OCTOBER 1978 / $2.00

ham radio magazine

SPECIAL
4th Annual Receiver Issue

More Details? CHECK—OFF Page 142

ham radio
magazine

OCTOBER 1978

volume 11, number 10

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates

United States: one year, $12.00
two years, $22.00; three years, $30.00

Canada: one year, $14.00
two years, $26.00; three years, $36.00

Europe, Japan, Africa:
(via Air Forwarding Service)
one year, $25.00

North America, South America, Australia
and Asia (except Japan):
(via Surface Mail)
one year, $18.00

foreign subscription agents

Foreign subscription agents are
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are available from
University Microfilms, International
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Order publication number 3076

Cassette tapes of selected articles
from *ham radio* are available to the
blind and physically handicapped
from Recorded Periodicals
919 Walnut Street, 8th Floor
Philadelphia, Pennsylvania 19107

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 233340

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a second look

by Jim Fisk

Our society is a very mobile one, and with the recently lowered air fares, more people than ever before are traveling by air. It's only natural for the vhf-fm operator, with his portable fm rig, to question the possibility of using his equipment on commercial flights.

It is popularly believed that all you have to do is obtain the captain's permission to operate; surely your little two-watt fm rig is not going to cause any interference with the high-powered radio equipment used on board the aircraft. However, this is not the case — according to the Federal Air Regulations, approval must be obtained from the air carrier (airline) and not the pilot in command. However, once approved by the air carrier, the permission of the captain in command must *also* be obtained to operate equipment aboard a particular flight.

Shortly after World War II, portable Japanese fm broadcast receivers appeared on the market, and passengers started using them aboard commercial flights. At the same time, aircraft navigation radios started doing strange things, and it didn't take long to determine that the interference was being caused by rf radiation from the portable fm receivers. The aircraft radios literally went wild, and at least two aircraft accidents were attributed to interference of this type.

When it was determined that this interference was present, the FAA promulgated new regulations, paragraph 91.19 of the Federal Air Regulations. This paragraph states that no electronic device may be operated aboard a commercial airliner *except* heart pacemakers, voice recorders, hearing aids, calculators, electric shavers and electric watches, unless the device has been approved by the air carrier or operator. The regulation further states that *the captain of the aircraft does not have the authority to authorize such operation*.

Consider, for a moment, what might happen if such operation were allowed. Suppose you have been operating all across the country, and your plane is about to land. A passenger with a briefcase telephone sitting across from you has been watching you operate. About 10 minutes before landing, he decides to call his wife. Unfortunately, his radio telephone transmits right in the middle of the glide slope spectrum. As soon as his transmiter is keyed, the glide slope indicator cross pointer goes up or down, and the autopilot follows it. That could be disastrous.

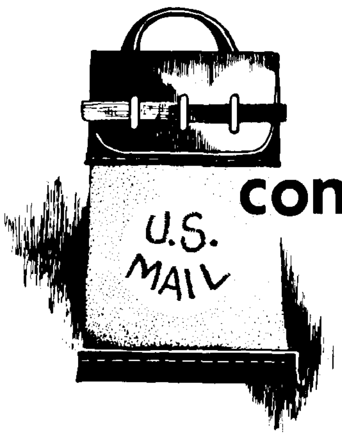
As an airliner flies across the country, the pilot changes frequency every 5 minutes or so. If several fm operators are on the same flight, only one can talk at a time, so some may decide to switch to other frequencies. When you figure out all of the i-f and carrier frequencies of the aircraft radios, plus the amateur gear, plus all the possible mixing products, you can appreciate the magnitude of the problem.

A few years ago a well known vhf-fm operator prevailed upon an airline to test his Motorola HT in one of their aircraft so he could operate during a flight he planned to take. After months of correspondence and personal meetings with airline communications people (many of whom were amateurs), the airline agreed to run the necessary tests. On the appointed day the aircraft was removed from line operation and the test began; it took three hours and four men to complete. The HT caused no interference, and the amateur received a letter authorizing the operation of *that* HT on *that* particular trip in only *that* type of aircraft. It's easy to understand why the airlines, who are trying to cut costs, are not enthusiastic about testing an individual's vhf-fm equipment.

Many fm operators continue to ask the captain's permission to operate, and he may give it, not realizing the possible bind he is putting himself in; he could have his license suspended or he could be fired. Don't put him in that position, and don't subject yourself and other passengers to a situation which could be hazardous to all on board.

Remember, you may not cause any interference during the trip, but the ILS glide slope receiver is used only during the last few minutes of flight, so interference may not be noticed until it's too late!

Jim Fisk, W1HR
editor-in-chief



comments

zip-cord feedlines

Dear HR:

I feel that the article by W7RXV (*ham radio*, April, 1978, page 32) deserves a rebuttal. First, I doubt the accuracy of Fruitman's measurements on the coax. As a practicing antenna engineer, I have made rather precise measurements of the loss of both RG-58 and RG-8. The numbers that have repeated themselves time and time again are very close to 2.0 dB/30 meters for RG-58 at 20 MHz and 1.0 dB at 4 MHz. Thus, for a 20-meter run, Fruitman should have measured 20 per cent (versus his claimed 58 per cent) transmission loss at 20 MHz. The other numbers are off by a proportionate percentage.

More serious, however, is Fruitman's apparent disregard of measurement and use techniques between coaxial cable and balanced-wire line. When he says "the zip cord came down and the RG-58 went up" one must be a bit suspicious that the two different types of transmission cable were fed from the same connector. Moreover, one must then ask if he used the same measurement technique for both coax and twinlead without the use of baluns. Had he made the blunder of not using the twinlead properly, the errors he reports would be expected. The loss is not in the twinlead, but in the VSWR mismatch.

I have used lamp (zip) cord on the lower bands with reasonable results. Of course, you can't go out and buy just any old zip cord and hook it up without going through a bit of investigation. The impedance of zip cord varies from about 70 ohms (for the light duty wire) to about 30 ohms for the super-heavy duty cord. Loss is a bit better than the normal 75-ohm twinlead.

Jim Weir, WB6BHI
VP Engineering
Radio Systems Technology
Grass Valley, California

51J product detector

Dear HR:

I would like to thank Bill Orr, W6SAI, for the fine article in the February, 1978, issue of *ham radio* which showed how to add a product detector to a Collins 51J receiver. I made this modification to my receiver and am very pleased with the change.

However, I did run into a problem which evidently did not crop up when Mr. Orr made his conversions. I found that when I turned up my audio gain, the receiver immediately started motor-boating. Investigation revealed it was not motor-boating, but was hash pick up from the filter unit of the power supply. I tried replacing the plug-in type filter capacitor, but this did not cure the problem. I finally had to add a 10-μF filter capacitor at the junction of the screen and plate resistors.

Frisco Roberts, K5CE
Corpus Christi, Texas

pi networks

Dear HR:

The article by Irv Hoff, W6FFC, in the June, 1978, issue of *ham radio* on pi network design was well done and fulfilled a need for those interested in building equipment. However I'd like to point out one minor discrepancy in an otherwise excellent article that may cause some confusion. On page 63, Irv points out that an rf choke should be used at the antenna output of any pi or pi-L network. All well and good, but nowhere in the article does the author indicate *how* this rf choke should be connected. Perhaps it is implicit in the text, but I think that a simple addition to **fig. 3** showing how to connect the choke would be appropriate. The rf choke should be strapped between the antenna output connector and *ground*. The choke is essential in such a circuit in series with the network output!

A. Wilson, W6NIF
Encinitas, California

phased antenna

Dear HR:

After looking at some of the beams (especially 40 meters) I came up with a cheaper way to get a signal to the West Coast, especially at night. The idea was copied from some of the other antennas — it's very easy to make. It uses half-wave dipoles in phase, and gives a nice bi-directional antenna. Taking in the velocity factor of RG-8, which is 0.66, I took 492×0.66 and divided that by the frequency $7.250\ MHz = 44'\,9''$ for my phasing lines, one for each dipole.

I use a minimum of 1/8-wavelength spacing (16′1″). I found very little difference between 1/4-wave and 1/8-wave spacing. I've checked it with several stations and, with a friend to compare signal reports with, I ran 500 W PEP while he ran the full 2 kw PEP. I averaged 2 to 3 S-units more than he did all the time. This same thing was tried on 75, and I got the same results. The stations on the West Coast gave me the best signal report. I hope this works as well for others as it has for me.

Jerry Thacker
Francisco, Indiana

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BALLOON MOBILE AMATEUR RADIO provided the Double Eagle II with vital communications as the 112 foot high balloon became the first free-air craft to cross the Atlantic. The balloonists, operating as W5OCP on 14325 kHz, maintained contact with their ground crew in Bedford, Massachusetts, using at Atlas transceiver.

Though None Of The Three balloonists were themselves Amateurs, there were Amateurs (including W5OCP) in the Eagle II's ground support crew. The Atlas had been taken along for compact, lightweight backup communications, and when the crew found themselves without other communications halfway through their trip, their emergency use of Amateur frequencies under paragraph 1381 of the International Radio Regulations was not inappropriate.

Another Adventurer, Naomi Uemura, JG1QFW, was in Washington in late August for a press conference and a celebration of his accomplishment: reaching the North Pole solo. With much of the trip's communications burden carried by Amateur Radio, lots of good PR should result.

THE FCC'S BAN ON 10-METER LINEARS came an important step closer to being challenged when the ARRL filed, in late August, a Petition for Review of the controversial decision in the United States Court of Appeals, District of Columbia Circuit. This in effect gives the League 45 days in which to prepare a brief on the matter, and, with the League Executive Committee meeting at mid month, the final decision on just how far they'll carry the matter should be made then.

Amateur Use Of ASCII for RTTY is the sole subject of the Notice of Inquiry and further Notice of Proposed Rule Making on the "Bandwidth Docket" — 20777. The NPRM will propose adding ASCII to the permitted emissions for Amateurs, while the NOI will ask what specific standards should go into the new rules. Comment Due Date on this Notice of Inquiry and NPRM is November 15, with Reply Comments due December 16.

The FCC-Amateur Media Meeting proposed for this month will probably be devoted to Amateur exams — content, study guides, exam administration both in and out of the FCC, and the like. Dates are available at Gettysburg, but it could be further delayed if a 10th Notice of Inquiry on WARC 79 requires a meeting of the Advisory Committee on Amateur Radio, which could be held at the same time.

A SIGNIFICANT ANTENNA VICTORY has been achieved by W6QOL. The pro-Amateur Radio decision in this case, rendered by the Federal District Court, prohibits the city of Placentia, California, from limiting W6QOL's antenna to 25 feet.

By Its Decision, that court directly contradicted the California State Court's decision that Cerritos, California, had the right (in the N6QQ case) to limit an Amateur's antenna height. Thus it appears that the Supreme Court, which declined to rule last spring on the N6QQ decision, will again become involved in the battle, as there are now contradictory decisions that must be resolved in the same area of law.

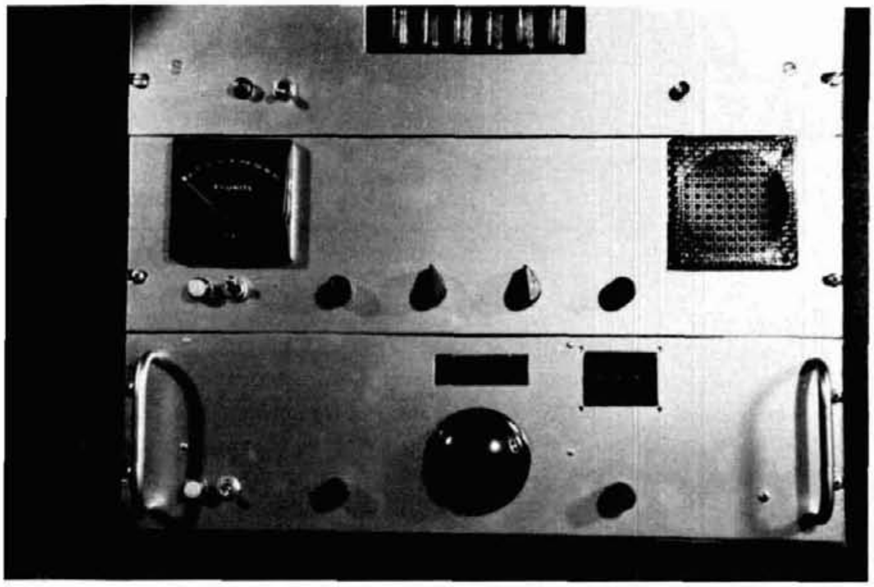
Another Antenna Battle appears to have been won in Farmington, Michigan, a Detroit suburb. Despite strong objections from home owners, the Farmington Heights Planning Commission has come out in favor of tripling the permitted height of radio towers from the present 25 to 75 feet. In a new ordinance sent to the city council, the planning commission proposed increasing the maximum to 75 feet, but with the proviso that the tower height could be no more than half the width of the lot.

1979 ARRL DX CONTEST, having been reduced by the Board of Directors to one weekend per mode, will be held in March with the first weekend for phone and the third for CW. Preference for February dates from large numbers of U.S. stations has been strong, citing building low-band noise from spring storms as a major problem. But, an informal poll of overseas participants indicated a preference for March dates.

DURING TORRENTIAL RAINS that brought death and destruction to south and central parts of Texas, Amateurs served a vital communications role. Nearby Amateurs received first word of the disaster at 6:00 AM on August 2, when K5RZD called into San Antonio via repeater to ask for helicopter evacuation of flood survivors from Medina. At about the same time, the U.S. Whether Service was calling EC WA5RNV asking for Amateur communications help. Until the waters started to recede late in the weekend, an estimated 100 Amateurs worked around the clock providing much needed communications for rescue workers and survivors. In addition to providing disaster communications, a number of Amateurs monitored flood gauges throughout the affected countries to provide Weather Service Chief Hydrologist George Kush with vital data for predicting which areas were threatened and needed to be evacuated.

THE FIFTH 2-METER "WAS" has been earned by K1WHS thanks to K9SS and his Idaho operation.

ANYONE DESIRING CATV cable and/or connectors per the Woods article in September, ham radio, should send a self-addressed, stamped envelope to Box 7111, Phoenix, Arizona 85011.



digitally programmable high-frequency communications receiver

High-frequency receiver for 1-30 MHz features up conversion, frequency synthesis, and novel digital control of the rf and mixer circuits

The high-frequency communications receiver described here includes a unique digital interface that provides it with unusual capability. The receiver covers selected portions of the high-frequency spectrum between 1.8 and 30 MHz with the aid of a frequency synthesizer that is an integral part of the digital display. Coverage includes all of the amateur bands, two bands which include WWV, several of the international short-wave broadcast bands, and most of the CB band. Coverage is not limited to these bands. In fact, the basic scheme is such that *the receiver can be set up to cover the entire range from 1.8 to 30 MHz.*

development

This unusual receiver was developed strictly as a hobby over a period of about a year. The object was to design and build a high-performance breadboard model that could be controlled from the front panel through a digital interface. In many respects, the result represents a radical departure from conventional receiver design. Once the basic scheme was conceived, it was only a matter of building up the

By **Norman J. Foot, WA9HUV**, 293 East Madison Avenue, Elmhurst, Illinois 60126

various functional blocks, integrating them, connecting the interface, and checking out the entire system. At first blush this seemed easy, but assuredly it was not.

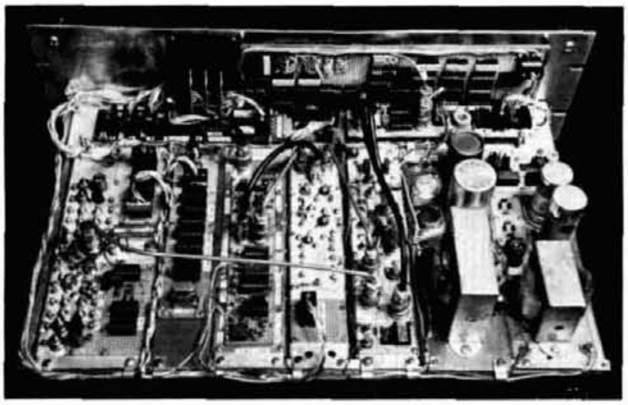
It is difficult to describe all of the details of the programmable receiver in one magazine article. Therefore, this article will include a general description of the overall scheme with the aid of functional block diagrams, so that a clear understanding of the basic idea will be gained. An overall wiring diagram of the frequency counter is not included, but special emphasis will be given to the circuits associated with the frequency synthesizer. Other circuits that will be described are those considered to be unique, such as logic control and the high-frequency oscillator.

general description

To simplify the development process, the decision was made to divide the receiver into physical subassemblies or modules, each including circuitry that logically belongs together. Each individual module was fitted with BNC connectors for rf and i-f interconnections, and 14 pin DIP sockets were used for logic and power. While this modularized approach added considerably to physical size and circuit complexity, it provided a way to easily and quickly remove modules to facilitate debugging, modifications, and repair. Pertinent test points were included on each module. Extender cables were used to operate the receiver with one or two modules removed from the main frame.

