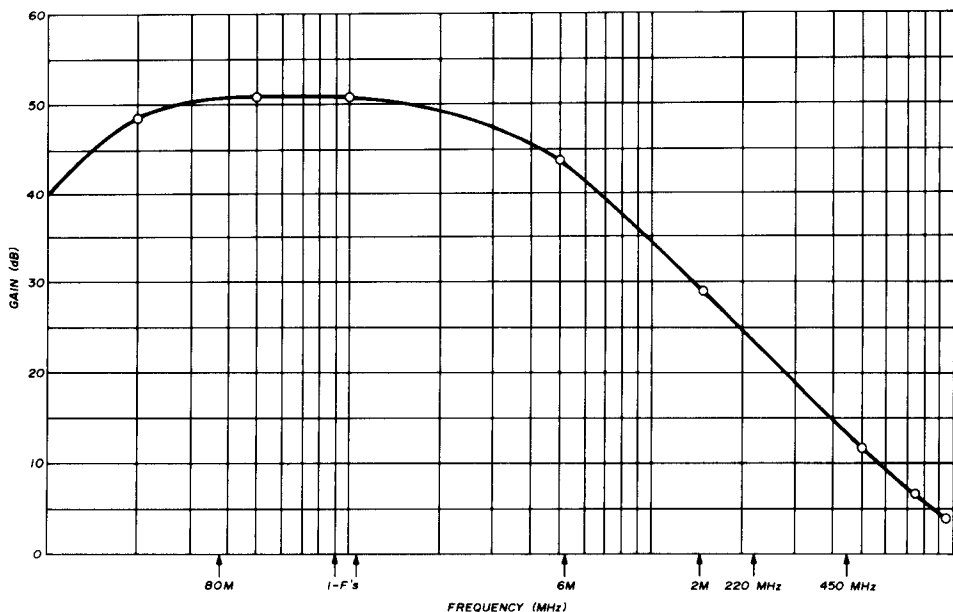


fig. 4. Frequency vs gain of the wideband rf and i-f amplifier shown in fig. 1. Circuit provides useable gain to 500 MHz and above.

amplifier because of small variations in transistor biasing. A typical value of maximum output level for 1 dB gain reduction is 300 mV.

The gain of this circuit should be treated with respect. When making connections to the board, keep the input and output leads separated. The two stages are identical so if the total gain isn't needed, connect only one stage and adjust resistor R5 for 3 to 4 volts drop across R2. I have built several of these amplifiers and none have oscillated, even with the input open, but it is a good idea to use a little caution.

tuned amplifier

Unless the amplifier will be used specifically as a broadband amplifier, tuned input and outputs should be used. This will reduce the possibility of undesired signals overloading the amplifier and will provide the desired bandpass characteristics. Since the input and out-

put are closely matched to 50- or 75-ohm systems the tuned circuits are used only to shape the frequency response.

The Q of a tuned circuit is a measure of its 3 dB bandwidth. Consider the tuned circuit in fig. 5. The two 50-ohm resistors represent the generator resistance driving the amplifier and the amplifier input resistance. The total parallel loading, R, of the tuned circuit in fig. 5 is 25 ohms. The loaded Q is given by

$$Q = \frac{R}{X}$$

where X is the reactance of the inductor or capacitor at resonance. This is ap-

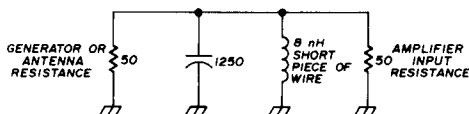
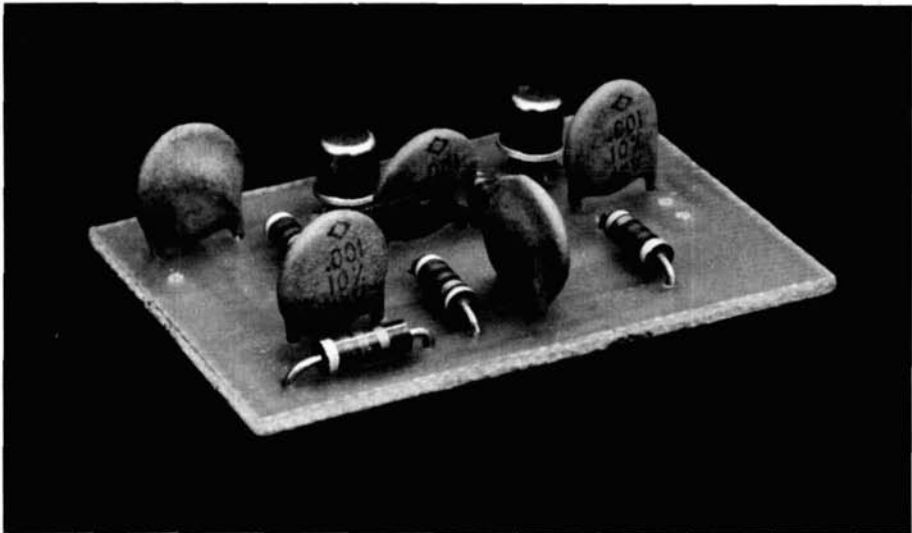


fig. 5. Model of an input tuned circuit for the wideband rf and i-f amplifier (see text).

proximately true if the loaded Q given above is lower than about one-tenth the Q of the inductor or capacitor. The Q of the circuit with no external loading is the unloaded Q . It is desirable to make the unloaded Q as much greater than the loaded Q as possible. In practice this means higher Q coils and wider loaded

To tune the circuit either the capacitor or inductor may be made variable. At high or low frequencies it might be difficult to obtain the proper reactance with reasonable values of L and C . In this case the inductor may be tapped or a capacitive divider may be used to increase the effective resistance of the



Complete wideband amplifier, shown here almost twice full size, is built on a 1.75x1.25" (45x32mm) printed-circuit board. In this view transistor Q1 is at left rear, Q2 to the right.

bandwidths. Narrowband tuned circuits with low quality coils and capacitors mean high loss in the tuned circuit. For further information on this subject see reference 2.

The resonant frequency of the tuned circuit in fig. 5 is 50 MHz. The reactance of the inductor at resonance is 250 ohms, so the loaded Q is 10. The 3 dB bandwidth is given by

$$BW = \frac{f_o}{Q_{loaded}}$$

and in this case is equal to 5 MHz. Since the inductor and capacitor should have an unloaded Q of at least ten times the loaded Q , it should be at least 100 for this case.

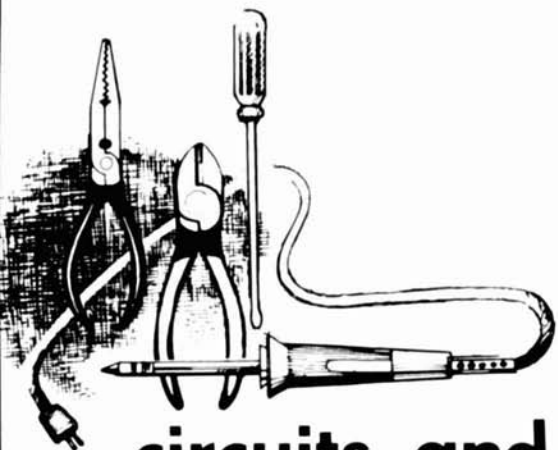
load. (For further reading on this subject see pages 160 and 161 of reference 2.)