The modules are mounted on a pair of 28 × 43 × 5 cm (11 × 17 × 2 inch) aluminum chassis, each equipped with a standard 13.3-cm (5-1/4-inch) high relay rack panel. The lower deck houses the circuits that function to select and display the frequency to which the receiver is tuned, and to convert incoming signals to 32 MHz. The upper deck contains the final mixer and all of the 1650 kHz i-f, mode select, detection, and audio circuitry. These two decks are referred to as the 32-MHz and the 1650-kHz decks.

Construction of the 32-MHz deck which includes the main tuning control, frequency display, thumbwheel switches, and rf gain control.



Rear view of the 32-MHz deck showing the perf-board construction.

Now that the design has been confirmed by the working breadboard model, it should be a relatively simple matter to physically reconfigure the receiver to fit into a single, standard-size enclosure.

No printed-circuit layouts have yet been prepared. All of the digital circuits were assembled by hand using pieces of perforated epoxy fiberglass with a hole matrix on 2.5-mm (0.1-inch) centers, such as Vector 169P84. Component parts were hand wired and soldered using a combination of wire wrap technique and no. 26 (0.4 mm) tinned copper wire insulated with Teflon sleeving.

A major challenge for those wishing to reproduce this receiver is to reduce the counter and other IC boards to printed circuits. The counter board alone includes 27 integrated circuits of which six are LED displays. The total IC count is 71. There are a total of 13 DIP sockets used for power, and three sets of 8 DIP switches. In addition, there are 34 transistors including series pass regulators in the power supplies, and 44 diodes including switching and power supply rectifiers and zener regulators.

The overall gain of the receiver from antenna to product detector output is approximately 122 dB when all controls are set for maximum. Overall sensitivity is 0.2 μV for 10 dB S + N/N ratio, which is more than adequate for most high-frequency requirements.

The front panel of the 32-MHz deck has a three-digit thumbwheel for band and band segment selection, a vfo knob for tuning 100-kHz band segments, and a six-digit seven-segment LED display which indicates the antenna frequency to within ± 100 Hz. The accuracy of the readout is based on a 1.0 MHz crystal clock which is zero beat with WWV. There is also a unique antenna trimmer, an rf gain control, pilot light, and power switch.

The 1650-kHz deck includes a mode switch which

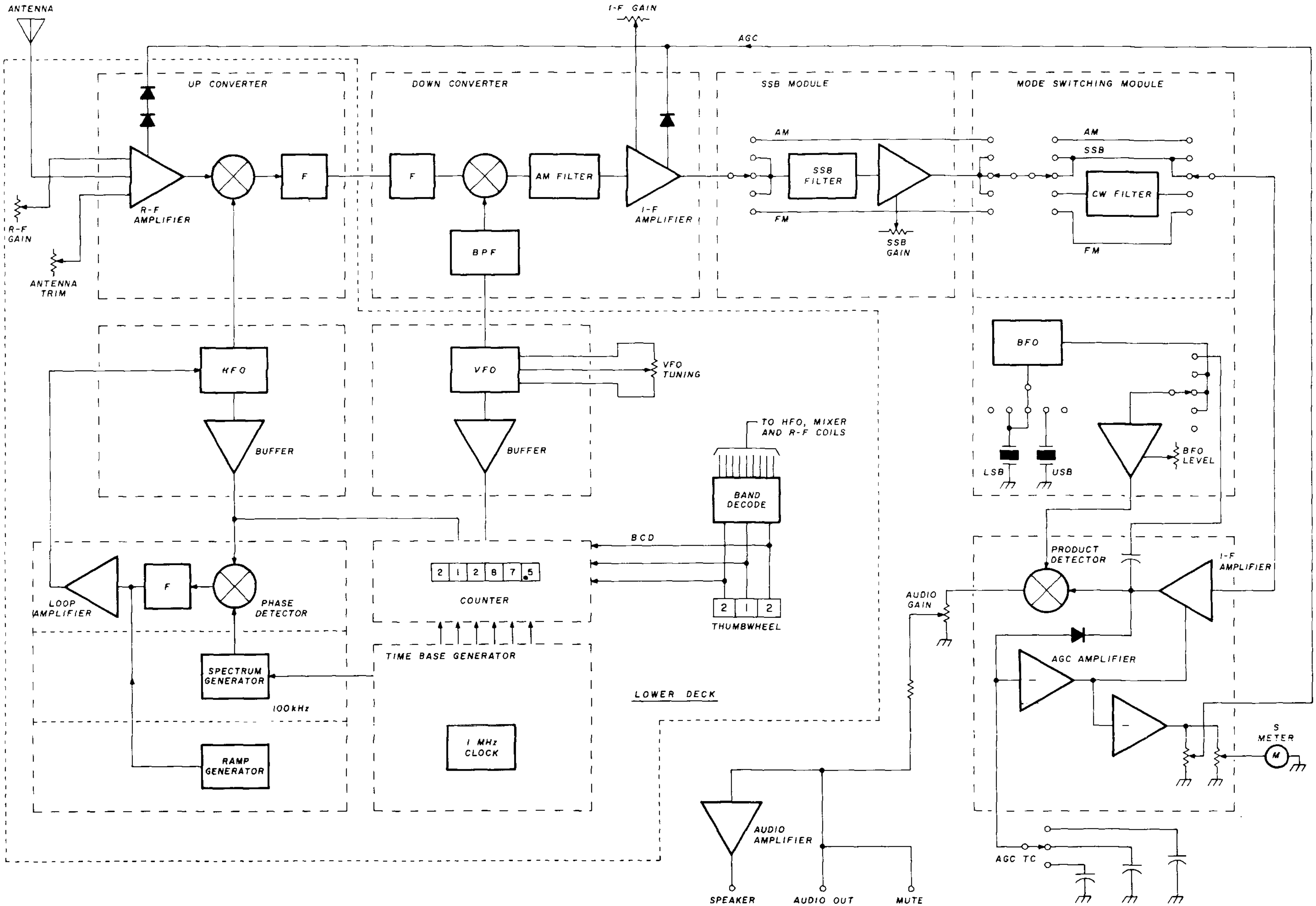


fig. 1. Simplified block diagram of the digitally programmable high-frequency communications receiver, showing the modular construction. The receiver features up conversion to 32 MHz, a frequency synthesizer, and diode-switched front-end and mixer circuits.

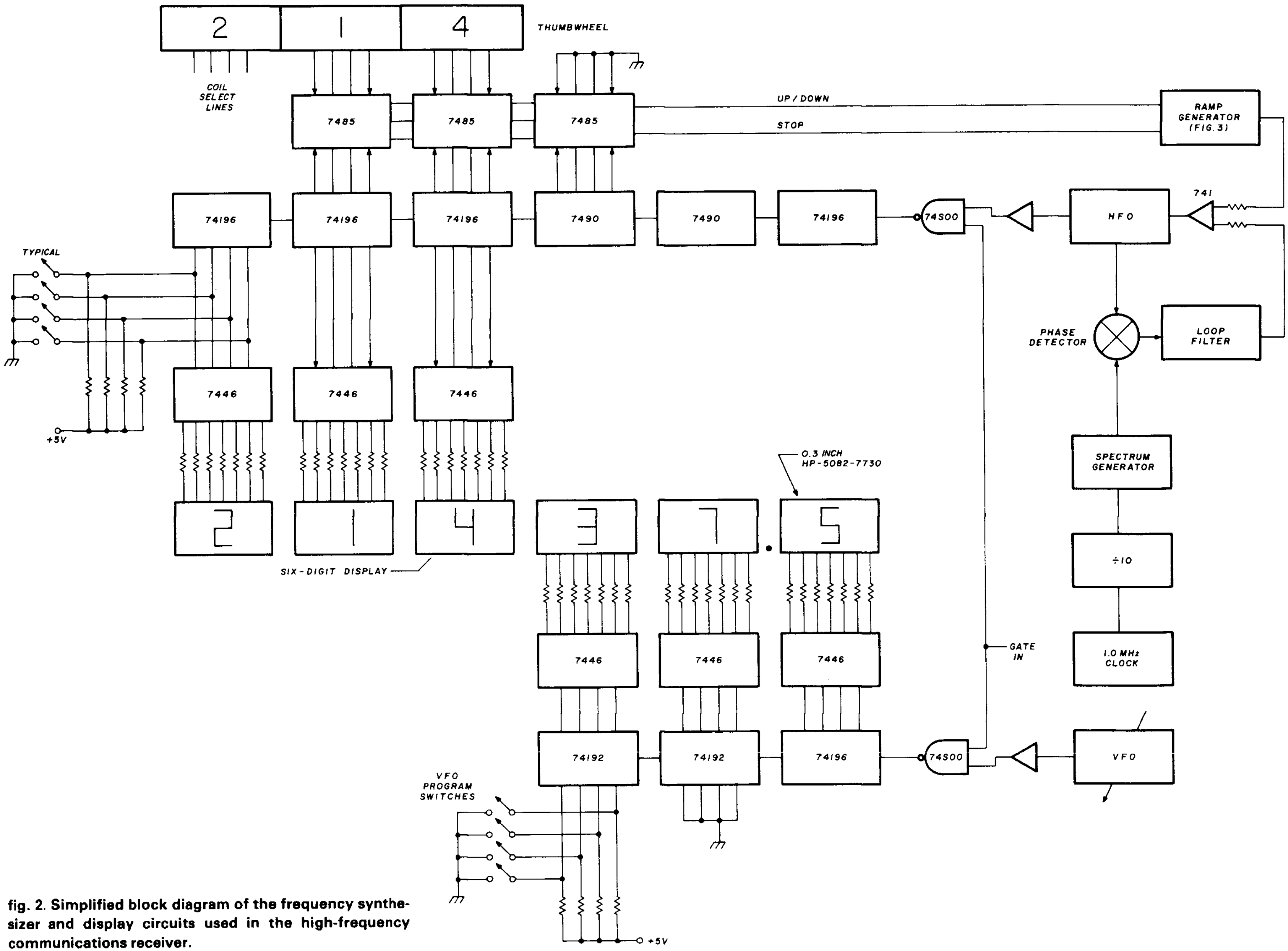


fig. 2. Simplified block diagram of the frequency synthesizer and display circuits used in the high-frequency communications receiver.

allows selection of a-m, USB, LSB, CW, or fm; agc release switch, S-meter, audio gain control, i-f gain control, pilot light, power switch, and a small loud speaker. Each deck includes its own regulated power supply.

Both the band-select thumbwheel switches and the mode switch output TTL compatible binary codes. The bandswitch thumbwheels provide a certain amount of redundancy, since the TENS, UNITS, and TENTHS MHz displays always agree with the decimal numbers programmed into the thumbwheel windows. This verifies that the receiver has indeed responded correctly to the command. If an invalid command is entered, such as 1.5 MHz, for example, the audio output and display are gated off until such time as a valid command is programmed.

basic receiver design

The ground rules for high performance in a general-coverage receiver were elegantly described by Wayne Ryder, W6URH, in his recent *ham radio* article.[1] He discusses the design criteria for general coverage, and gives illustrations of up-conversion schemes designed to minimize interference from the various radio services.

The programmable receiver described here is also designed for general coverage, although the model I built covers selected portions of the high-frequency spectrum. It is a double-conversion system with the first i-f above and the second i-f below the tuning range. The choice of 32 MHz as the first i-f places the first local oscillator (hfo) between 33.8 and 61.9 MHz; therefore, the image band is between 65.8 and

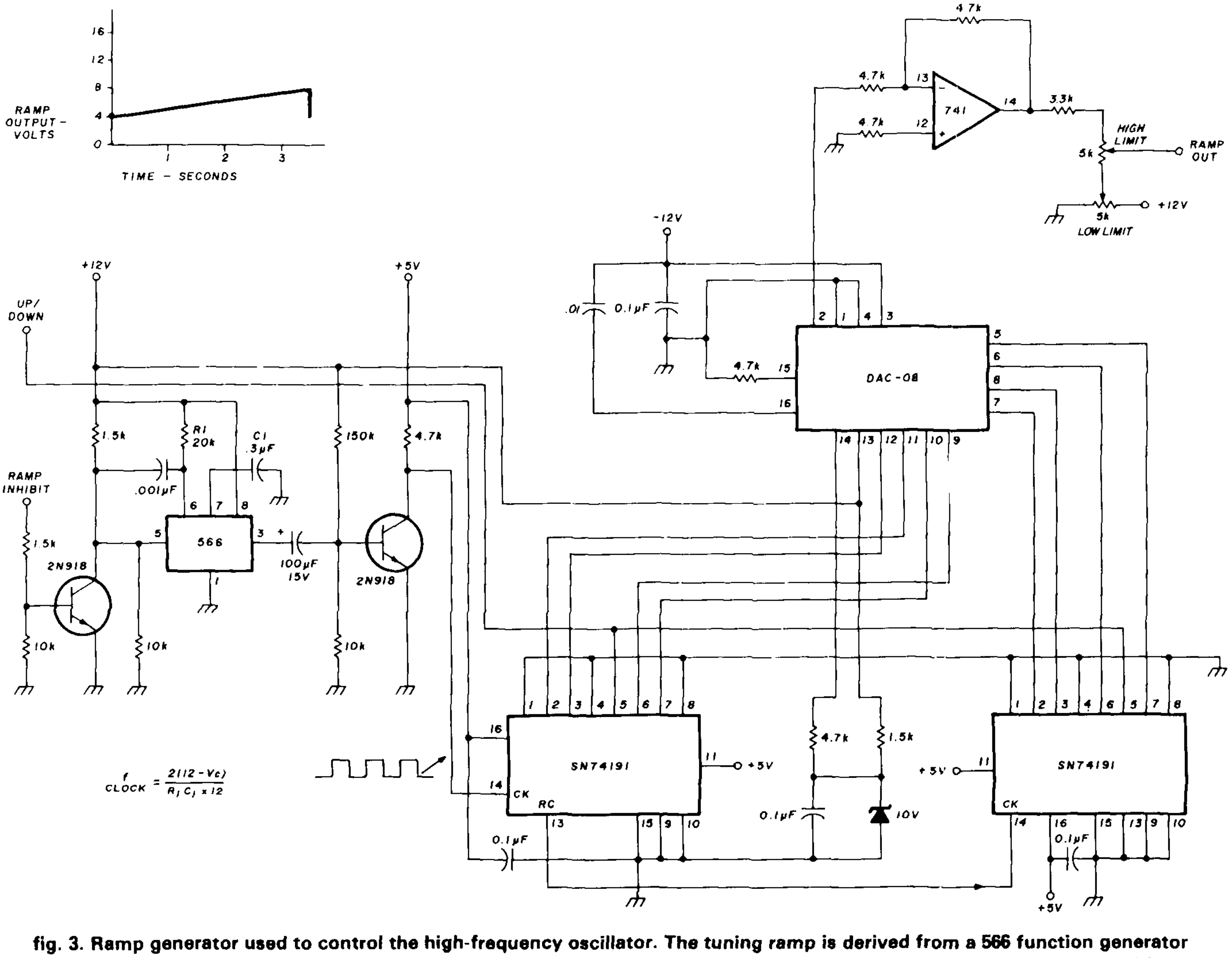


fig. 3. Ramp generator used to control the high-frequency oscillator. The tuning ramp is derived from a 566 function generator and a pair of 74191 up/down counters which drive an 8-bit digital/analog converter. All DIP packages are shown as viewed from the bottom.

High-frequency oscillator module. The band-select signals are connected through the DIP socket.

93.9 MHz. Although this choice may not be optimum in terms of possible interference from local TV or fm stations, the tuned front-end rf circuitry provides sufficient image rejection so that this kind of problem has not been encountered, even though I live near a major metropolitan area.

To adequately describe the programmable nature of this receiver, it is first necessary to explain the block diagram shown in **fig. 1**. The front end includes an rf amplifier and first mixer. The rf input signal is up-converted to the first i-f, which has a passband covering 31.9 to 32 MHz. The hfo signal is derived from the frequency synthesizer and operates from 33.8 to 61.9 MHz. The vfo for the second converter tunes continuously between 30350 and 30250 kHz to down-convert the 32-MHz signals to 1650 kHz.

The digital readout uses six seven-segment LED displays. The three left-hand digits respond to a counter driven by the hfo; the three right-hand digits are driven by the vfo. Programmable counters are used so the display corresponds to the antenna frequency rather than either the vfo or the hfo. The TENS, UNITS, and TENTHS of MHz are selected by the setting of the thumbwheel. The TENS, UNITS, and TENTHS of kHz digits read from 0.00 to 99.9 as the vfo knob is turned clockwise across its range. Thus, the thumbwheel serves as the bandswitch while the vfo knob bandspreads 100-kHz band segments.

The first conversion scheme is described analytically as follows:

$$f_{hfo} = f_s + 32000 \text{ (frequencies in kHz)}$$

where f_s is the signal frequency at the low end of a 100-kHz band segment. For example, if the receiver is tuned to 1800 kHz, the hfo will be phase-locked to 1800 + 32000 = 33800 kHz. The hfo setting for the 28.0 MHz band segment would be 60000 kHz.

The second converter down-converts the 32000 to 31900 kHz i-f band to 1650 kHz. This requires a vfo that tunes from 30350 to 30250 kHz, in that order, so that an up-converted signal that falls within the 32-MHz passband will be down-converted to 1650, depending on the setting of the vfo. Therefore, the overall conversion scheme is:

$$f_s(\text{low end}) = f_{hfo} - 1650 - 30350$$

$$f_s(\text{high end}) = f_{hfo} - 1650 - 30250$$

For example, if the digits set into the thumbwheels

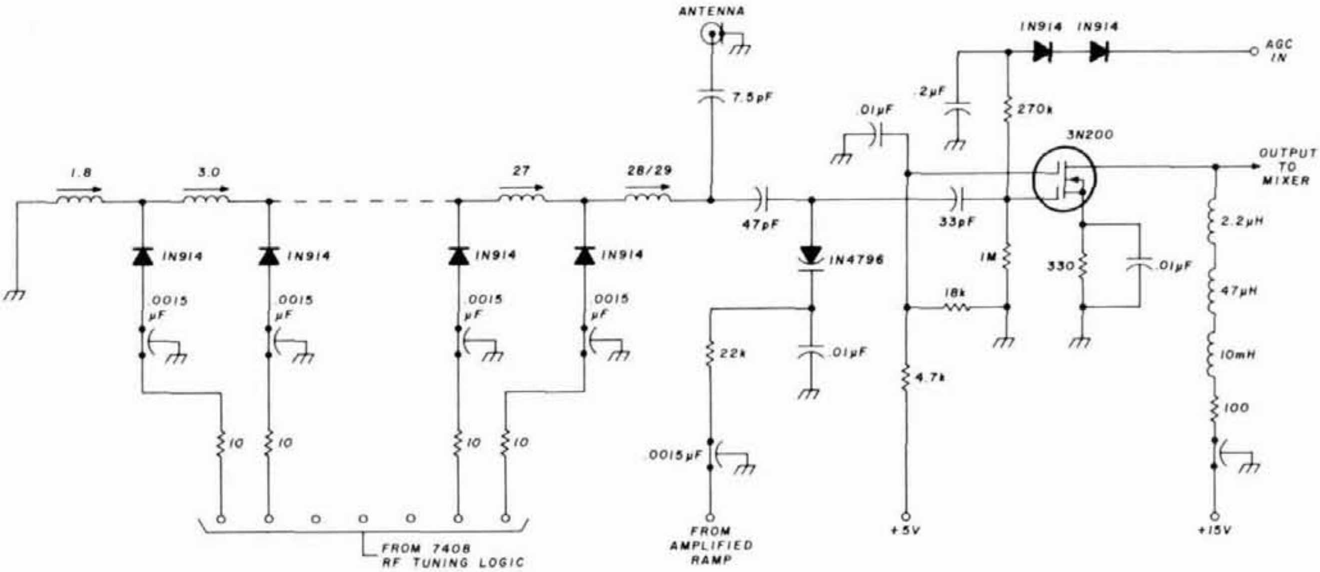


fig. 4. Schematic diagram of the diode-switched rf amplifier. Band selection is accomplished by providing forward current to the 1N914 diodes to short circuit portions of the inductor to ground. Similar arrangements are used in the mixer and high-frequency oscillator.

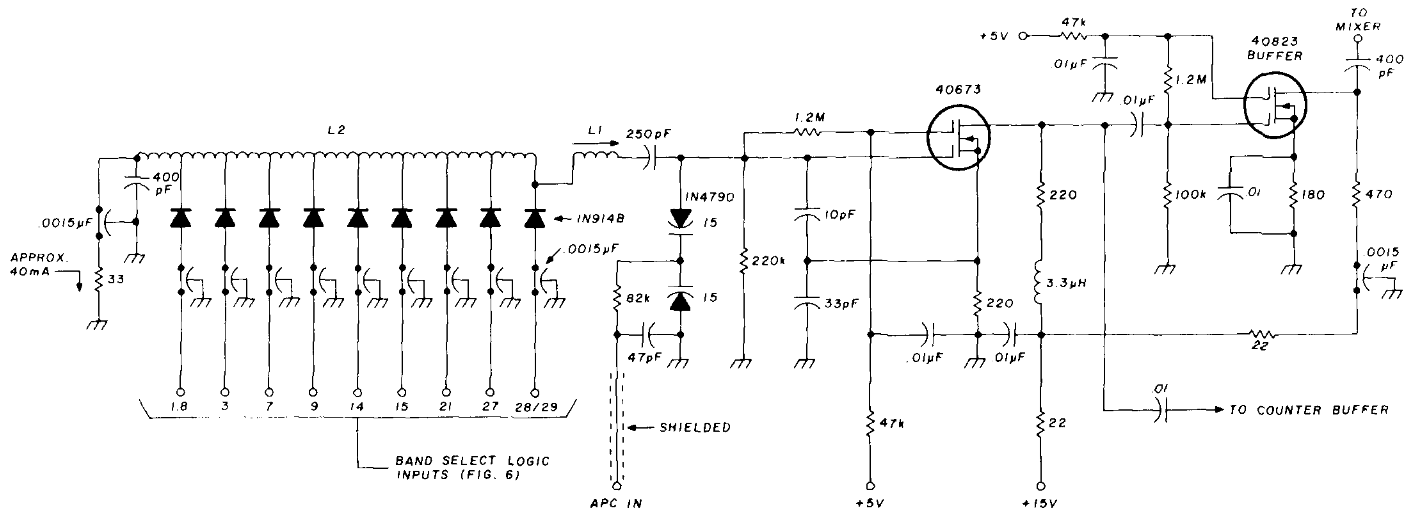


fig. 5. Circuit diagram of the digitally controlled high-frequency oscillator. Band-select logic is similar to that used in the rf amplifier and mixer circuits in that forward-biased diodes are used to change the value of L2.

are 141, the hfo frequency is 46100 kHz and the vfo tunes from 14100 to 14200 kHz.

Special care was given to the design of the vfo since it is free running and operates at a relatively high frequency. The design goal was for less than ± 2 kHz drift during the first ten minutes of warmup, and less than ± 100 Hz thereafter. Short-time drift and phase jitter were reduced to negligible values by careful bypassing and post regulation of the power supply.

A series of experiments were performed which were aimed at synthesizing the vfo, to provide it with 10-Hz steps in conjunction with a front-panel tuning knob. However, the approach using continuously variable vfo tuning was selected because of its infinite resolution which gives the operator that "smooth" feel. The ten-turn potentiometer used for vfo tuning gives more than adequate resolution, and allows ssb signals to be tuned in quickly and easily. One turn of the vfo knob represents 10 kHz of tuning range. The RIT control was added primarily for use in net operation.

The i-f circuits following the second mixer are fairly conventional, with the possible exception of the AM3705 analog multiplexers, which are used for mode switching. Changing modes is accomplished with a front-panel switch that outputs 3-bit binary words to drive the multiplexers located at appropriate places in the receiver. These multiplexers perform such functions as bypassing the ssb filter when operating a-m, selecting the upper or lower sideband bfo crystal, and inserting the narrow CW crystal filter into the rf chain in the CW mode. While the switching functions are illustrated on the block diagram by conventional switch symbols, all switching is actually done by binary control from the panel.

The i-f amplifier is a Motorola MC1590G which drives MC1596G product detector. This combination is ideal because it allows a single product detector to be used for a-m, ssb, and fm. The 1590 provides more than 60 dB of agc. In this receiver the agc range is extended by using delayed agc on both the rf and i-f preamplifier stages, the longest delay being associated with the rf stage.

The agc voltage is tapped off the agc amplifier and fed to an op amp; the op amp drives both the delayed agc circuits as well as the S-meter. This approach provides a simple way of using a 0-1 mA meter as the S-meter. Calibration of the S-meter was accomplished by inserting a Kay attenuator in the i-f circuit, feeding a signal to the antenna, and recording the deflection of the S-meter for various attenuation values. Full scale is adjusted by a potentiometer on the op-amp output, while zero is set by removing the antenna and adjusting the agc gain control associated with the MC1590.

A GE PA-237 integrated circuit is used as the audio amplifier; this drives a small speaker behind the panel of the 1650 kHz deck. This is an optional feature which admittedly was added for esthetic purposes to balance the front panel arrangement. Generally the output of the product detector is used to directly drive a stereo amplifier, although the built-in speaker allows the receiver to be used independently.

That part of the overall functional block diagram (**fig. 1**) which is inside the dotted lines belongs to the 32-MHz deck. It is primarily involved with the counters and display, the hfo and vfo, and the digital interface. The balance of this article will concentrate primarily on the details of these circuits, which collectively are the heart of the frequency synthesizer.

function of the counter

The counter plays an important role in controlling the hfo frequency, and it also provides the outputs to drive the display. **Fig. 2** is a block diagram which illustrates the basic scheme, but for clarity some of the functional circuits have been omitted. For example, DIP switches mounted on the counter board are used to program the TENS, UNITS, and TENTHS MHz counters. Only one set of these switches is shown in **fig. 2**. Logic inputs such as GATE, LOAD, and RESET circuits have been omitted. An excellent article by WB2DFA, which appeared in *ham radio*[2], illustrates many of the counter, logic, display, and time base generator circuits, some of which are used in this programmable receiver. Wiring diagrams of some of the more sophisticated up/down programmable counters will be described later. An article by Phillip Rand, W1DBM, in *QST*[3] is recommended if you want a better understanding of these circuits.

For the receiver described here to work properly, it is necessary that the hfo be phase locked to harmonics of 100 kHz derived from the 1.0 MHz clock. It is absolutely essential that the hfo be locked to the correct harmonic or line number. This is automatically accomplished by means of a ramp which tunes the hfo to within a few kHz of the correct spectral line. By definition, the hfo is part of a frequency synthesizer that has the capability to output any one of 281

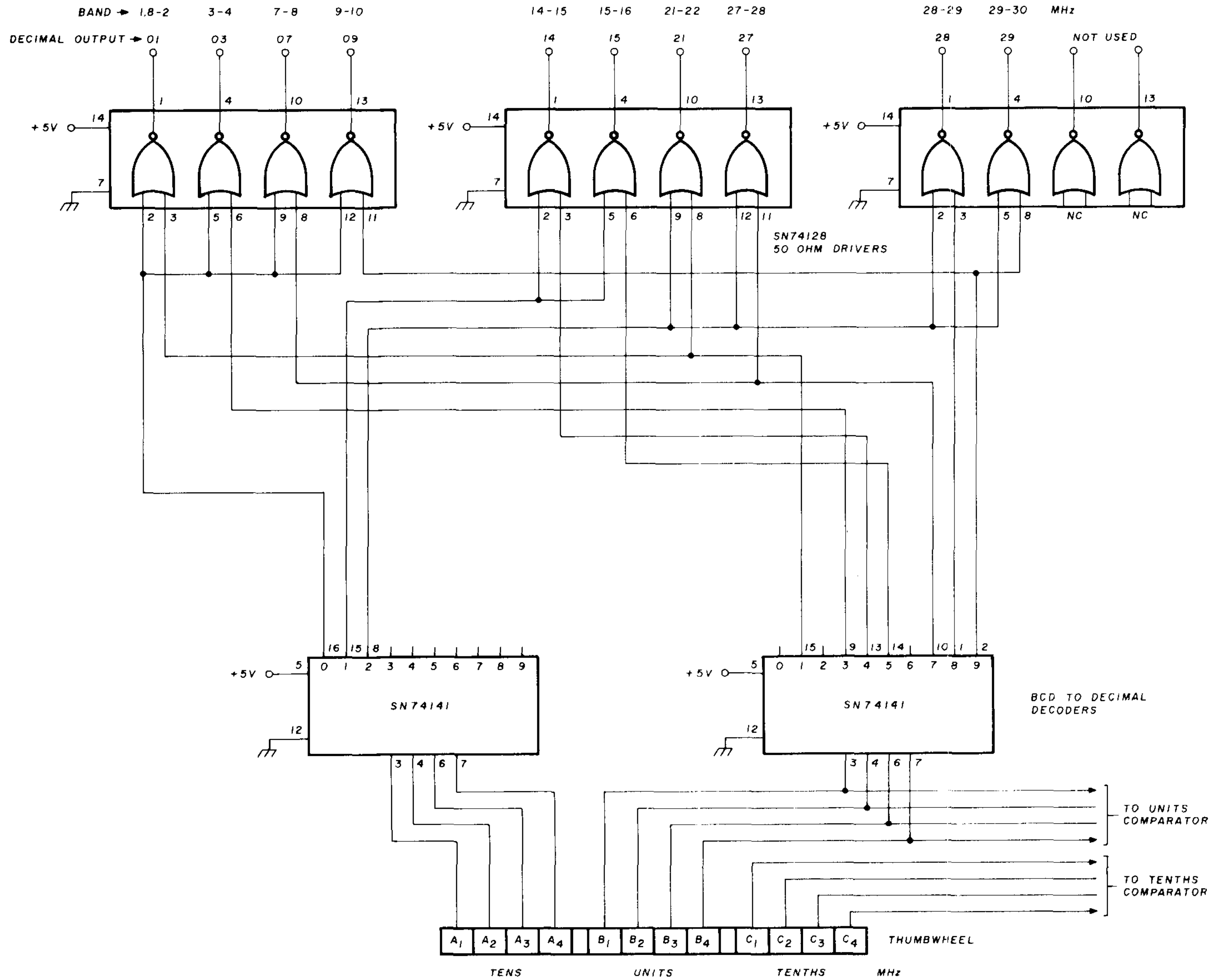


fig. 6. Digital band selector circuit provides binary commands for coil switching in the rf amplifier, mixer, and high-frequency oscillator. To maintain circuit tracking, the same voltage ramp is used to tune all three circuits.

table 1. Low band edge high-frequency oscillator (hfo) frequencies.

input signal frequency (MHz)	hfo frequency (MHz)
28.00	60.00
27.00	59.00
21.00	53.00
15.00	47.00
14.00	46.00
9.00	41.00
7.00	39.00
3.00	35.00
1.80	33.80

frequencies between 33.8 and 61.9 MHz. Specifically, these are 33.8, 33.9, 34.0, . . . 61.7, 61.8, and 61.9 MHz. In the breadboard receiver I built, 91 of these outputs are available, corresponding to a total frequency coverage of 9200 kHz. It is not difficult to modify the system to include additional bands or to change existing bands to cover other frequency ranges.

binary numbers counted, the ramp is inhibited. At this time, the hfo is close enough to the desired 100-kHz line that there can be no ambiguity. In theory, the frequency error immediately prior to lock-up is never more than 6.19 kHz, although in practice it may be slightly larger. **Fig. 2** includes functional blocks which illustrate how coarse tuning is accomplished in association with the display counter. Note that the third comparator which is associated with the HUNDREDTHS hfo MHz is hard wired to respond to zero. This counter stage is not associated with the display.