Adequate space is left at the input and output of the circuit board for compact tuned circuits. The coils at the input and output should be placed at right angles to avoid feedback caused by mutual coupling. Better still, the use of toroids will eliminate this problem and provide compact coils.

references

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2. *The Radio Amateur's Handbook*, 50th Edition, American Radio Relay League, Newington, Connecticut, 1973, page 41.

ham radio



circuits and techniques

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vhf/uhf single-frequency conversion

The manufacturers of commercial radio communications equipment are not only giving thought to the advantages of single conversion, they are using it in two-way fm gear (fig. 3). In most commercial (and practically all amateur vhf units) the dual-conversion technique is common with its high and low i-f

narrow-band second i-f. Recently, with the development of crystal filters for the high i-f range, the selectivity of the high i-f has been vastly improved. In addition, the intermodulation characteristics of receivers have been improved through the use of fet mixer stages.

The above improvements suggest that not much thought be given to the old-fashioned single-conversion receiver. However, modern monolithic crystal techniques suggest a second look. What are some of the problems of dual conversion? Additional spurious responses for one thing, possible because of the presence of a second mixer. Also, the signal level is rather high at the second mixer and strong, nearby interfering signals can desensitize the receiver. Although the interfering signal may not be heard, it operates the limiter and/or automatic gain control system and reduces receiver sensitivity.

A modern, single-conversion design, on the other hand, offers several advantages. In addition to eliminating some of the problems caused by the dual-conversion system, the single-conversion design provides added reliability and simplified receiver alignment. It can also provide a notable size reduction — this is especially important in modern portable equipment.

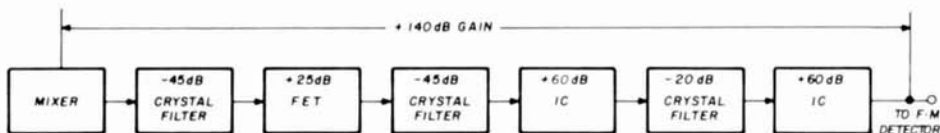
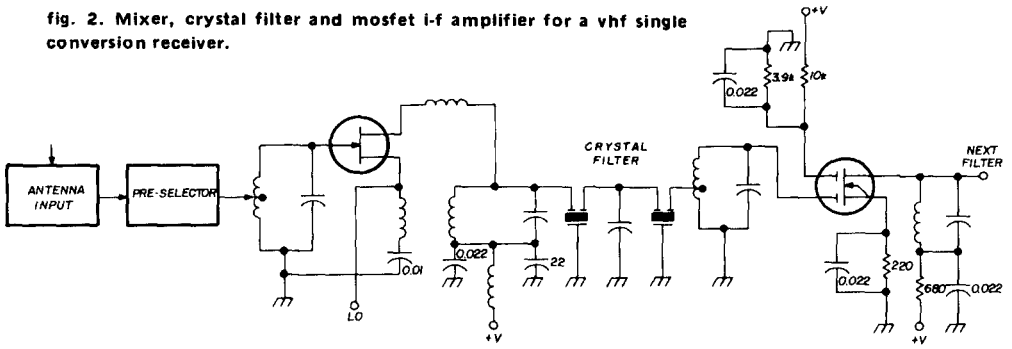


fig. 1. Block diagram of a single-conversion receiver showing the use of monolithic crystal filters and IC gain blocks.

sections. The high intermediate frequency is responsible for reducing image responses, preventing them from reaching the second mixer. Gain and selectivity are the responsibility of the

Modern monolithic techniques provide an answer to the two basic single-conversion problems of image rejection and instability which result from the necessary high gain required at the

fig. 2. Mixer, crystal filter and mosfet i-f amplifier for a vhf single conversion receiver.



single i-f frequency. The structure of the monolithic crystal filter design provides high stability, reliability and excellent selectivity. Thus the single-conversion i-f can be made relatively high so that

incorporated in the single-frequency i-f system shown in fig. 1.

The *non-monolithic* crystal filter with which most amateurs are familiar contains numerous discrete inductors,

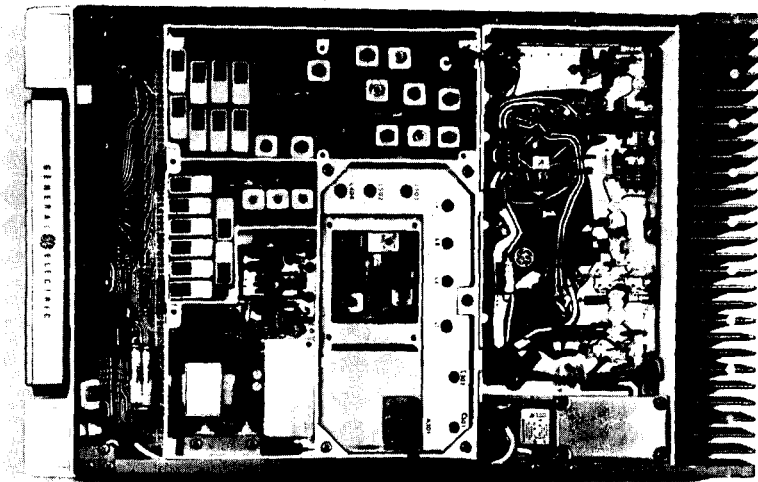


fig. 3. The GE Master II commercial vhf fm radio uses a single-conversion receiver (photo courtesy General Electric).

image response is reduced, just as it is in dual-conversion receivers with high intermediate frequencies. High-gain performance is obtained by the unique arrangement of filter and gain blocks

resistors and capacitors in addition to the crystal. These components are used primarily for coupling, but they must be mounted and then carefully aligned. The *monolithic* crystal filter, however,

incorporates coupling as an inherent part of the quartz assembly (accomplished by placement of electrodes). The electrodes provide coupling and a means of converting an electrical signal

frequency limit for design of such filters to accommodate a 15-kHz bandwidth is approximately 5 MHz. At the opposite end of the spectrum the practical limit at today's state of the art falls in the 25-

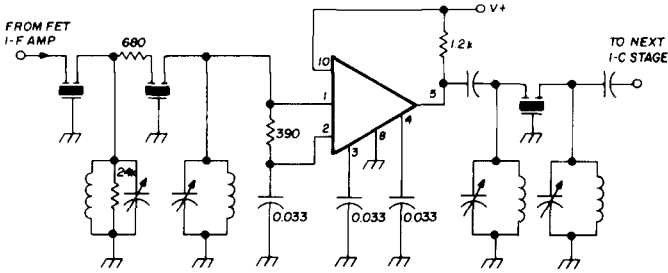


fig. 4. IC i-f amplifier with monolithic crystal filters at the input and output.

at the input to a self-resonant vibration of acoustic energy through the crystal-line material. Acoustic energy is converted back to electric signal at a resonant output converter. Selectivity and coupling are controlled by the electrode geometry of the quartz blank.

Pairs and other groupings of these basic monolithic crystal filters can be assembled in modules and used to form higher-order filters. Presently the low-

to 30-MHz range. (Although monolithic crystals have been made to operate to near 200 MHz, the cost is extravagant.)