The tuning ramp is derived from a 566 function generator and a pair of 74191 up/down counters which drive an 8-bit DAC (digital to analog converter). The DAC provides infinite memory. When the ramp is interrupted, the ramp voltage corresponding to the value needed to coarse tune the hfo is retained indefinitely, or until a new frequency is programmed. **Fig. 3** is a schematic diagram of the ramp generator. Up/down and inhibit logic signals are derived from the cascaded comparators. With this approach, it is

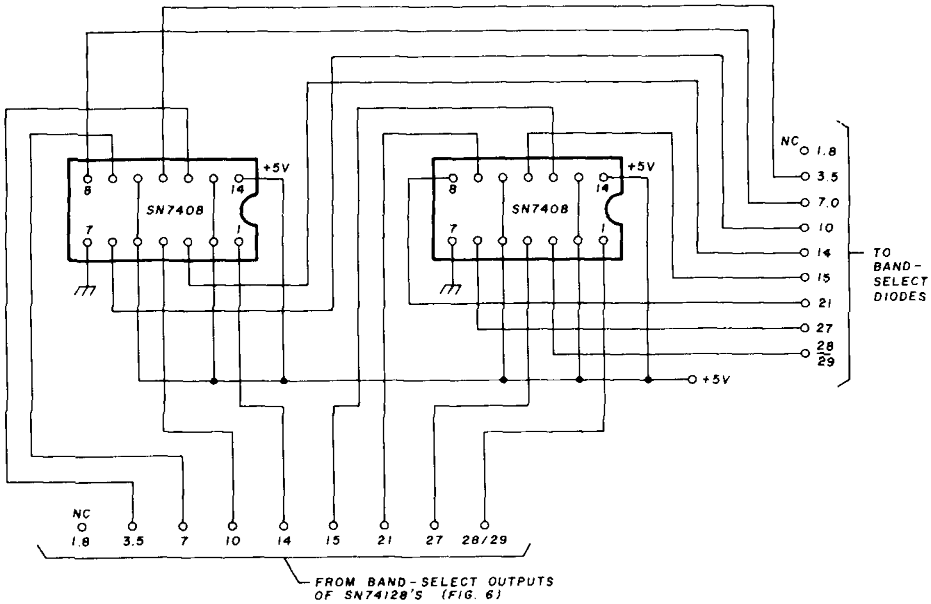


fig. 7. Schematic showing the use of 7408 two-input AND gates to drive the switching diodes in the rf amplifier and mixer circuits to reduce fan-out loading on the 74128 drivers in the band selector (fig. 6). Since the same tuning ramp controls the control signals to the hfo, mixer, and rf amplifier, tracking is maintained.

The key to synthesizer operation is the combination of programmable hfo counters and associated comparators. An 8-bit number is fed to the UNITS and TENTHS MHz comparators from the front panel thumbwheels when the desired frequency is dialed. A voltage ramp is enabled which tunes the hfo in the direction of the dialed frequency. When coincidence occurs between the binary numbers dialed and the

not possible for the hfo to lock to any 100 kHz line other than the correct one. A momentary interruption of power, or any other condition that unlocks the hfo, causes a repeat of the phase lock action.

front end tuning

A schematic diagram of the rf amplifier is shown in **fig. 4**. The rf coils are wired in series with 1N914s

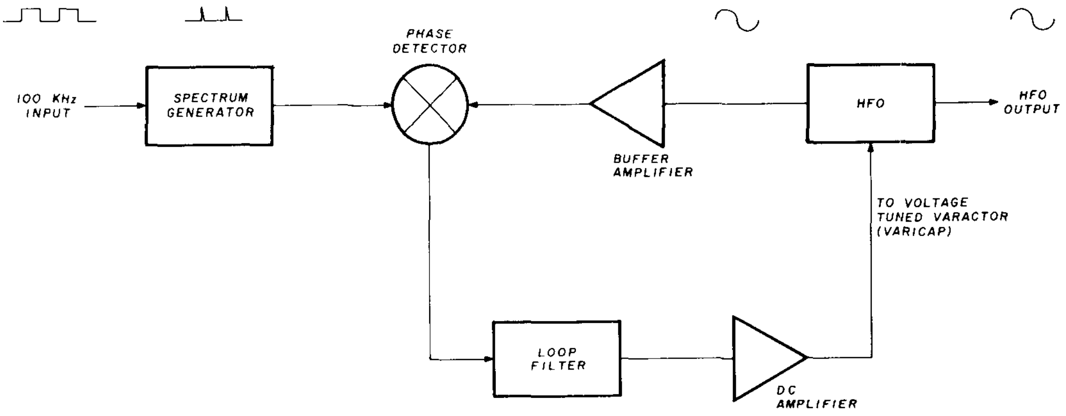


fig. 8. Functional block diagram of the phase-locked loop used in the receiver. The hfo provides one of two signals to the phase detector; the other input from the spectrum generator provides 100-kHz signals throughout the hfo tuning range. Loop bandwidth is about 10 kHz.

located at coil junctions which provide short circuits to ground. A particular band is selected by providing forward current to one of the diodes by band-select logic circuits which will be described. The rf circuit is typical of the mixer and the hfo, each having similar band selecting diodes. The ramp which tunes the hfo is also used to tune the rf and mixer coils.

hfo tuning and control

Since the hfo operates at a much higher frequency than the rf and mixer stages, rather than using separate coils for each band, a single coil wound with no. 14 (1.6 mm) tinned wire is used as shown in **fig. 5**. The inductance is self-supporting except for a polyethylene foam strip which is glued to one side of the coil with epoxy cement to make it rigid. The switching diodes are soldered to the coil at appropriate locations. Because of the inductance of the diode leads and the manner they are dressed away from the hfo coil, the correct diode locations are found by cut and try at first, those corresponding to the highest frequency bands being most critical. Each tap is adjusted so that lock up occurs with a ramp voltage of approximately 4.0 volts at a frequency corresponding to the low end of the particular band. **Table 1** lists the low end hfo frequencies corresponding to each of the 1.0 MHz bands in the receiver.

Start by dialing 28 MHz, and adjust the inductance of L1 until the hfo locks up at 60000 kHz with approximately 4.0 volts of ramp. Next, with 27 MHz programmed, adjust the position of the 27 MHz diode on L2 until lock up occurs with about 4 volts of ramp. Repeat this process, proceeding toward the 160-meter band. An auxiliary counter can be used to con-

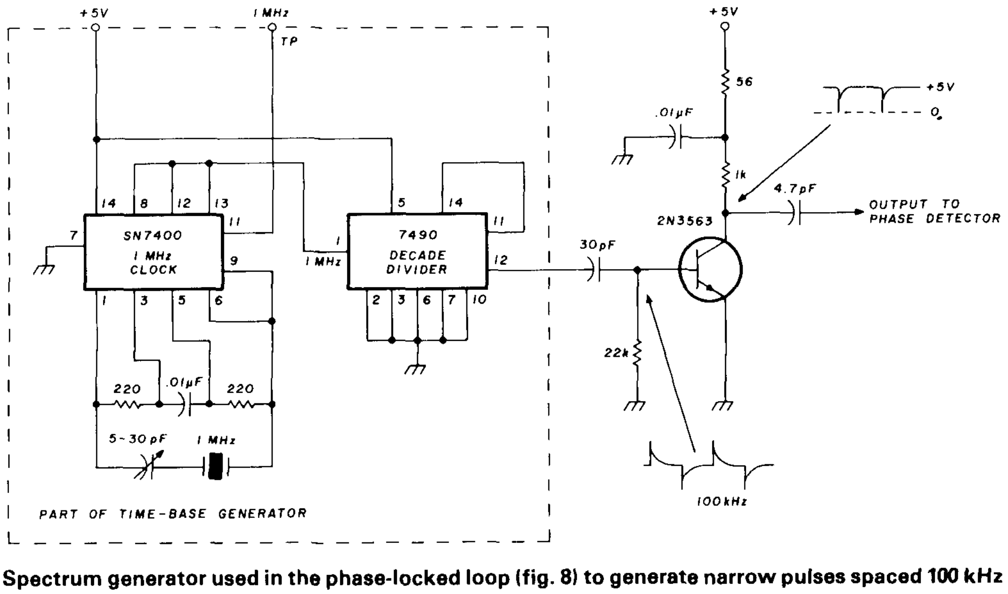


fig. 9. Spectrum generator used in the phase-locked loop (fig. 8) to generate narrow pulses spaced 100 kHz apart.

firm that the hfo locks up at the proper frequencies. If the hfo counters are properly programmed, the digital readout will agree with the thumbwheel numbers.

programming the counters

The TENS and UNITS MHz hfo counters are programmable. Two sets of DIP switches mounted on the back of the counter board contain the programming switches. Four of these switches are used to program each hfo counter stage.

Each counter can be programmed to initiate its count at any number from zero through 9. This is a means of advancing each digit to agree with the corresponding antenna frequency digit. If the heterodyne scheme described here is used, the TENS MHz counter should be programmed to 0111 (7) and the UNITS MHz stage for 1000 (8). The TENTHS MHz stage is hard wired to 0000 (zero).

The TENS kHz vfo counter is programmed for 0101 (5) while the UNITS and TENTHS kHz stages are hard wired for zero. In some heterodyne schemes, such as one which uses a 455 kHz i-f, the UNITS kHz stage should also be programmable.

band-select logic

The traditional way to change bands and select signals is by means of mechanical devices such as multi-wafer switches and ganged variable capacitors, although some CB and 2-meter receivers have employed frequency synthesizers for channel selection. More recently, high-frequency receivers have come on the market which use frequency synthesizers and digital displays. These are relatively expensive, but ultimately most high-frequency receivers will be manufactured this way.

Equipment for CB and 2 meters covers relatively narrow frequency bands. For example, a 40-channel

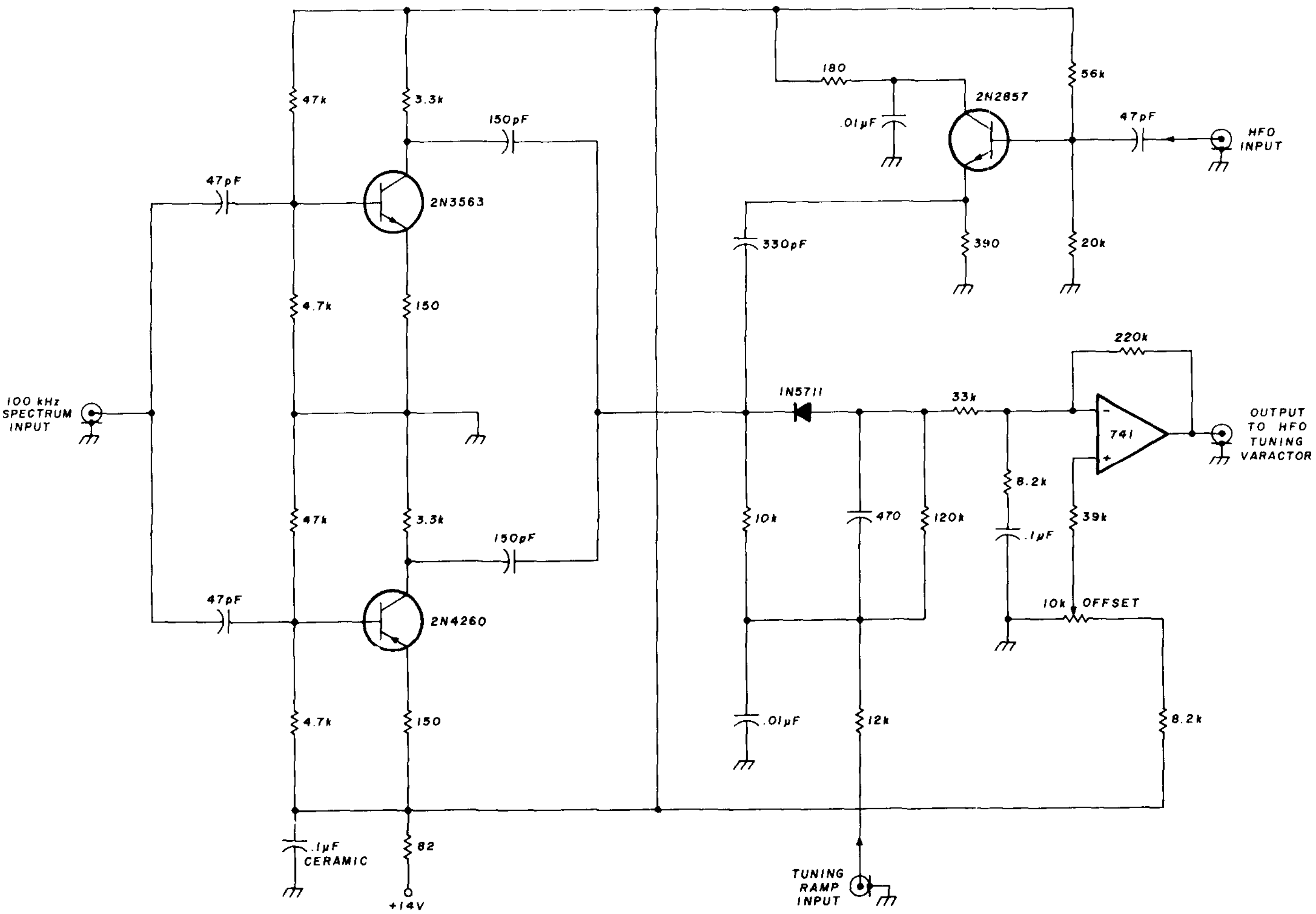


fig 10. Parametric phase detector developed by WA9HUV. The circuit uses a pair of fast complementary switching transistors to provide the required phase inversion without transformers. The gain of the 741 op amp is set at about 15 dB.

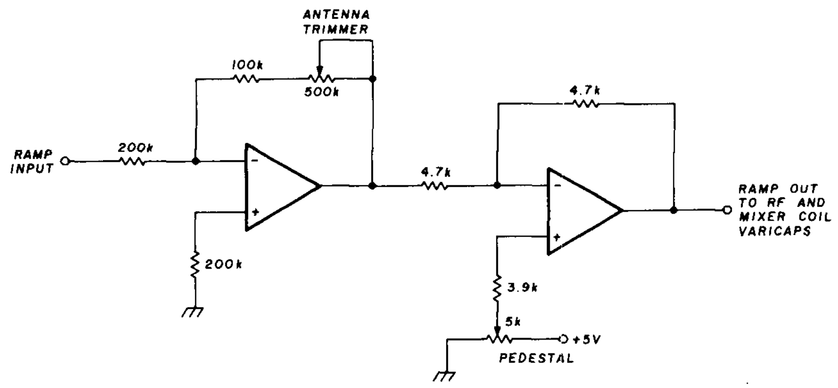


fig. 11. Unusual antenna trimmer circuit used in the receiver maintains tracking with the rf amplifier and mixer circuits.

CB receiver covers a frequency band having a high to low frequency ratio of about 1.6 per cent. By contrast, the high-frequency portion of this programmable receiver covers 33 per cent of the high-frequency band between 1.8 and 30 MHz. The technique of band changing through front panel thumbwheel switches, together with band-selecting diodes, allows a wide range of frequencies to be covered with relative ease.

The binary commands not only perform the rf, mixer, and oscillator coil switching function, but they program the comparators as well. A pair of SN74141 BCD to decimal decoder drivers and three SN74128 current drivers are used to provide TENS and UNITS MHz band-switching logic signals. The circuit of **fig. 6** shows how this is done. Note that the TENS 74141 provides a MHz logic output of 0, 10, or 20. The UNITS MHz 74141 provides a logic signal output corresponding to any number between zero and nine. The 74128s are current sources which are wired so that any one of 12 possible sets of coil selections can be made, although only ten of these are used.

To select 80 meters, for example, 03 is dialed. This provides BCD logic signals which turn on the 03 driver. The current supplied from this driver forward biases the appropriate 1N914 hfo diode, thus selecting the hfo coil corresponding to the 3.0 to 4.0 MHz band. Assuming 3500 kHz (035) is programmed, the ramp tunes the hfo until its frequency reaches 35500 kHz where it becomes phase locked. The vfo now tunes the band segment, 3500 to 3600 kHz.

The same 74128 logic signals are used indirectly to select the 3.0 to 4.0 MHz rf and mixer coils. SN7408 quad two-input AND gates driven from the 74128s supply current to the front end coil diodes to reduce fan out loading on the 74128s, as shown in **fig. 7**.

Note that the rf and mixer coils are tuned with the same ramp that tunes the hfo. This is how the rf and hfo circuits are made to track; details will be given later.

phase-locked loop

Having developed a way to coarse tune the hfo, the task of phase locking the hfo to the correct harmonic of 100 kHz becomes a relatively simple matter. **Fig. 8** is a block diagram which illustrates a second-order phase-locked loop such as the one employed here. The hfo, which is part of the closed loop, provides one of two signals for the phase detector. The other input to the phase detector is external to the loop and includes a spectrum of signals spaced 100 kHz apart. This spectrum extends across the hfo frequency range from 33.8 to 61.9 MHz.

The bandwidth of the closed phase-locked loop is limited to about 10 kHz so that there can never be an ambiguity between line selection. The loop can recognize only one line at a time, in spite of the fact that there are at least 281 individual sine-wave signals fed into the phase detector from the spectrum generator. Since it is not necessary to provide a way for selecting individual spectral lines, the spectrum generator, illustrated in **fig. 9**, is quite simple and consists of only a single transistor and a few discrete components.

The crystal clock and decade divider are part of the time-base generator. The 2N3563 transistor serves as a very fast switch which provides very narrow (16 ns) pulses to the phase detector. These narrow impulses include fairly uniform distribution of individual sine-wave signals extending from 100 kHz to well above 60 MHz. R-C coupling circuitry is arranged to reject most of the unneeded signals below 33 MHz; pulse shaping and parasitic capacitance causes the amplitudes of these signals to roll off rapidly above 60 MHz. It is necessary to shield the spectrum generator to prevent these signals from getting into the front end of the receiver. Otherwise, they will appear as markers at each end of each band segment.

phase detector

A survey of the various types of PLL ICs available to perform the phase detection function was disaptable 2. Tuning voltage for the front end and high-frequency oscillator.

input frequency (kHz)	tuning voltage (volts)	hfo frequency (MHz)	tuning voltage (volts)
7000	4.00	39.0	4.00
7100	4.51	39.1	4.25
7200	5.09	39.2	4.53
7300	5.75	39.3	4.85
7400	6.47	39.4	5.20
7500	7.29	39.5	5.60
7600	8.11	39.6	6.00
7700	9.14	39.7	6.50
7800	10.16	39.8	7.00
7900	11.40	39.9	7.60
21000	4.00	53.0	4.00
21100	4.16	53.1	4.20
21200	4.32	53.2	4.40
21300	4.50	53.3	4.64
21400	4.65	53.4	4.82
21500	4.86	53.5	5.09
21600	5.08	53.6	5.36
21700	5.30	53.7	5.64
21800	5.55	53.8	5.95
21900	5.80	53.9	6.27

pointing in terms of the requirements imposed by the programmable receiver. In many cases frequency response was the limiting factor — in others the cost was too high. Some earlier PLL chips that might have had promise were no longer available; more recent types are not only expensive, but also require considerable peripheral circuitry. I decided to settle for a homebrew design.

The phase detector used for the phase-locked loop is an original circuit which I developed. It uses a pair of fast complementary switching transistors which perform the required phase inversion without the need for transformers. This makes the circuit simple and very broadband as well. The circuit is shown in **fig. 10**. The complementary transistors would normally produce zero output, since the transistors are complementary and one output is 180° out of phase with the other. However, the hfo signal, which has a relatively large peak-to-peak amplitude, modulates the transistor collectors. If the relative phase of the hfo is other than 90° relative to the reference signal, one of the outputs tends to be suppressed while the other is enhanced. For this reason, the phase detector is referred to as a parametric phase detector. The hfo is the element that controls the amplitudes of the other two signals. The amount of unbalance depends on the relative phase of the hfo signal and the reference signal. When that angle is 90°, the outputs are equal and cancel.

The input to the Schottky diode acts as a common summing junction for the three signals: the zero reference, the 180° reference, and the hfo. If the output of the detector is plotted as a function of the phase angle, a discriminator type of curve results. Note that the phase detector curve sits on top of a pedestal which results from rectification of the relatively large hfo signal. This is unimportant since it is compensated for by the offset potentiometer associated with the op amp at the phase detector output. Furthermore, the offset adjustment allows the voltage for the tuning varactors to be set to the proper value for the low end of each hfo band.

When the hfo is not locked the output of the phase

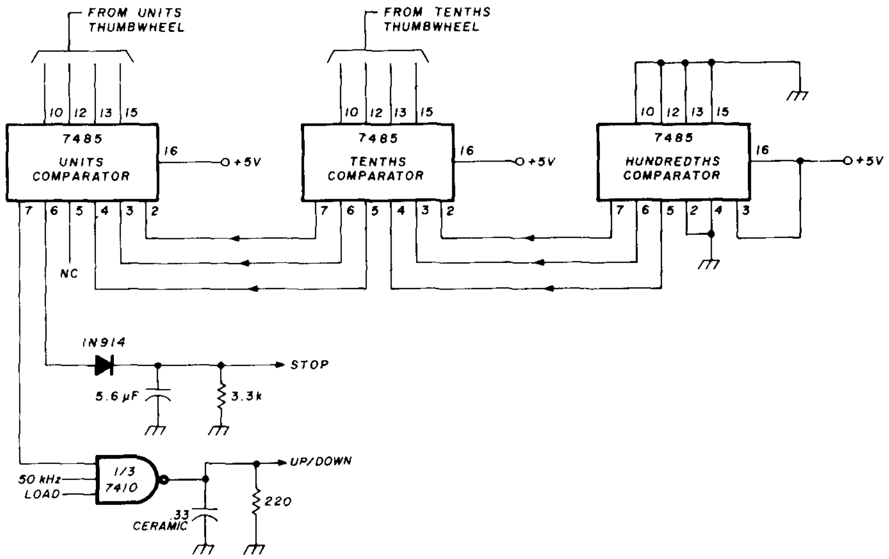


fig. 12. Simplified diagram of the cascaded comparators and their output filtering. This circuit provides the ramp sweep stop and directional steering commands to the 74191 up/down counters in the ramp generator (fig. 3).

detector is a sine wave at a frequency equal to the difference between the reference signal and the hfo. Lock up occurs almost instantaneously and is difficult to see on an oscilloscope.

The gain of the op amp is set to about 15 dB. If the phase-locked loop has a tendency to oscillate, the gain of this stage can be reduced by lowering the value of the 220k feedback resistor. It is not recommended that any of the other component values in the phase detector be changed.

spurious signals

In spite of care in shielding the spectrum generator, it was necessary to shield the rf section of the receiver carefully and to apply bypassing capacitors rather extensively at each terminal of the DIP power connector and at the BCD inputs to reduce these signals to levels below the ambient antenna noise.

Spurious signals which were much more difficult to control resulted from products of the hfo and vfo. Even though these signals are applied to separate mixers and are not intended to be associated, both the hfo and vfo signals are converted to TTL compatible levels in the counter where intermixing and harmonic generation results. These spurs can be classified as follows:

$$M(vfo) - N(hfo) = 1650\ kHz$$

The strongest spur occurs where $M = 2$ and $N = 1$ at 27025 kHz. A smaller spur occurs at 14330 kHz where $M = 3$ and $N = 2$; a relatively weak spur was found at 7012.5 kHz where $M = 4$ and $N = 3$.

The ideal way to eliminate these spurious signals is to install a bandpass filter between the first mixer output and the second mixer input. This filter should cover a band from 31.9 to 32.0 MHz and should have sharp skirts. Some excellent filter design articles are included in the list of references.[4,5]

antenna trimmer

The circuits associated with the antenna trimmer bear little resemblance to conventional antenna trimming circuits. However it is a very effective technique which has the added advantage that tracking between the hfo and the rf circuits is accomplished at the same time.

The voltage ramp that tunes the hfo is also fed to the rf and mixer tuning varactors. A front panel potentiometer control serves as the antenna trimmer. It operates in conjunction with the circuit described in **fig. 11** which includes a pair of op amps.[6]

The way in which the antenna trimmer works is best illustrated by an example. First, set the thumbwheels to select 7.0 MHz (070). The ramp voltage

frequency to be displayed	hfo frequency
01800	33800
02000	34000
03000	35000
04000	36000
05000	37000
06000	38000
07000	39000
08000	40000
09000	41000
10000	42000
11000	43000
•	•
•	•
•	•
16000	48000
17000	49000
18000	50000
19000	51000
20000	52000
21000	53000
•	•
•	•
•	•
26000	58000
27000	59000
28000	60000
29000	61000

Programmed to advance 8 (1000)

Programmed to advance 7 (0111)

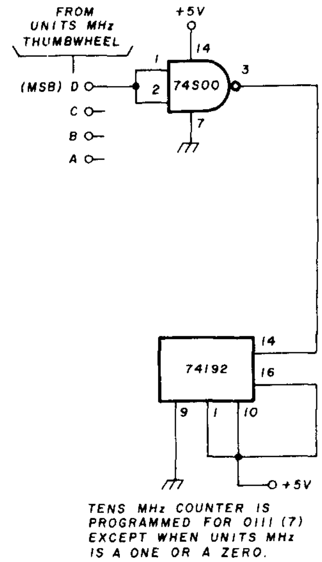
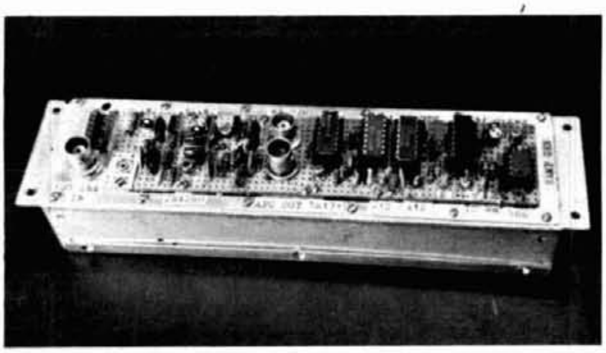


fig. 13. Logic correction circuit which resolves a minor problem with the TENS MHz counter. When an 8 or 9 is programmed, the TENS MHz counter and corresponding display advances one extra digit because the TENS MHz digit advances. An example is given in the text.

corresponding to the low end of each band is approximately 4.0 volts, which is the value required to phase lock the hfo to 39000 kHz. As can be seen in **table 2**, as the hfo is incremented from 39000 to 39900 kHz (which tunes the receiver from 7000 to 7900 kHz), its tuning voltage increases from 4.0 to 7.6 volts. Based on the sizes of the varicaps used and the total circuit capacitance, the rf and mixer varicaps require a tuning voltage range of 4.0 to 11.4 volts to tune from 7000 to 7900 kHz. Thus, the change in ramp voltage for the front end must be amplified by the op amp by a factor of 2.055 to achieve tracking with the hfo.

The situation is different in the 21000-21900 kHz band where the hfo tuning ramp varies from 4.0 to 6.27 volts. However, the rf and mixer circuits only need a variation of 4.0 to 5.8 volts. Therefore, for the amateur 15-meter band, the ramp slope must be amplified by a factor of 0.7929. The antenna trimmer control automatically makes this gain adjustment when it is adjusted for resonance or maximum signal strength. The 5k pedestal control is set to provide 4.0



Construction of the module containing the spectrum generator (left), phase detector (center), and ramp generator (right).

volts output corresponding to the low end of each band. Once this adjustment has been made, it does not have to be readjusted.

front end alignment

The antenna trimmer circuit provides a convenient way to align the front end coils. Start the alignment with 10 meters and proceed in numerical order down to 160 meters. This will avoid interaction between coil inductance settings. The pedestal potentiometer is first set for 4.0 volts output corresponding to the low end of the band. The front end coils are then aligned. Next, the high end of the band is programmed by means of the thumbwheel switches and the antenna trimmer is adjusted for maximum rf gain. Without changing the setting of the antenna trimmer, realign the rf coils at the low end of the band. Again, check the antenna trimmer setting at the high end. Several iterations of this kind are required until the inductance of the coils is set so that tracking is achieved without having to radically change the antenna trimmer setting when going from the low to high end of the band.

The same procedure is repeated for each of the other bands. Once the coils have been aligned in this manner, very little adjustment of the trimmer should be required after the initial adjustment when a new band segment is selected.

comparators

The ramp sweep stop and directional steering commands are derived from the three hfo comparators as previously explained. Sweep control is accomplished by cascading the SN7485s as shown in **fig. 12**. Since the 74191 up/down counters in the ramp circuit need only a single up/down command, pin 5 of the 191s are tied high. However, when the ramp is enabled, if the up/down logic level is low, the counters count up, which increases the ramp voltage. Therefore, the MORE THAN cascaded output of the 7485s is connected to pin 5 of the 191s. The EQUALS output which stops the ramp is high only when all of the 12 counter bits agree with the bits programmed, including the hard wired HUNDREDTHS MHz bits.

Since the counters operate in two modes, namely COUNT and DISPLAY, the bits are changing during the COUNT mode, causing the comparators to output fluctuating logic data. To overcome this problem, both of the cascaded comparator outputs are modified to provide suitable output logic signals. These relatively simple circuits are shown in **fig. 12**. Because of the requirement to filter these control signals, there is a finite delay between comparator coincidence and application of the logic commands. This sets an upper limit to the ramp sweep rate because some overshoot of the hfo frequency results. If the sweep rate is too fast, overshoot will be sufficient to change the comparator output signals before the sweep circuit is stopped, which means that the sweep lock circuit will keep oscillating and lock up will never occur.

A total ramp sweep interval of about two to four seconds is satisfactory. Only when changing bands is there any noticeable delay in hfo lock up. When incrementing the thumbwheel switches from one band segment to another, there is no noticeable delay.

There is a trivial problem that is related to the TENS MHz counter. Whenever either an 8 or 9 MHz is programmed, the TENS MHz counter and corresponding display advances one extra digit because the corresponding TENS MHz digit advances. To avoid this problem, the simple logic circuit shown in **fig. 13** is used. This circuit prevents the TENS MHz counter

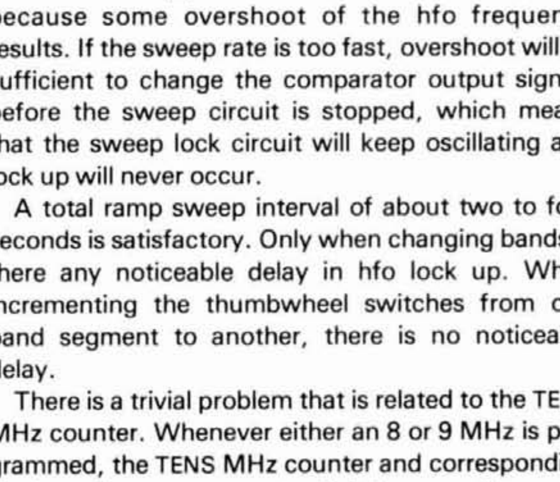


fig. 14. Resistance of a forward-biased 1N914 diode at 1 kHz, as measured with a GenRad 1650B impedance bridge. Diode resistance is important because it affects the Q of the tuned circuits used in the receiver.