Integrated circuits fit in ideally with the use of monolithic crystal filters. An integrated-circuit gain block can provide high stability with 60-dB gain and provide adequate performance up to 20 MHz or so with no instability and regeneration problems in sensible designs. Note that a single-conversion gain

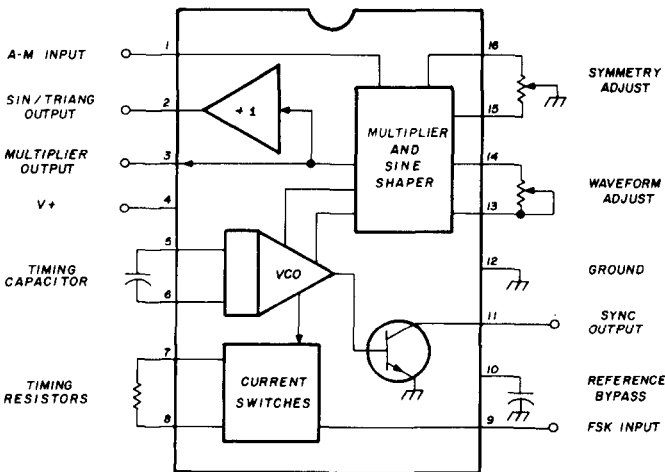


fig. 5. The Exar 2206C function-generator IC.

of 140 dB can be obtained with the use of three crystal filters and three IC gain blocks. The staggering of crystal frequencies and the alternate gain-loss plan provide a high-gain overall bandwidth with low noise content.

Keeping the noise figure down in the i-f section minimizes the requirement for receiver front-end gain. As an aid in this objective the first amplifier of the i-f amplifier can be a field-effect transistor.

In practical vhf/uhf single-conversion operation a good i-f choice would be in the 10- to 12-MHz range. This matches well the standard 10.7-MHz i-f. When using a much higher i-f there may be problems with low-order spurious components, especially if the receiver is to operate over the 25- to 60-MHz range. At lower intermediate frequencies there is an additional cost factor and the 5-MHz limit of present monolithic crystal-filter science.

A simplified schematic of the mixer and first i-f amplifier is shown in fig. 2. The vhf signal from the antenna is applied to a matching input transformer and then to a five-section helical pre-selector that sets the rf selectivity of the receiver. The output of the preselector is applied to the gate of an fet mixer. Local-oscillator power is injected to the source of the fet.

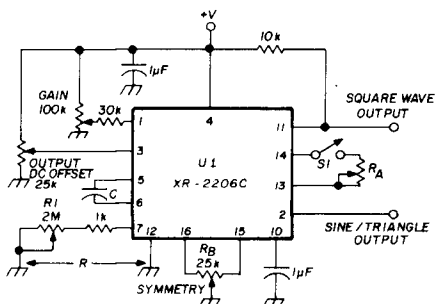


fig. 6. Simple sinusoidal generator using the XR-2206C IC provides sine, triangle or square-wave outputs. Switch S1 must be closed for sine output.

The drain output is linked to the input of the first monolithic crystal filter. An LC network does the appropriate matching to the four-pole filter. A second network matches the output

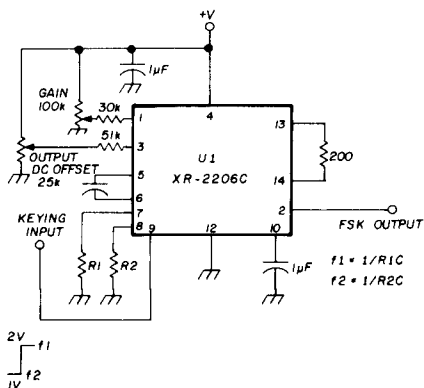


fig. 7. Simple FSK generator using the XR-2206C function-generator IC. Mark (f1) and space (f2) frequencies are given by the equations. The keying input is applied to pin 9.

of the crystal filter to the input of the i-f amplifier. The intermediate frequency is 11.2 MHz and i-f gain is 20 dB.

The first i-f amplifier is followed by a second crystal filter along with the input and output impedance-matching networks, fig. 4. This is followed by an IC i-f amplifier with a gain of 60 dB. The integrated circuit is similar to the RCA CA3014 wideband amplifier IC. No agc system is required. Dynamic signal range, freedom from significant intermodulation distortion components and limiting action down to minimum signal level eliminate this circuit function.

function generator

In the October, 1972, column the Exar 205 and S200 function generators were introduced. These integrated-circuit generators can form a number of basic sine, ramp and pulse waveforms. Generated waveforms can also be ampli-

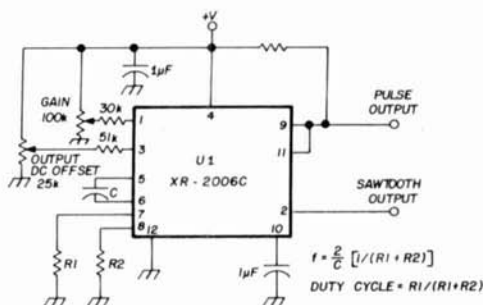


fig. 8. Pulse and sawtooth generator using the XR-2206C function-generator IC.

tude or frequency modulated. Exar now makes a low-cost generator, the Exar-2206C, shown in fig. 5. Similar to the previous function-generator ICs, it comprises four function blocks — voltage-controlled oscillator, multiplier and shaper, buffer amplifier, and current switches. The upper frequency limit is about 1 MHz, less than the previous devices. However, the device has greater stability with temperature change and

can be swept over a larger frequency range.

As with the previous devices, only a few external components are needed. A sine-wave generator, fig. 6, can be adjusted to provide a pulse, sinusoidal or triangle wave output. The output frequency is determined by the value of the capacitor connected between pins 5 and 6 and the resistance connected between pin 7 and common. The output level is set by the gain control while the offset dc voltage at pin 2 can be adjusted with the 25k offset potentiometer connected to pin 3.

To obtain a sine-wave output with minimum harmonic content switch S1 must be closed. Potentiometers R_A and R_B are then adjusted for minimum distortion in the sine-wave output. The output frequency is given by

$$f = 1/RC$$

where R is the approximately 2 megohms from pin 7 to ground, and C is connected between pins 5 and 6.

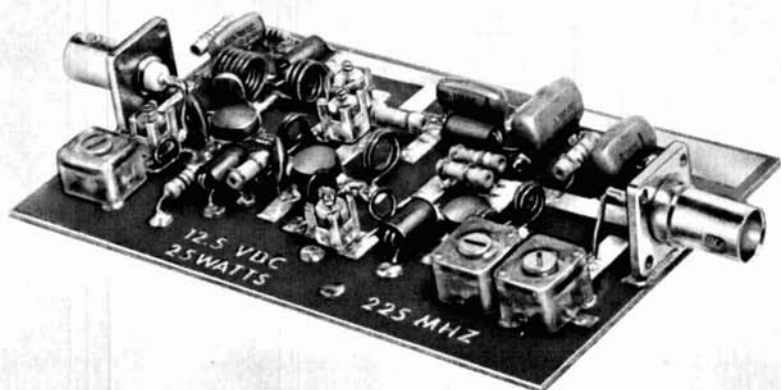


fig. 9. A 25-watt amplifier for 225 MHz which requires 100 mW of drive. Complete circuit of this unit is shown in fig. 10 (photo courtesy Amperex).

A frequency-modulated output can be obtained by applying a modulating wave to either pin 7 or pin 8. Amplitude modulation is obtained by applying the modulating wave to pin 1.