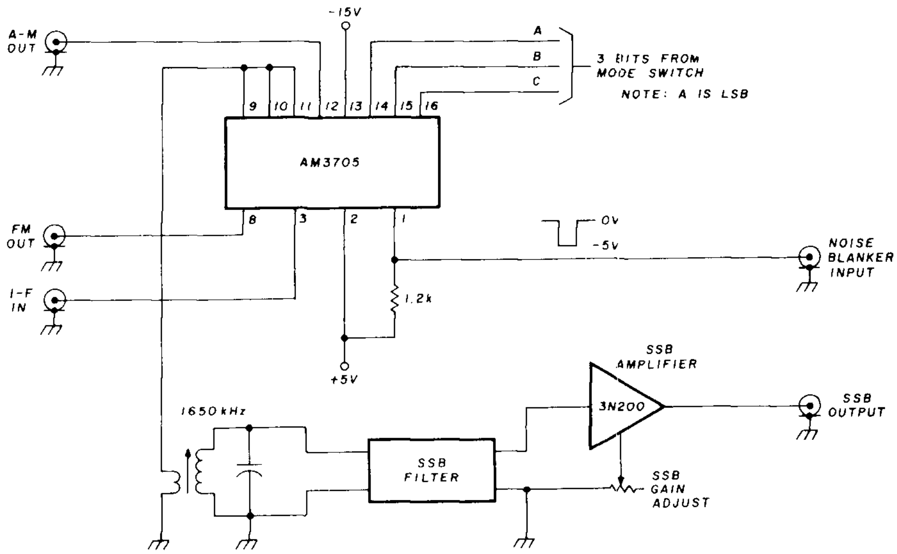


fig. 15. Diagram of the ssb mode select used in the receiver. This is typical of the analog multiplexing circuits operated from the front-panel.

from advancing whenever the UNITS hfo digit is a one or a zero. When dialing 27.0 MHz, for example, the hfo frequency is 59 MHz, but when 28 MHz is dialed, the hfo advances to 60 MHz. This change in the most significant digit from 5 to 6 would normally advance the display from 27 to 38. The logic circuit of **fig. 13** solves this problem.

switching diodes

As previously explained, 1N914 diodes are used for front end and hfo coil switching. This approach was found to be both simple and effective. It should be recognized, however, that the forward resistance of the diode has a tendency to reduce the *Q* of the tuned circuits. To minimize this effect, the L/C ratio of the front-end coils has been made as large as possible. **Fig. 14** shows the diode resistance as a function of forward current. Note that the resistance drops to about 6 ohms at a current of 40 mA, and any further increase in current has a small effect on resistance. Some simple calculations show that with a circuit capacitance of 20 pF at 28 MHz, if the loaded circuit *Q* is 50 without the diode, it drops to 23 with the diode; this both increases the rf bandwidth and reduces gain. It was found that *Q* and gain could be increased to acceptable levels if parallel-connected diode pairs were used with the 10-meter coils.

The effect of the switching diodes is less pronounced on the lower bands (it is practically negligible at 1.8 MHz). It is possible to control front-end gain so that it is nearly the same on all bands by adjusting the currents in the diodes.

mode-select circuitry

The circuit of the ssb mode-select circuit is shown in **fig. 15**. This is typical of each of the analog multiplexing circuits operated from the front-panel mode switch. While only five outputs are used, the switch has 8-pole capability.

summary

The development of this programmable receiver was a much more formidable task than I originally envisioned. The receiver as it exists presently represents a first phase effort, and much yet remains to be done in terms of refinement. The basic idea has been proven to be sound, however, and the result is a high performance breadboard receiver of advanced design.

It is hoped this article will provide other experimenters with new ideas and incentives to try their hand at something radically new. Additional circuit details can be made available to those hearty experimenters who are interested in duplicating this receiver in part or as a whole. Please send me a self-addressed, stamped envelope for further details. Readers' suggestions and constructive criticism are welcomed.

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ham radio

super low-noise 432-MHz preamplifier

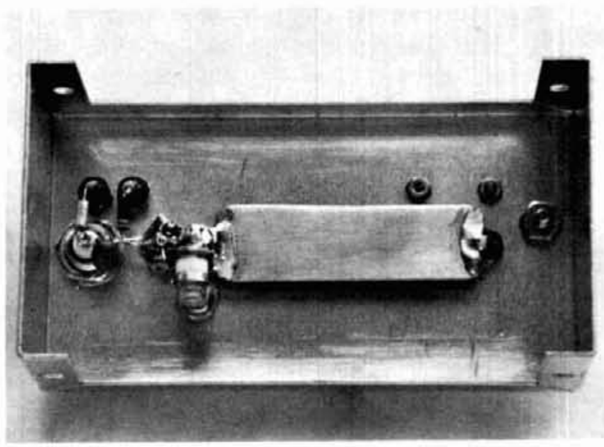
Construction of a low-noise bipolar transistor preamplifier which offers 0.8 dB noise figure with 15 dB gain

The heart of any successful moonbounce (EME) system is the low-noise preamplifier which precedes the receiving converter. On 144 MHz, an overall system noise figure of less than 1.0 dB seldom increases receiver sensitivity significantly because of the high sky noise temperatures of 300° to 400° K. On 432 MHz, however, sky noise temperatures of 10° to 20° K are possible; this means that decreasing the overall system noise figure below 1 dB significantly improves receiver sensitivity.

Until recently the popular Fujitsu FJ203 and the Fairchild FMT4575 bipolar transistors have yielded the lowest noise figures at 432 MHz — typically 1.25 dB. The Texas Instruments MS2110, although not as well known, also produces noise figures as low as 1.1 dB at this frequency. Shigeru Sando, JH1BRY, introduced an excellent 432-MHz preamplifier using NEC's V244 GaAs fet which opened the eyes of many moonbouncers.[1] With its impressive 0.6-0.7 dB noise figure, it beat everything else at the noise-figure contests. However, its cost of $120 makes it a luxury for most amateurs.

Recently NEC introduced a new bipolar device, the NE64580, which is rated at 0.8 dB noise figure at 500 MHz. At $92 each, it appeared to be a mighty competitor to the V244. Even better, the people at NEC came out with the NE64535 at a cost of only $17 in single quantities;* it is rated at 1.6 dB noise figure at 2 GHz with an f_T of 8.5 GHz. At 500 MHz, the NE64535 has a rated noise figure of 0.8 dB.

The NE64535 uses the same chip as the NE64580 but is mounted in a less expensive hermetically sealed *Micro-X* package. This article discusses the design of a 432-MHz preamplifier that uses this



Construction of the low-noise 432-MHz preamplifier, showing the placement of the stripline resonator. Output connector J1 is to left; input SMA connector is partially hidden by the piston capacitor C1.

By Al Ward, WB5LUA, RR2, Box 65A, McKinney, Texas 75069

*The device is manufactured by the Nippon Electric Company (NEC) in Japan and is being marketed by California Eastern Laboratories (CEL), Post Office Box 915, One Edwards Court, Burlingame, California 94010. Cost is $17 each in quantities of 1-9, decreasing to $15 each for quantities of 10-99.

device to obtain measured noise figures as low as 0.8 dB. The original design was based on the more expensive NE64580, but identical results have been achieved using the NE64535.

At the 1977 convention of the Central States VHF Society in Kansas City, Missouri, this preamplifier measured only 0.1 dB higher noise figure than the V244 GaAs fet entered by K2UYH; all other bipolar entries had approximately 0.3 to 0.4 dB higher noise figures. At 432 MHz, this decrease in noise figure over other bipolar devices results in a significant increase in receiver sensitivity.

Since I was intrigued by the fact that Shigeru Sando was able to use a parallel tuned circuit on the input to his V244 preamplifier and still achieve a low noise figure, I decided to try something similar. I wanted to obtain a low-loss match for minimum noise figure and still achieve adequate selectivity so I wouldn't require an external cavity.

The final design uses a parallel tuned circuit with capacitive coupling on the input (**fig. 1**). To minimize circuit losses I used a low-loss microstrip line rather than a lumped inductor. Resistive loading is used on the output and will be discussed later. I am presently using this preamplifier without an external cavity and have experienced no problems with intermodulation. Since the preamplifier is capacitively coupled at the input, greater rejection of unwanted signals will occur below 432 MHz. When the preamplifier is adjusted for minimum noise figure at 432 MHz, 10 dB of rejection is typically obtained at 200 MHz, 26 dB at 100 MHz, and 40 dB at 50 MHz. In all but the worst rf environments this should be adequate.

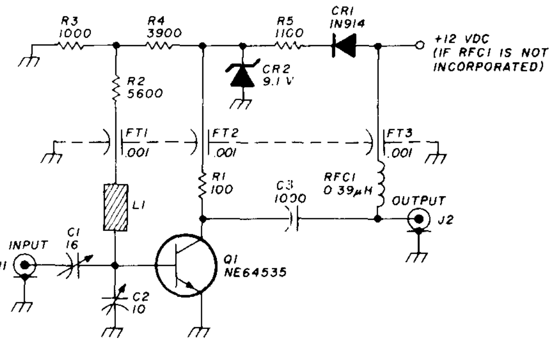
At my location, for example, a broadband FJ203 preamplifier cannot be used without a cavity, whereas the NE645 preamplifier has given no problems at all (my location is within 15-20 km of fm and TV transmitting antennas). If external filtering is required, a cavity filter with 0.2 dB loss described by Joe Reisert[2] will increase the noise figure to only 1.0 to 1.2 dB, which is still a worthwhile improvement.

design

Lacking any sort of tables describing the optimum source impedance required for minimum noise figure, I initially designed the input circuit for maximum gain using the published *s* parameters for the device. The input circuit was then optimized for lowest noise figure with a Hewlett-Packard HP-342A automatic noise figure meter. The resultant circuit uses a microstrip line with a characteristic impedance of 70 ohms.

An attempt was made to tune the output circuit to increase selectivity, but, as expected, completely stable operation was not obtained. As with most microwave bipolar transistors operated in this frequency range, maximum available gain is so high that oscillations are common. Stable operation of this device was finally achieved with resistive loading. This method of stabilizing 432-MHz preamplifiers was suggested in an earlier article on the FMT4575.[2] However, the selection of the 100-ohm collector resistor was not arbitrary; in addition to stabilizing the preamplifier, it also serves to provide a better match to the post-amplifier. Resultant output vswr of this preamplifier should be less than 2.0:1. When using a post-amplifier with a variable match at the input, no problems have been incurred in obtaining stable operation.

While optimizing noise figure, I required some



C1	0.8-16 pF air variable (Johanson 5200 series)
C2	0.8-10 pF air variable (Johanson 5200 series)
C3	1000-pF mica or ceramic disc (not critical)
CR1	1N914 or 1N4148 silicon diode
CR2	9.1-volt zener (1N757)
FT1-FT3	470-1000 pF feedthrough capacitors
J1	SMA coaxial connector
J2	BNC coaxial connector
L1	microstrip line 15 mm (0.6 in.) wide, 53 mm (2.1 in.) long, mounted 5 mm (0.2 in.) above chassis
RFC1	0.39-µH miniature rf choke

fig. 1. Schematic diagram of the low-noise preamplifier for 432 MHz; typical noise figure is less than 1.0 dB. The NE64535 transistor is manufactured by Nippon Electric Company (NEC) and costs $17 in small quantities.

method of varying the dc bias conditions to determine their effect on noise figure. The bias circuit I used allows optimization of both V_{CE} and I_C for minimum noise figure by varying resistors R2 and R5. This method uses both voltage feedback and a constant base current source to ensure dc stability.[3] (Bias circuit design using this arrangement is discussed on page 39.)

The data sheet for the NE64535 specifies $V_{CE} = 8.0$ volts at $I_C = 7.0$ mA. It was found experimentally that minimum noise figure at 432 MHz occurs at a V_{CE} of

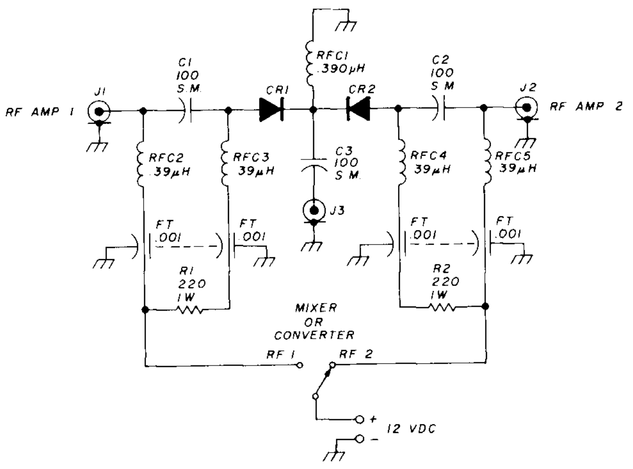


fig. 2. Combination bias tee/PIN diode switch used to operate separate preamplifiers for 432-MHz EME and tropo. The PIN diodes (CR1 and CR2) are Microwave Associates type MA47110.

6.0 volts with I_C approximately 4-5 mA. The new operating conditions decreased noise figure by about 0.2 dB. Although not as critical as the input match, the dc operating conditions can also be optimized for lowest noise figure at a particular operating frequency.

Zener diode CR2 was incorporated mainly to protect the collector-emitter junction, since the device has a maximum V_{CE} rating of only 12 volts. Not only does it protect the device as you optimize bias conditions, but it also protects the transistor from transients that may be present on the 12-volt supply line.

table 1. PIN diode switch performance on three vhf amateur bands.

frequency	insertion loss	isolation
144 MHz	0.25 dB	22 dB
220 MHz	0.60 dB	20 dB
432 MHz	1.20 dB	16 dB

For maximum effectiveness, the preamplifier should be mounted at the antenna. With RFC1 installed in the circuit, +12 volts can be conveniently run up the receive coax. At my station I use separate preamplifiers for EME and tropo operation feeding a common mixer, so I needed a convenient method of switching between preamplifiers. Since good mechanical coaxial relays are expensive, I devised the combination bias tee/switch arrangement shown in **fig. 2**. The switching elements are inexpensive, readily available PIN diodes.

When the PIN diodes are forward biased with 50 mA of current, their insertion loss is slightly less than 1.25 dB at 432 MHz. With no bias applied, the isolation between ports is 16 dB. Isolation is defined as the insertion loss to the *off* port. Since +12 volts is switched between preamplifiers at the same time that the control bias is transferred, tuning interactions between the EME and tropo preamplifiers are kept to a minimum.

The isolation can be improved by applying reverse bias to the PIN diode in the off port leg. The reverse bias decreases the diode capacitance, thereby increasing the isolation to the off port. The amount of reverse bias that can be applied is limited by the reverse breakdown voltage specified for the diode. Since a dual polarity power supply is not available at my station, I chose not to reverse bias the PIN diode. There are many versions of the PIN diode switch that could be used to increase isolation — but they are beyond the scope of this article.

The usefulness of the PIN diode switch can be extended to switching various local oscillators that supply injection to a broadband double-balanced mixer for multiband operation. As shown in **table 1** the combination bias tee/PIN diode switch arrangement performs even better at lower frequencies. The PIN diode used here is a Microwave Associates MA47110 available for 99 cents in small quantities.

construction

The preamplifier is built in a 108 × 57 × 38 mm (4-1/4 × 2-1/4 × 1-1/2 inch) Minibox, which is both inexpensive and readily available. The rf circuitry is mounted on the inside while all the bias components are mounted on the top of the minibox (**figs. 3** and **4**). This allows greater isolation between the rf and dc components than if all components were mounted inside the enclosure.

The microstrip line, 15 mm (0.6 inch) wide and 53 mm (2.1 inches) long is mounted approximately 5 mm (0.2 inch) above the chassis. The corners are rounded off to a radius of 1.5 mm (1/16 inch) to minimize the discontinuities at the end of the microstrip line. The 0.001-µF feedthrough capacitor is used as a

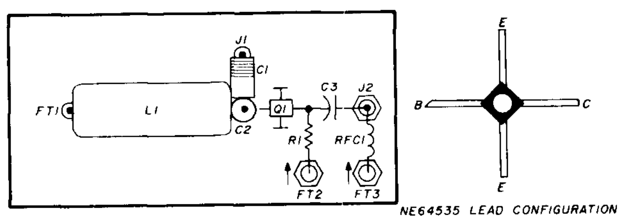


fig. 3. Layout of the rf components in the low-noise 432-MHz preamplifier. The input network is formed by C1, C2, and L1; the output network consists of R1 and C3. Bias circuitry is installed on the outside of the Minibox enclosure (see fig. 4).

support for one end of the microstrip line; the opposite end is soldered to the Johanson variable capacitor, C2. The input matching capacitor, C1, is soldered directly between the input connector and capacitor C2. An SMA-type connector is used on the input. Its small size and low loss make it a must for low-noise operation. A less expensive BNC type connector is used on the output.

To facilitate direct grounding of the two emitter leads, I used two solder lugs bolted to the chassis to serve as tie points. The lugs are cut off so they stand up from the chassis about 3 mm (0.1 inch). This allows enough area to conveniently solder the emitter leads. Keep the emitter leads on the device full length.