The device can also be used as a

vhf power amplifier

Amperex has announced a 25-watt, 225-MHz power amplifier module designed for fm circuits, **fig. 9**. This module uses the 2.5-, 8- and 25-watt

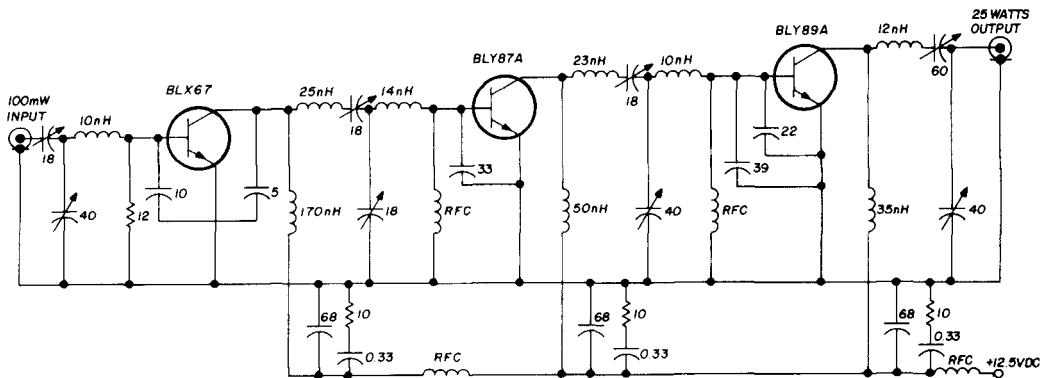


fig. 10. Three-stage solid-state 25-watt power amplifier for 225 MHz (see **fig. 9**).

simple FSK generator using the circuit of **fig. 7**. In this case the *mark* and *space* frequencies are determined by resistors R1 and R2. The actual FSK keying waveform is applied to pin 9. The mark and space frequencies (f_1 and f_2) are given by

$$f_1 = 1/R_1C$$

$$f_2 = 1/R_2C$$

A circuit for using the 2206C as a pulse and sawtooth generator is shown in **fig. 8**. In this mode pin 9 is shorted to the squarewave output at pin 11. The ramp rise and fall times or pulse duty cycle are regulated with resistors R1 and R2. The pulse width and duty cycle can be adjusted from 1% to 99% by proper selection of values for resistors R1 and R2 as given by

$$f = 2/C \left(\frac{1}{R_1 + R_2} \right)$$

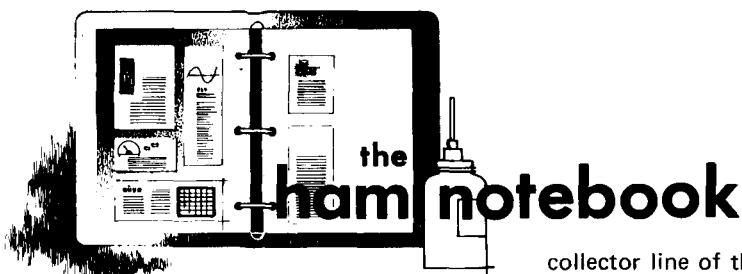
$$\text{duty cycle} = \frac{R_1}{R_1 + R_2}$$

Amperex semiconductors shown in the schematic diagram of **fig. 10** and features 50-ohm input and output. With a 100-mW input signal, the module will deliver 25-watts output. The circuit is straightforward, using four capacitive dividers for input, output and interstage matching. The collectors are shunt-fed and three decoupling networks avoid self-oscillation. The amplifier can withstand an output mismatch as high as 50:1 without harm.

metric tapes

Some months ago I mentioned the convenience of a metric tape for calculating antenna and transmission-line lengths. Steel tapes with lengths up to 100 meters can be purchased from Forestry Suppliers, Box 8397, Jackson, Mississippi 39204. Standard lengths of 3, 20, 30, 50 and 100 meters are available. Some models are calibrated in feet and inches as well.

ham radio



more power from the Standard 826M

Recently I offered to change the final transistor in a Standard 826M for a RTTY enthusiast friend. The output stage had "passed on," not from accidental abuse, but from overwork. Duty cycle or not, 40 minutes of RTTY was a normal day's work, before breakfast!

When I received the unit and instruction manual, I began to study the schematic. Companies such as Standard, that also produce marine and commercial land mobile gear, tune the slugs a little further into the coils and call the set an amateur radio. This is good in many respects. It not only assures that the design is fairly sound; it also lends itself to the plagiarizing of features from one model to another.

The case in point is the power selector system. This feature was originally designed into the marine units by FCC directive. Marine vhf sets must have a power selector system which will select either full power out or slightly less than 1 watt out for inner harbor communication. When this feature was incorporated into the amateur line, it was labeled as a "battery-saver circuit."

The circuitry is similar to the tune/operate switch in older tube and hybrid radios which reduces power output during tuneup. In the Standard, the power selector switch inserts a resistor in the

collector line of the final and predriver. Changing the value of this resistor, within reason, makes changing the power output and increasing the duty cycle easy.

My decision on what level of power output to allow was hampered by several obvious and other, not-so-obvious, constraints. First, physical constraints: as the allowed power rises, the value of dropping resistor increases in size due to the higher power dissipation. Second, electrical limitations: the wire connecting the dropping resistor to the front panel switch is not suitable for more than several hundred milliamperes.

After some bench and field tests, an output of 2 to 3 watts in the low power position was found to be a satisfactory level. By Ohm's law, decreasing the resistor by a factor of two doubles the current and power output. In the Standard 826M the original resistor value of R005 is 22 ohms. The same value is used in the 806M; in the newer 826MA, the value of R363 was reduced to 15 ohms (consult individual instruction manual for value and position).

The resistor will probably be located across the two 9-pin test jacks on the rear of the chassis or on the transmitter board between the chassis wall and output transistor heatsink. A word of caution when working around the final area: take care not to deform the air-wound coils; sometimes they don't bend back. I know from experience!

Simply halving the present value will

produce a 2-watt output. Reducing the value slightly again will bring this to 3 watts. Reducing the resistance below this will increase the heat dissipation of the resistor to too great a value. The replacement resistor should be a Brown Devil or similar type, with a 12 watt, or slightly higher, rating. Mount the replacement in exactly the same fashion as the original unit.

This small modification adds much to the versatility of the 826. First, it provided the longer duty cycle originally sought. Second, some of the new 2-meter bricks need only 2 to 3 watts in to give 40 or more watts out. This means a selectable power of 2, 10 or 40 watts. Third, the power output stability and thermal characteristics of the final device are greatly increased, even in key-down sessions that last 45 minutes.

Although this article has discussed only the Standard line, examination of any 2-meter rig with a low-power switch will produce the same findings. Again, don't shoot for more than 3 watts at the low power setting. It can be done, but not without pulling heavier wires. And watch out for those air-wound coils.

John Pakusich, WB6KVF

open filament pins on power tubes

During a recent operating session, it suddenly appeared that my 3-500Z had died, with an apparent open filament. I replaced it with an elderly 4-400A, kept for emergencies, and let a few choice words out.

Close examination of the 3-500Z was puzzling, since the filament appeared intact. A check with an ohmmeter indicated continuity. Since the 4-400A replacement was working fine, the filament transformer was above suspicion.

A close look at the filament pins of the 3-500Z, however, lead to the solution. One of the two pins appeared to

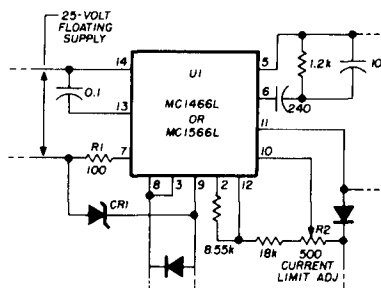
have lost its solder. The other seemed okay. The bad pin was heated with a 40-watt iron and removed from the heavy filament lead. The filament lead was cleaned and scraped and then the pin was fitted back over it and resoldered. The tube has been successfully restored to full service.

The solder connection was probably not too good to start with, and the heavy starting surge current undoubtedly caused the connection to deteriorate further. Since discovering this malady I have been advised of the same problem with a 4-1000A, and I later saved another 3-500Z from becoming a lamp base. If that expensive tube in your final amplifier quits, be sure to check the solder connections to the filament pins before consigning it to the trash can.

Bob Locher, W9KNI

short circuit

In the circuit for the regulated solid-state high-voltage power supply described in the January, 1975, issue (fig. 1, page 42), the 18k resistor connected between pin 2 and the 8.55k resistor should be connected to pin 12; the 8.55k resistor should be connected between pins 2 and 12 as shown below. Also, the 0.33



μ F capacitors connected from the secondary of transformer T3 to ground should be rated at 600 volts.



new products

Heath SB-104 series

The new SB-104 series of equipment from Heathkit is one of the most advanced designs in amateur radio today and includes such features as digital readout, all solid-state circuitry, and broadband rf design. Key units of equipment in the new Heath line are the SB-104 five-band transceiver, SB-644 remote vfo, SB-634 station console, SB-614 station monitor and SB-230 conduction-cooled kilowatt linear.

transceiver

The new SB-104 ssb and CW transceiver is completely solid-state from front end to rf output and runs a cool 100 watts output. The four final transistors are completely protected against high vswr and thermal runaway so there is practically no danger of damaging them in normal operation. The Heath Company is so sure of the design, in fact, that they have placed a one-year warranty on the rf output board and final transistors. Since all of the rf circuits of the SB-104 are broadband, you can move instantly from one end of the band to the other, or from band to band; there is no need to adjust any preselector, drive, load or turn controls — just choose

the band you want, dial in your operating frequency, and go on the air.

The digital dial of the SB-104 provides direct 6-digit readout of your operating frequency with 100-Hz resolution on all bands. And, unlike some digital readout systems that actually read only the vfo frequency (and interpret the output frequency), the digital dial in the SB-104 accounts for the vfo, bfo and high-frequency oscillator signals. With this arrangement there is no need for a frequency calibrator.

The large spinner dial provides about 30 kHz per revolution, a rate that seems ideal for all types of operating habits. Of course, the 100-Hz accuracy of the digital dial means that when you want to be on a certain frequency, that's where you'll be. If you only need 1-kHz accuracy, the last, 100-Hz digit can be turned off with a front-panel switch.

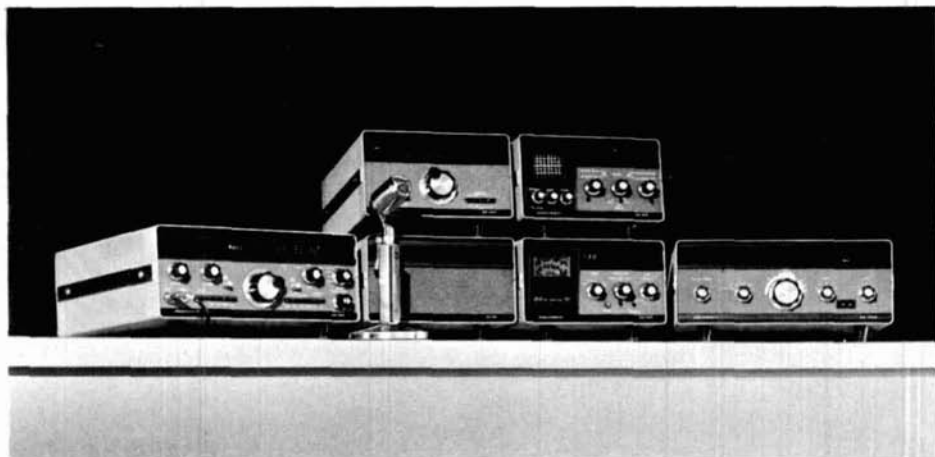
Although the final amplifier of the SB-104 is rated at 100 watts *output* (ssb or CW), for QRP operation the output can be instantly switched to one-watt with a front-panel pushbutton. The carrier and unwanted sideband are suppressed 55 dB, harmonic radiation is down 45 dB, spurious radiation within ± 3 MHz of the carrier is down 50 dB or more, and third-order intermodulation distortion is 30 dB down from two-tone output.

The broadband receiver has been designed for minimum cross-modulation and intermodulation distortion products — important aspects on today's crowded amateur bands. Adjacent signal overload is negligible with the SB-104 yet sensitivity on all bands is 1 μ V or better for 10 dB signal-plus-noise-to-noise ratio. Selectivity is 2.1 kHz minimum at 6 dB down and 5 kHz maximum at 60 dB down (2:1 shape factor). With the accessory 400-Hz CW filter, selectivity is 400 Hz at 6 dB down and 2 kHz maximum, at 60 dB down. Audio output is rated at 2.5 watts into 4 ohms at less than 10% THD (1 μ V input signal provides 0.5 watt audio output). Agc is switch select-

able for release times of 100 μ s or 1 millisecond; attack time is less than 1 millisecond.

Front panel controls include agc (fast, slow, off) audio and rf gain, main tuning, mic/CW level, vox gain, vox delay and bandswitch. Pushbutton switches are used to control vox, noise blanking, mode, tune, high/low power,

designed for use with a 13.8-volt dc power supply so it is a natural for mobile operation. Current drain is 2 amps on receive and 20 amps on transmit (3 amps when switched to low power). The HP-1144 fixed-station power supply (\$89.95) provides the necessary voltages from 120/240 Vac lines. Other accessories for the SB-104 include the



and metering. On the rear panel are controls for vox anti-trip and sidetone level. Rear panel connections are provided for phone patch, auxiliary audio output, speaker, key, alc, external vfo, i-f output and separate receive and transmit antennas.

The more than 2800 parts used in the SB-104 transceiver are mounted on 15 separate printed-circuit boards for easy assembly and test. Eleven of the boards plug-in and seven of these may be extended out of the chassis for adjustment or troubleshooting while the transceiver is operating. Two large wiring harnesses eliminate 95% of the point-to-point wiring in the SB-104. Total construction time is about 50 hours. Alignment of the completed transceiver is fast and simple, requiring only a dummy load, microphone and vtvm.

The SB-104, priced at \$699.95, is

SB-604 station speaker (\$29.95) SBA-104-1 noise blanker (\$24.95), SBA-104-2 mobile mount (\$34.95), and SBA104-3 400-Hz CW filter (\$34.95).

remote vfo

The SB-644 remote vfo (\$119.95) was designed specifically for the SB-104 transceiver and provides serious DXers with complete split transmit/receive capability. The transceiver vfo can be at one end of the band while the remote vfo is at the other. Furthermore, the system is designed for transceive operation on the remote vfo *or* internal transceiver vfo, transmit on the SB-104 and receive on the remote, or receive on the SB-104 vfo and transmit on the remote. There are also provisions for two crystals in the SB-644 for fixed-frequency control.

Although the linear dial on the front panel of the SB-644 remote vfo places



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you in the right ballpark, actual frequency readout is provided on the transceiver's digital display. The display automatically changes to the correct frequency as you switch from receive to transmit so there's never any doubt as to the frequency you are operating on.