The leads on C3 and R1 must be as short, and as far away from the input circuitry, as possible to reduce any chance of feedback. If the preamplifier is built according to the layout in **fig. 3** no shield will be required between the input and output circuitry.

When using a Minibox as an rf enclosure, be sure to scrape off any paint or any other nonconductive film that may be on the areas where the two halves of the enclosure meet. This is best done before assembly has been started, and is necessary to achieve a good rf-tight enclosure. Be sure to use all four screws supplied with the Minibox.

operation

Connect the output of the preamplifier to the post-amplifier or converter with a short section of 50-ohm coaxial cable. Terminate the input with 50 ohms. With +12 volts powering the preamplifier, the total current drain should be 5 to 6 mA for lowest noise. The actual collector current will be about 1 mA less than the total current due to the current being drawn by resistors R3 and R4 in the bias circuit. At the rated 5-6 mA of current, V_{CE} will be 5-6 volts. Since the actual collector current drawn from the power supply is a function of the dc current gain (h_{FE}) of the device, the value of R2 may have to be adjusted slightly to achieve the desired amount of collector current. To date, all of the devices tested have achieved similar operating parameters and lowest noise figure with-

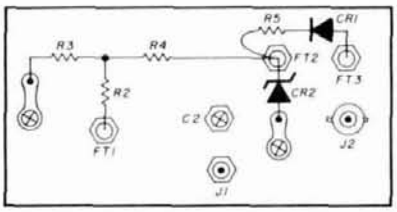
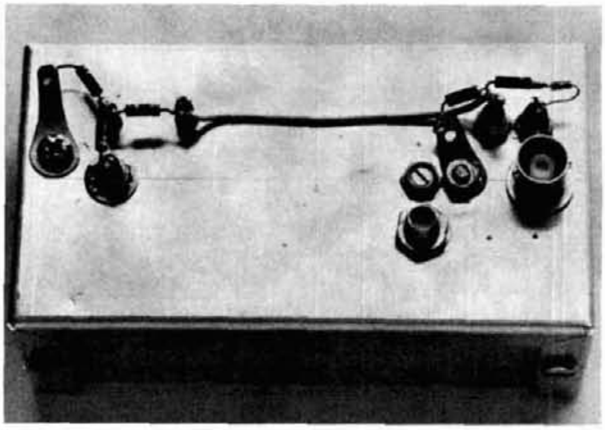


fig. 4. Top view of the 432-MHz preamplifier showing the layout of the bias circuit.



Top view of the 432-MHz preamp showing the layout of the bias resistors (see fig. 4). BNC output connector is at far right; SMA input connector is to the left.

out any modifications to the bias network shown in **fig. 1**.

Optimizing the input network for lowest noise figure is most easily done with an automatic noise-figure meter, but precise tuning can still be achieved by using a weak-signal source or a simple noise generator. Start with C1 at about half capacitance and then minimize noise figure with C2. Increase C1 slightly and then repeak C2 for minimum noise figure. Do not increase the capacitance of C1 past the point where minimum noise figure occurs. Overcoupling with C1 broadens the frequency response of the preamplifier with no improvement in noise figure. Finally, peak the post-amplifier stage into the preamplifier for minimum noise figure.

All devices I have tested so far have yielded noise figures between 0.8 and 1.0 dB; associated gain at minimum noise figure varies from 14 to 16 dB. With the added selectivity and lower noise figure obtainable with this device, this preamplifier should make for a significant improvement in the reception of EME signals as compared with other bipolar devices presently on the market. It also does a good job of challenging users of GaAs fets at noise-figure measuring contests! After nine months of operating 432-MHz EME with this preamplifier, I am convinced that it has made a worthwhile improvement in the reception of weak signals.

references

1. Shigeru Sando, JH1BRY, "Very Low-Noise GaAs Fet Preamplifier for 432 MHz," *ham radio*, April 1978, page 22.
2. Joseph Reisert, W1JR, "Ultra Low-Noise UHF Preamplifier," *ham radio*, March 1975, page 8.
3. Kenneth Richter, "Design DC Stability Into Your Transistor Circuits," *Microwaves*, December 1973, page 40 (see also Hewlett-Packard Application Note 944-1).

ham radio

superhet tracking calculations

How to choose component values for tuning rf and local-oscillator stages in superhet receivers

Superheterodyne receivers which have ganged tuned capacitors for simultaneous tuning of the local oscillator and signal frequency circuits require a special design approach. When such circuits are correctly designed and adjusted, they are said to "track," meaning that each resonant circuit is correctly tuned for any frequency setting of the receiver's tuning dial. Errors in tracking, if large enough, cause loss of receiver sensitivity.

The following method of calculating component values for superhet tuning circuits is not new, but I've tried to reduce the procedure to its essentials. Interested readers who want to pursue the topic should review **reference 1**, which lists other works on the subject.

the problem

Fig. 1 illustrates what is to be accomplished. As the tuning capacitor is rotated, the receiver's input circuits must tune from the lowest signal frequency, f_1, to the highest signal frequency, f_2. At the same time, the local oscillator (LO) must tune to a frequency which is always equal to the signal frequency plus the intermediate frequency, the i-f being a constant fixed frequency. Although circuits can be designed so the LO frequency is lower than the rf or signal frequency, the method described here requires that the LO be higher than signal frequency.*

Fig. 2 shows the component arrangements for the signal and oscillator tuning circuits. In the signal circuit, C_T represents the distributed capacitance of the coil, plus the minimum capacitance of the variable tuning capacitor, plus any fixed capacitance necessary to adjust the circuit. Capacitance C_G is the variable capacitance of one gang of the tuning capacitor used for the signal frequency. C_{Gmax} is the difference between minimum and maximum values of the variable capacitor. If a variable capacitor section can be adjusted from a minimum value of 10 pF to a maximum value of 365 pF, for example, then C_{Gmax} for that capacitor is 355 pF.

Capacitor C_{TL} in the oscillator circuit represents the distributed capacitance of the oscillator coil; its value is found by measuring the self-resonant frequency of the coil with a grid-dip meter; then, knowing the inductance, the capacitance may be calculated. In many cases, however, C_{TL} will be so small, compared to the other circuit capacitances, that it may be neglected. C_p is called the padder capacitor; C_{Tc} represents the minimum capacitance of the oscillator section of the tuning capacitor plus any capacitance needed for correct adjustment; C_{Go} is the variable capacitance of the oscillator gang on the tuning capacitor. C_{Gomax}, used in the equations, is the difference between minimum and maximum values of C_{Go}. It is not required that the oscillator section of the ganged tuning capacitor have the same capacitance as the rf sections, but, for the equations here, its percentage of maximum capacitance *vs.* angle of shaft rotation should be the same as for the rf sections. In other words, the rotor plates of the capacitor should all have the same shape.

design equations

Units for the following equations are microhenries for inductance, MHz for frequency, and picofarads

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*To reduce problems with spurious signals, the local oscillator should be above the signal frequency. **Editor**.

for capacitance. Equations are listed in the order in which they must be solved, so that values needed for a particular equation will have already been determined. It is first necessary to define the signal and i-f frequencies, then make a few preliminary calculations regarding their relationships. Also, the difference between minimum and maximum values of the tuning capacitor sections must be determined; this is best done with an accurate capacitance meter or bridge. I am hesitant to accept vendor's ratings, especially when buying variable capacitors on the surplus market.

f_1 = minimum signal frequency

f_2 = maximum signal frequency

f_i = intermediate frequency (i-f)

$$A = \frac{f_2}{f_1} \qquad A^2 = \left(\frac{f_2}{f_1}\right)^2$$

$$B = \frac{f_2 + f_i}{f_1 + f_i} \qquad B^2 = \left(\frac{f_2 + f_i}{f_1 + f_i}\right)^2$$

$$C_{Gmax} = C_{max} - C_{min}$$

$$C_{Gomax} = C_{omax} - C_{omin}$$

Values of components for the *signal tuning circuit* may be calculated as follows:

$$C_T = \frac{C_{Gmax}}{A^2 - 1}\ pF$$

$$L = \frac{25330}{C_T f_2^2}\ \mu H$$

For the oscillator circuit, there are two methods of calculation; one is for arithmetical-mean tracking, and the other is for geometrical-mean tracking. Arithmetical-mean tracking is probably best if the receiver tunes a relatively narrow range of frequencies, while geometrical-mean tracking should be used if the receiver tuning range is large, such as $f_2/f_1 = 3$. Again, some preliminary calculations are needed. For arithmetical-mean tracking, calculate:

$$r = \frac{A^2}{B^2}\left(\frac{3+A}{3+B}\right)\left(\frac{1+3B}{1+3A}\right)$$

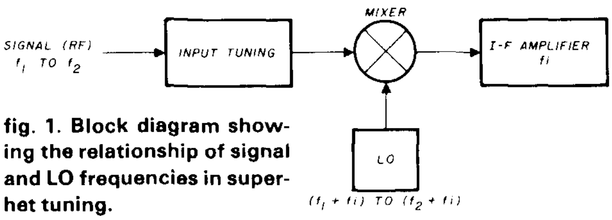

fig. 1. Block diagram showing the relationship of signal and LO frequencies in superhet tuning.

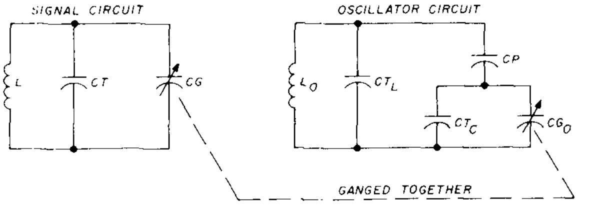

fig. 2. Component arrangement in superhet oscillator and signal tuning circuits. In some receiver designs two or more identical signal circuits may be used for preselection.

For geometrical-mean tracking, calculate:

$$r = \frac{A^2}{B^2}\left(\frac{2B + (1+B)(\sqrt{A})}{2A + (1+B)(\sqrt{A})}\right)$$

With the proper value of r determined, proceed as follows:

$$C_{Pmax} = \frac{C_{Gomax}}{r - 1}$$

$$C_{Tcmax} = \frac{C_{Gomax}}{rB^2 - 1}$$

$$C_{Pmin} = C_{Pmax} - C_{Tcmax}$$

$$C_P = (C_{Pmin} + C_{TL})\ pF$$

$$C_{Tc} = (C_{Tcmax} - C_{TL})\ pF$$

$$L_o = \frac{25330\ C_{Pmin} C_{Pmax}}{C_{Tcmax} C_P^2 (f_2 + f_i)^2}\ \mu H$$

example

A receiver is wanted which will tune from 2.5 to 3.5 MHz, and the i-f is to be 0.455 MHz. To add a little safety factor, it is decided to make the tuning range 50 kHz wider on each end.

$$f_1 = 2.45\ MHz$$

$$f_2 = 3.55\ MHz$$

$$f_i = 0.455\ MHz$$

$$A = \frac{3.55}{2.45} = 1.449$$

$$A^2 = 2.1$$

$$B = \frac{3.55 + 0.455}{2.45 + 0.455} = 1.3787$$

$$B^2 = 1.9$$

A three-gang variable capacitor is available, and each section is measured to have a range of 10 to 365 pF. Therefore

$$C_{Gmax} = 365 - 10 = 355\ pF$$

table 1. Results of calculations to prove validity of the design approach (see figs. 3 and 4).

C_G (pF)	f_{osc} (MHz)	f_{osc}-0.455 (MHz)	f_{sig} (MHz)	error (Hz)
0	4.0050	3.5500	3.5500	0
45	3.7812	3.3262	3.3257	500
90	3.5946	3.1396	3.1392	400
180	3.2994	2.8444	2.8443	100
270	3.0746	2.6196	2.6195	100
355	2.9054	2.4504	2.4497	700

The rf or signal section components may now be calculated:

$$C_T = \frac{355}{2.1-1} = 322.73\,pF$$

$$L = \frac{25330}{322.72\,(3.55)^2} = 6.228\,\mu H$$

For proper alignment, and to allow for tuning out stray circuit reactance, the inductor should be slug-tuned, and C_T should have an adjustable component. Remember also that the calculated value of C_T includes the minimum capacitance of the variable tuning capacitor. With these things in mind, the rf tuning circuit could be designed as shown in **fig. 3**. The 10-pF minimum capacitance of the tuning capacitor, plus the 43-pF setting of the trimmer, plus the 270-pF fixed capacitor add up to the calculated value for C_T of 323 pF. Distributed capacitance of the coil has been ignored, but its value can be no more than a few pF ad can be easily compensated for by slight adjustment of the trimmer during alignment.

Arithmetical-mean tracking is chosen for the oscillator circuit because of the modest tuning range.

$$r = \frac{2.1}{1.9}\left(\frac{3+1.449}{3+1.3787}\right)\left(\frac{1+3\times 1.3787}{1+3\times 1.449}\right) = 1.0787$$

$$C_{Pmax} = \frac{355}{1.0787-1} = 4510.8\,pF$$

$$C_{Tcmax} = \frac{355}{1.0787\,(1.9)-1} = 338.2\,pF$$

$$C_{Pmin} = 4510.8 - 338.2 = 4172.6\,pF$$

Ignoring the distributed capacitance of the oscillator coil gives the following:

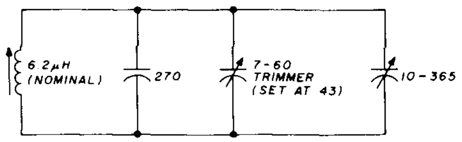

fig. 3. Design of the signal tuning circuit given in the example.

$$C_{TL} = 0$$

$$C_P = 4172.6 + 0 = 4172.6\,pF$$

$$C_{Tc} = 338.2 - 0 = 338.2\,pF$$

$$L = \frac{25330 \times 4172.6 \times 4510.8}{338.2(4172.6)^2\,(3.55+0.455)^2} = 5.0478\mu H$$

Using these calculated values, the actual oscillator tuning circuit could be set up as shown in **fig. 4.**

That takes care of the paper design, but is it correct? To find out, I used the calculated values of the capacitors and inductors, then chose several discrete values for C_G and C_{Go}, the sections of the variable capacitor, and calculated the resonant frequencies of the rf and LO circuits for each value. **Table 1** shows the results.

For perfect tracking, each oscillator frequency minus 0.455 MHz should exactly equal the corresponding signal frequency. The errors are so small

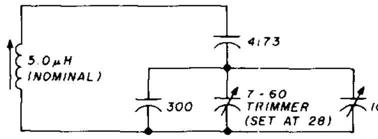

fig. 4. Design of the oscillator tuning circuit.

that they can be attributed to rounding off calculations in each step of the procedure; therefore, the overall method appears valid.

remarks

No allowances are made for effects of coupling the resonant circuits to other circuit components, which will certainly have some impact. If the range of the adjustable components does not allow proper alignment and tracking, then the values of some components may have to be slightly changed. The small distributed capacitance (C_{TL}) of the oscillator coil, which was ignored, causes slight errors in the calculated values of the oscillator padder, C_P, and trimmer, C_{Tc}, but C_P is so large that the error is negligible there, and C_{Tc} may be adjusted to compensate for the error.

Information on correct superheterodyne alignment techniques is available to amateurs elsewhere,[2] but the design equations presented here have, in my opinion, been too long absent from contemporary amateur literature.

references

1. F. Langford-Smith, *Radiotron Designer's Handbook*, 4th edition, 1952, page 1002.
2. William I. Orr, W6SAI, *Radio Handbook*, 20th edition, 1975, page 19.31.

ham radio

More Details? CHECK—OFF Page 142

CW signal processor

A simple means for eliminating QRM from CW signals by using a phase-locked loop and audio oscillator

Almost from the very beginning of ham radio, there has been interference. With varying degrees of success, numerous devices have been designed to combat this problem. Modern technology has provided us with such things as narrow-bandwidth crystal filters, active audio filters, *Q*-multipliers, and acoustically resonant transducers, to name just a few.

There is another method of providing interference rejection for the CW operator, though it has been largely ignored except by a scant few. This involves the use of narrow-bandwidth, integrated-circuit tone decoders, or as shown in this article, the LM567.