Four front-panel pushbutton switches on the SB-644 control all transceive, transmit and receive modes on both the transceiver and the remote vfo. No switching is necessary at the transceiver — just push a button to turn on the vfo, and push another to select the vfo or crystal. Status lamps behind the window indicate whether the frequency is controlled by the transceiver vfo, remote vfo or crystal.

The vfo circuitry in the SB-644 is the same as that used in the new SB-104, and thanks to the true digital readout used in the transceiver, concern about vfo dial linearity is a thing of the past. The vfo kit is assembled on two circuit boards and only two simple adjustments are required for alignment. Frequency drift is less than 100 Hz per hour after a 30-minute warmup; dial backlash is 100 Hz maximum.

station console

The new SB-634 station console (\$179.95) is actually five station accessories in one: a 24-hour digital clock, ten-minute ID timer; phone patch, rf wattmeter and swr bridge. The digital clock indicates hours, minutes and seconds with digits large enough that they can be read from across the room. The clock runs continuously as long as the console is plugged in, completely independent of the other functions.

The ten-minute ID timer uses three digits to indicate minutes and seconds up to 9:59. At the ten-minute mark the timer automatically recycles and provides a visual alarm or both visual and audible alarms (selectable from the front panel). If you elect to identify before the 10-minute mark, as you

should, the ID timer can be reset by a pushbutton on the front panel. Accuracy of both the 24-clock and ten-minute timer is determined by the accuracy of the power-line frequency.

The hybrid phone patch which is built into the station console allows either manual operation or voice control without switching connections. Isolation between transmit and receive circuits is at least 30 dB and can be adjusted with a rear-panel control. When you are running a phone patch, the meter can be used to indicate VU, and transmitter and receiver gain can be set with separate front-panel controls.

The rf power meter covers the frequency range from 1.8 to 30 MHz in two ranges, 200 or 2000 watts full scale, with accuracy of $\pm 10\%$. If you wish to measure swr, simply press a button on the front panel. Swr sensitivity, which is less than 10 watts, is adjustable.

station monitor

The Heath SB-614 station monitor (\$139.95) will display transmitted ssb, CW, a-m (trapezoid) and RTTY (cross) signals up to one kilowatt from 80 through 6 meters. The flat-face CRT uses push-pull drive for a keystone-free, sharp, clean trace that can be used to diagnose a wide variety of operating problems: non-linearity, insufficient or excess drive, poor carrier or sideband suppression, regeneration, parasitics and key clicks. The operating manual includes 40 CRT display illustrations and explanations.

The station monitor has all standard scope control functions with a recurrent, automatic sync-type sweep generator that is adjustable in three ranges from 10 Hz to 10 kHz. For limited test applications the SB-614 can also be used as a normal scope with 10 kHz to 50 kHz bandwidth, good sync and high input sensitivity (60 mV rms for $\frac{1}{4}$ -inch vertical deflection). A rear panel 10:1 attenuator provides extra operating convenience.

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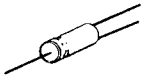
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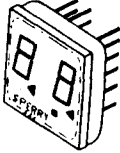
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conduction-cooled linear

The new SB-230 1-kilowatt conduction-cooled linear uses a rugged Eimac 8873 triode in a proven grounded-grid circuit to deliver up to 1200 watts PEP input on single sideband and 1000 watts on CW with less than 100 watts drive. The linear is also rated at 400 watts input for slow-scan television or RTTY. Third-order intermodulation distortion is -30 dB or more. And since the SB-230 is conduction cooled, there is no need for a noisy blower — the massive heatsink on the rear of the unit takes care of all the cooling requirements.

The cabinet of the SB-230 features microswitch interlocks on both the top and bottom to shut down the primary power when either of the covers is removed. The temperature of the heat-sink is monitored so if the temperature rises too high a thermal circuit breaker opens and the amplifier shuts down. To allow the 8873 sufficient time to warm-up when it is first turned on, a delay circuit is built in. When warmup is completed the Delay light goes out. Lights are also provided to indicate when the exciter is running straight through and when the amplifier has been shut down because of high heat-sink temperature.

Bandswitching of the linear is accomplished with a single knob and the load and tune controls are clearly marked so you can return to a favorite operating frequency by simply noting the control positions. The back-lighted meter is used to indicate relative rf power, plate current, grid current or high voltage. Relative power sensitivity is adjustable from the front panel. Price, including power supply, is \$319.95.

For more information on the exciting new Heath SB-line, the first complete high-frequency amateur communications system to be offered in some time, write to the Heath Company, Benton Harbor, Michigan 49022, or use check-off on page 110.



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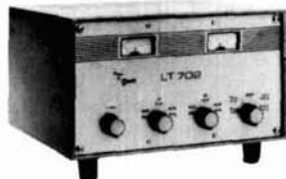
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An all mode transverter (SSB, AM, FM, CW, RTTY, SSTV) for transposing a 2 m signal to the 70 cm band and vice versa. The full 10 MHz between 430 and 440 MHz is covered in five bands of 2 MHz each. Each of these bands can be selected individually for transmit and receive so that it is especially suitable for operation over repeaters and transponders. The receive converter is synchronized to the transmit oscillator during transceive operation. Several coaxial relays are provided for dual band operation 2 m/70 cm and 70 cm/2 m. Output

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we've offered. Over 2800 parts, yet all but a handful mount on one of the 15 glass epoxy boards, and two wiring harnesses eliminate most of the point-to-point wiring. Eleven boards plug-in and 7 can be extended out of the chassis while operational; the other boards are accessible without dismantling. Alignment is fast and simple due largely to the broadband design; only a dummy load, mike, and VTVM are needed.

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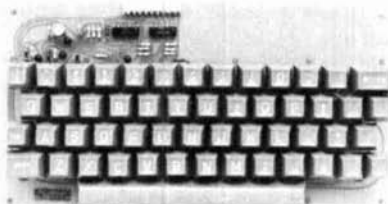
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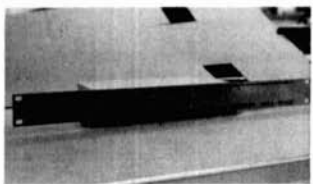
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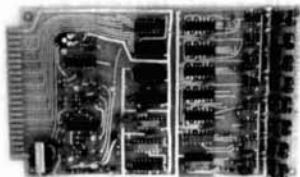
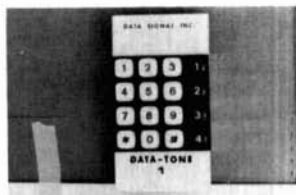
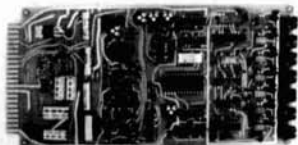
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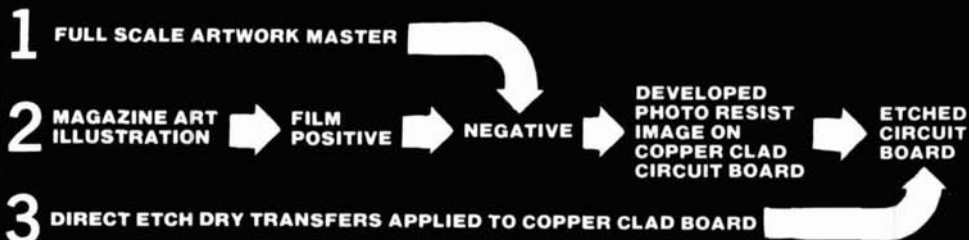
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ROUNDING TO TEN DIGITS	YES	YES	NO
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DEGREE/RADIAN KEY	YES	YES	NO
MEMORY (OTHER THAN STACK)	1	1	1
KEYS	26	40	35
LOGIC	ALGEBRAIC	ALGEBRAIC	POLISH
LOG, ln	YES	YES	YES
TRIG (ARC, SIN, COS, TAN)	YES	YES	YES
DEGREE \leftrightarrow RADIAN CONVERSION	YES	YES	NO
DEG/RAD MODE SELECTION	YES	YES	NO
y^x , a^x	YES	YES	YES
x^y	NO	YES	NO
\sqrt{x}	YES	YES	YES
$\sqrt[y]{x}$ $x \leftrightarrow y$	YES	YES	NO
$\frac{1}{x}$	YES	YES	YES
EXCHANGE X WITH Y	YES	YES	YES
BIGGEST DISPLAY	YES	NO APPROX. 20% SMALLER	NO APPROX. 25% SMALLER
2 PARENTHESIS LEVELS (BRACKETING)	YES	NO	NO
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SCIENTIFIC NOTATION	YES	YES	YES

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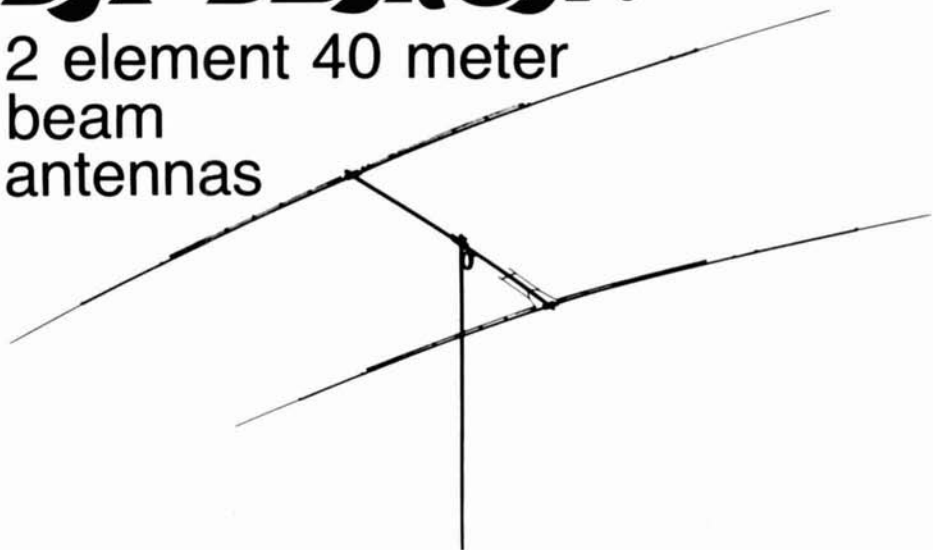
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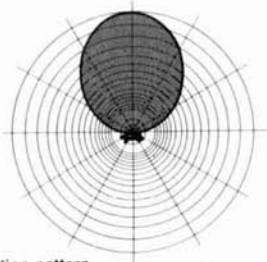


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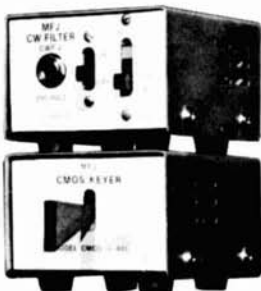
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from Last Listing	93.00	93.00	
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REDUCED from Last Listing	145.00	145.00	
Gertsch FM-3 Frequency Meter	207.00	176.00	
Heath HW-202 w/Six Sets of Crystals	179.00	179.00	
Icom IC-2F Transceiver with 6 Sets of Crystals and IC-3P AC Supply — REDUCED	136.00	136.00	
Ken KP-202 with Nicads, Case, Flexible Antenna, Home-Brew Charger, 6 Sets	275.00	275.00	
2 1/2" Rack Cabinet w/Carrying Handles	17.00	17.00	
Realistic PR0-4 Pocket Scanner with Four Crystals and Flexible Antenna	95.00	81.00	
Regency HR-6 w/Four Sets of Crystals	245.00	208.00	
Regency HR-2A with BARRA Preamp and Three Sets of Crystals	150.00	128.00	
Regency HR-212 w/5 1/2 Sets of Crystals	189.00	161.00	
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This is just a partial listing. Send SASE for our full used equipment list. Call or write for a quote on your favorite gear.

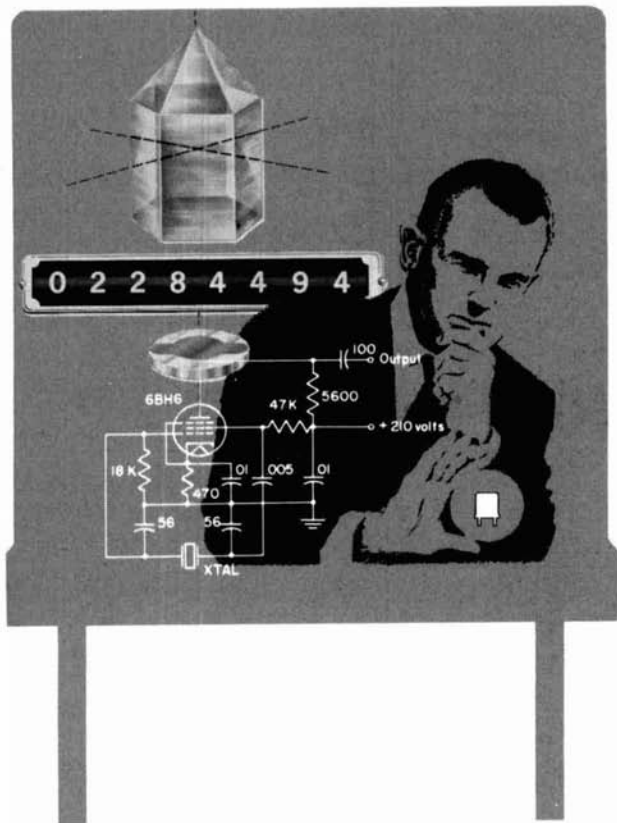
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MODEL 4 X 6-C WIRED and TESTED \$270.00 f.o.b.

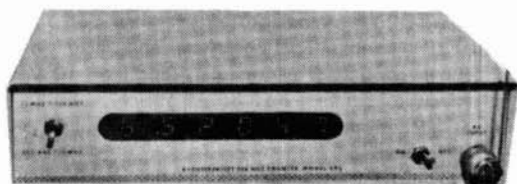
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SPECIFICATIONS

Frequency Range	500 kHz - 250 MHz
Sensitivity	Less than 80 mV at 150 MHz
Input Z	50 ohms
Max. Input Voltage	15 V rms, 50 V dc
Time Base	Crystal Clock plus-minus 10 ppm 0 C to 40 C ambient
Readout	6 Digit 7 Segment LED
Power	120 V ac
Dimensions	2 1/4" H, 10" L, 7" D
Cabinet	Light blue

K - ENTERPRISES

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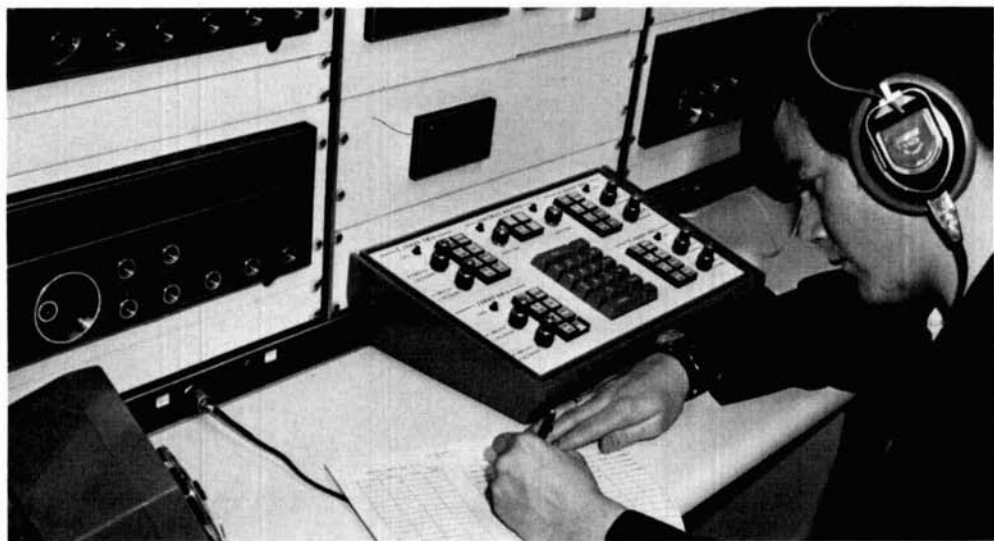
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	MMc 50	MMc 144	MMc 220	MMc 432	MMc 1296
RF Freq. (MHz) †	50-54	144-148	220-224	432-436	1296-1300
IF Freq. †	28-32	28-32	28-32	28-32	28-32
N.F. (typical)	2.5dB	2.8dB	3.4dB	3.8dB	9.0dB
Nom. Gain	30dB	30dB	26dB	28dB	20dB
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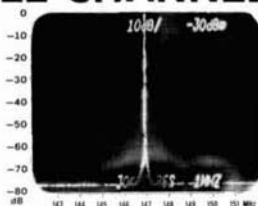
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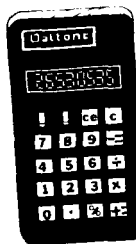
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□ 2102 1000-bit "static" Ram for above. \$6.95



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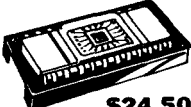


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1.75 Each

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*35 LED matrix

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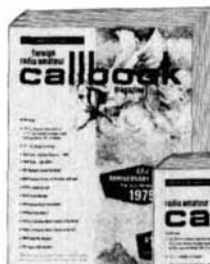
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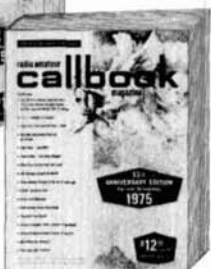
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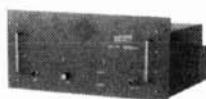
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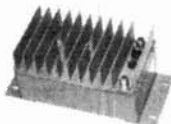
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TRADE HANDY-TALKIE Regency HRT-2, new, never opened, for solid state gear or \$169. WØKGI, 303-447-8692 (Colorado).

EXCLUSIVELY HAM TELETYPE 21st year, RTTY Journal, articles, news, DX, VHF, classified ads. Sample 30¢. \$3.00 per year. Box 837, Royal Oak, Michigan 48068.

HALLICRAFTERS FPM-300 with 10 meter xtals and fan, \$340. Great condition. Joe Roth, WN3VJK, 99 Thompson Dr., Pittsburgh, PA 15229.

PC's, Send large S.A.S.E. for list. Semtronics, Rt. #3, Box 1, Bellaire, Ohio 43906.

THE SIXTEENTH ANNUAL HAMFEST, sponsored by the Southern Tier Amateur Radio Clubs, is scheduled for 10 a.m., April 19, 1975, at St. John's Ukrainian Hall, Johnson City, New York. Admission to lectures and flea market only, \$1.00 for adults. Total admission including awards and excellent dinner, \$6.50. For tickets or further information, write to: STARC, P. O. Box 11, Endicott, N. Y. 13760. Advance ticket sales only by April 16, 1975.

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THE GREATER BALTIMORE HAMBOREE, Calvert Hall College, Putty Hill and Goucher Boulevard, Towson, Maryland (one mile south of Exit 28 of Beltway-Interstate 695), Sunday, April 6, 1975, 9 a.m. Food service, flea market, contests, prizes. Registration: \$2.00. Complete table set-ups indoors. INFO: Joe Lochte, 5400 Roland Avenue, Baltimore, Maryland 21210 or Brother Gerald Malseed, 8102 La Salle Road, Towson, Maryland 21204. (301-825-4266).

WORLD QSL — See ad this issue.

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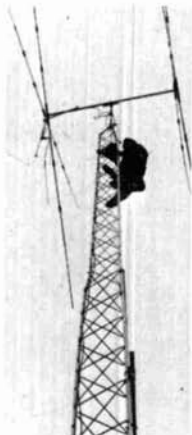
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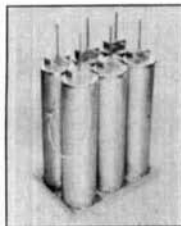
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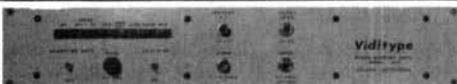
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WANTED: Someone with a 1919 or 1920 Canadian license list or some other means of proving that John R. Nesbit was licensed as 5BM at that time. Please contact VE7AS.

EVANSVILLE TRI-STATE ARS will hold their annual hamfest on May 18, 1975 at the 4-H fairgrounds, US41, 3 miles north of town. Overnight camping, auction, flea market, door prizes, and ladies bingo. For information contact Jay, WB9ICL, R#1, Box 56M, Wadesville, In. 47638.

POLICE CALL. A complete listing including frequency and location of VHF-UHF public service stations. Nine volumes cover the whole USA. Order the volume for your state. Only \$4.95 plus 25¢ shipping. HAM RADIO, Greenville, N.H. 03048.

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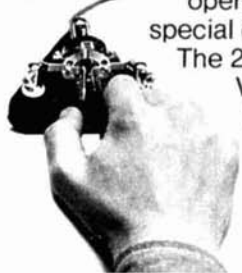
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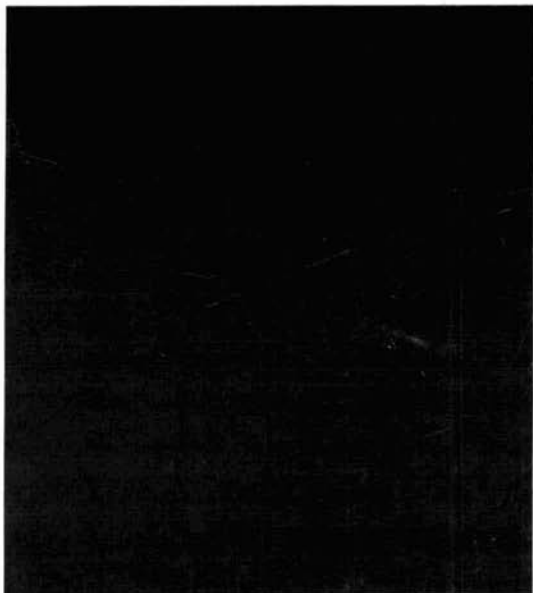
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