The LM567 is a phase-locked-loop tone decoder which can be made to respond to a tone anywhere from less than 1 Hz to approximately 500 kHz. For my use, the range is adjustable from roughly 500 Hz to 1100 Hz. The bandwidth has been set to about 50 Hz either side of the center frequency. In other words, if the LM567 is set to a center frequency of 750 Hz, it will respond to any signal from 700 Hz to 800 Hz and ignore virtually all others.

circuit description

A CW signal from the phone jack of a receiver is fed to the 8-ohm winding of T1 (see **fig. 1**). This transformer presents a low-impedance termination for the receiver audio stage, as well as providing a voltage step-up for the input of the LM567. The two 1N34A germanium diodes across the secondary limit the audio voltage to near the optimum value for the tone decoder. The three resistors and the capacitor connected to pins five and six determine the frequency range over which tones can be decoded. When a tone of the proper frequency is present at the input terminals, the output (pin 8) goes to ground and causes the LED to light.

The waveform at the output of the LM567 is sometimes a little ragged, and, for that reason, it is fed through one half of a 7413 Schmitt trigger. This stage transforms the output waveform to a square wave with very fast rise and fall times and also performs the inversion necessary for the following stage.

It wasn't until a prototype was constructed that a problem became known. The output of the tone decoder stays low for a few milliseconds after the input signal stops. The net result is to increase the "weight" of the keyed signal. That is, it decreases the spacing between code elements. To counteract this problem, a 1000-ohm resistor was connected between the output of the LM567 and the input of the 7413. Also, there is a 3.3-μF capacitor from the input of the 7413 to ground. This combination delays

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the switching time of the 7413 after the tone decoder goes low. The end result is to restore normal weight to the keyed signal. The values specified were experimentally derived and may be adjusted to suit individual tastes. The third stage is the familiar NE555 timer, wired as a keyed audio oscillator.

In operation, a CW signal is tuned in on the receiver and the frequency control of the CW PROCESSOR is varied until the LED begins to blink in unison with the incoming signal. This indicates that the LM567 is tuned to the proper frequency and is decoding the CW being presented to it. Activating S1 will replace the live audio with the tone generated by the 555.

It takes approximately 10 to 15 millivolts of audio from the receiver to activate the tone decoder. This corresponds to a rather weak signal (S3 to S4 would be a fair guess) in most receivers. Obviously then, the CW PROCESSOR is quite sensitive and does not take a "block-buster" signal to make it work.

From time to time, you will encounter signals that shift frequency, fade into the noise, or become obliterated by stronger adjacent signals. As these situations occur, the CW PROCESSOR will stop responding to the signal, with the resultant loss of audio. The circuit shown in **fig. 2** was added to automatically switch the output from the receiver back to the headphones, after an adjustable delay.

Once again, the NE555 timer is pressed into service, this time as a monostable multivibrator. The addition of a single PNP transistor transforms the circuit into a negative recovery monostable.[1] If this circuit is incorporated into the CW PROCESSOR, S1 will be replaced by the contacts on K1.

It should be pointed out that if a relay with a 5- or 6-volt coil is unavailable, it is quite permissible to substitute one with a 12-volt coil and operate the second NE555 from the 12-volt supply line. The circuit performance is identical in either case. Of course, the additional NE555 is not absolutely necessary to the performance of the CW PROCESSOR, but it does add a considerable amount of operator convenience. Whether it is included or not is entirely up to the individual builder.

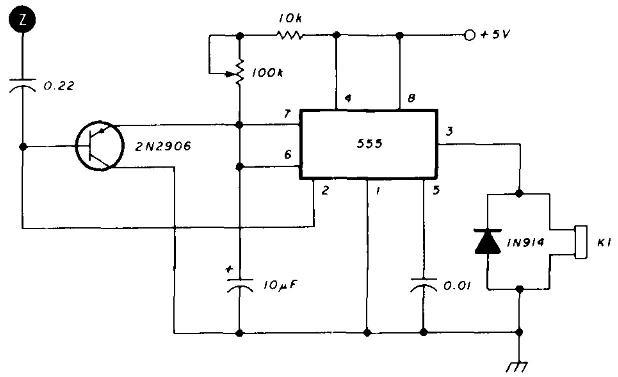


fig. 2. For periods when you experience loss of signal, this circuit will automatically switch back to live receiver audio after a suitable delay. If a relay with a 5-volt coil is not available, the circuit can also be powered from +12 volts.

construction

Because everything is operating at audio frequencies, layout and construction are definitely not critical. My version was constructed on a printed circuit board with hole spacings adjusted to suit the size of components on hand. Alternatively, perfboard and hardwired connections could be used with equally reliable results.

On the subject of parts and pieces, note that all of the necessary parts to build your own version of the CW PROCESSOR are listed in the Radio Shack catalog. Reasonable amounts of latitude may be taken

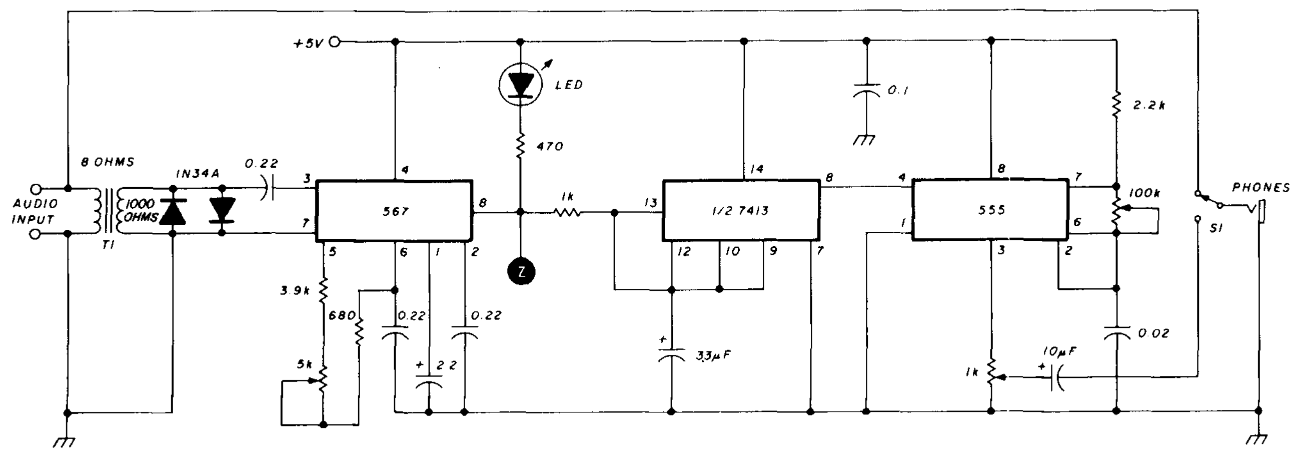
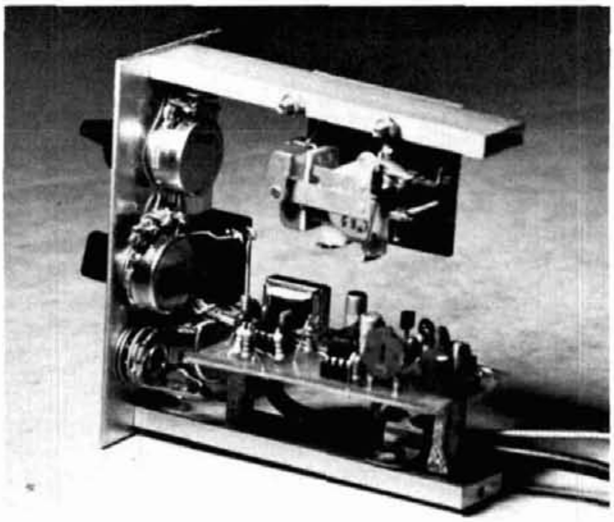


fig. 1. Schematic diagram of the signal processor. The 567 is a phase-locked loop which is configured to respond to tones from 500 to 1100 Hz. The Schmitt trigger reduces the weighting effect caused by the output of the PLL remaining low after removal of the audio signal. Since the processor requires a +5 volts, a simple 3-terminal regulator can be used to power the unit from a 12-volt line.



CW signal processor with the case removed, showing layout and construction. Relay is mounted on separate circuit board installed on top chassis rail; other circuit board is mounted on standoffs on bottom rail.

with regard to parts values since few are critical. However the capacitors associated with the LM567 and LM340 should be low-leakage types; tantalums are recommended as the best choice in this case.

The enclosure was garnered from the local surplus emporium and is pretty much a one-of-a-kind item. Any box large enough to accommodate the parts will be suitable. In the early developmental stages, the CW PROCESSOR was operated with no enclosure at all and no problems were encountered. This would suggest the possibility of using a nonmetallic box instead of the more common aluminum cabinet.

summary

The CW PROCESSOR is not a cure-all for all QRM problems; it does have its limitations. For example, the 100-Hz bandwidth talked about earlier is valid only for signals not exceeding approximately 300 millivolts at the input of the LM567. As is sometimes the case, you will be trying to copy an S5 signal with the CW PROCESSOR and an S9 signal only 200 Hz away will disrupt reception. Also, if you are trying to copy a heavily weighted CW signal, the CW PROCESSOR will aggravate the situation unless an absolute minimum of signal is presented to the input.

All things considered, however, using the CW PROCESSOR has been pure joy. Operator fatigue is greatly reduced by not having to listen to all of the garbage normally associated with ham band signals. It is amazing how well this device snatches a barely readable signal out of the noise and transforms it into the kilowatt-next-door category. Build one for yourself and see what a difference it makes at your station.

reference

1. Don Lancaster, *TTL Cookbook*, Howard W. Sams, Indianapolis, 1974.

ham radio

birdie suppression in the Swan 160X

The Swan 160X is a 400-watt PEP input transceiver for 160 meters. Unfortunately, it is no longer made. During its lifetime, there was little interest in the 160-meter band. Now, that interest is growing, the 160X is a prized piece of equipment for the "top band" enthusiast. A few 160Xs are available second hand, but they are quickly snapped up and do not stay on the dealer's shelves for any length of time.

A minor problem with the 160X is a birdie or crossover product which falls in the passband of the receiver. It can be heard as a carrier, or heterodyne, at about 1834 kHz. Though not particularly loud, it can be very annoying when you're looking for a weak DX signal.

An investigation of the mixing technique in the 160X shows that the spurious signal is a result of unwanted mixer products from the VFO and the carrier oscillator, Q3. At spurious frequency of 1834 kHz, the carrier oscillator is at 5.500 MHz and the VFO is at 7.333 MHz. The third-order product of these two frequencies is:

$$Birdie = 3f_2 - 2f_1$$

where,

$$f_1 = 7.333\ MHz$$

$$f_2 = 5.500\ MHz$$

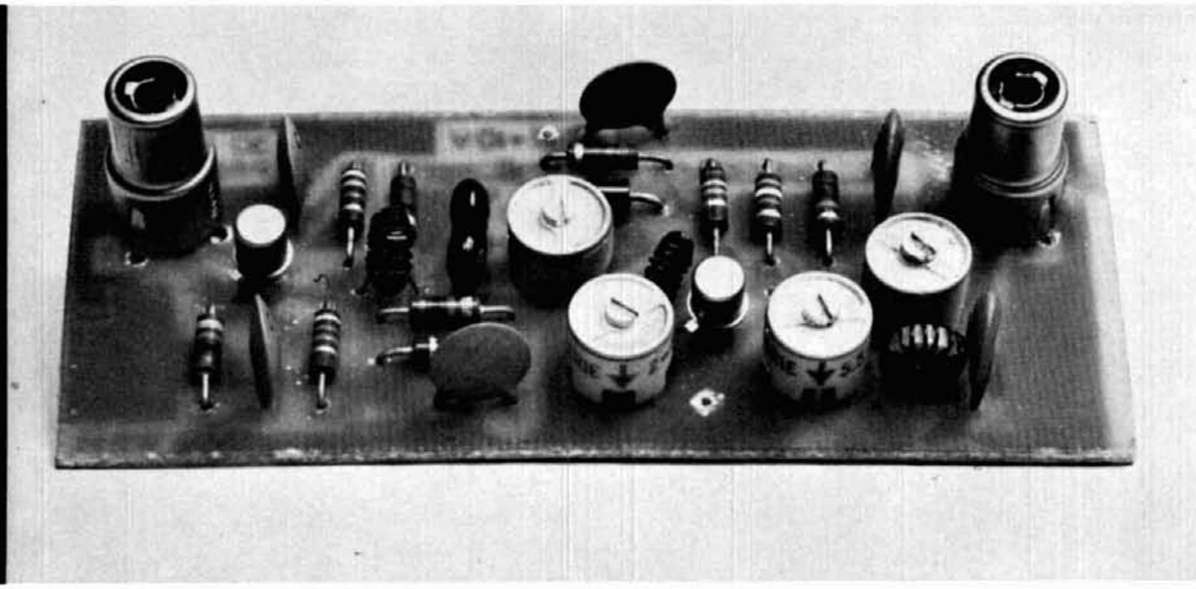
or, $$16.500 - 14.666 = 1.834\ MHz$$

The birdie may be reduced to an amplitude by placing a trap tuned to 14.666 MHz in the output lead of the VFO. This is easy to accomplish since the output signal from the VFO appears at the accessory socket (J6) located on the rear apron of the 160X. It is merely necessary to break the lead in the plug and insert a small trap as shown in **fig. 1**. The trap can be made up of very small components and mounted directly to the pins of the plug, which should always be in place when an auxiliary VFO is not used. (Since, to my knowledge, an auxiliary VFO was never built for the 160X, this is a moot point!)

If a compression-type capacitor is used, the trap is easily adjusted by tuning the transceiver to 1834 kHz and adjusting the capacitor for minimum birdie response in the receiver.

Bill Orr, W6SAI

More Details? CHECK—OFF Page 142



low-noise 30-MHz preamplifier

A low-noise preamplifier which may be used to improve 10-meter receiver sensitivity, improve OSCAR communications, or extend Gunnplexer range

The apparent sensitivity of many communications receivers seems to fall off above about 25 MHz because of the lower levels of external galactic and external noise at these frequencies.[1] When the 10-meter band is wide open, this isn't particularly noticeable, but when propagation conditions are marginal additional sensitivity makes a big difference in DX performance.

A receiver which has an adequate noise figure on 20 and 15 meters may be marginal on ten; also, front-end circuits which have been optimized for the lower amateur bands don't always work as well as they should at 28 MHz. This is especially true with vacuum-tube rf amplifiers. Since 10 meters is open perhaps three years during the 11-year sunspot cycle, and then for only a few hours each day, it is understandable why the designers don't pay more attention to 10-meter performance.

If you operate on the vhf-uhf bands and use your receiver as a tunable i-f, noise figure is extremely important because it affects the noise performance of your vhf/uhf converter. Satellite communications can also be improved by better receiver sensitivity, and if you operate on 10 GHz with a Gunnplexer, you

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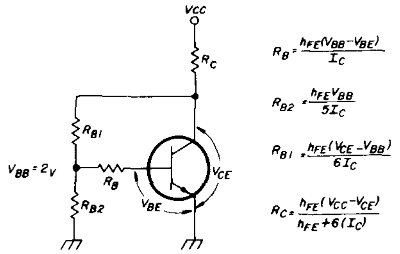

fig. 2. Dc bias circuit for vhf/uhf applications stabilizes collector current with voltage feedback through resistor R_B and maintains constant base current with R_{B1} and R_{B2}. A design example is given in the text.

can double your effective range by lowering your system noise figure by 6 dB.

30-MHz preamp

The 30-MHz i-f preamplifier shown in **fig. 1** is based on a design by engineers at Microwave Associates using low-noise npn silicon planer transistors.* These transistors exhibit excellent noise figure *vs* current characteristics, which results in extremely low noise figure and wide dynamic range. The circuit provides 19 dB gain with a noise figure of about 1.1 dB; compression of 1 dB occurs at an output of −7 dBm. The 3-dB bandwidth of the preamplifier is 10 MHz, and the input is designed to match the 200-ohm source impedance of the Gunnplexer mixer diode. Circuits for matching the preamp to 50 ohms are discussed later in this article.

The noise figure of the Schottky mixer diode in the Gunnplexer is specified at 12 dB maximum, but many units are better than this. With careful design, proper impedance matching, and the use of an i-f preamplifier with a 1.0 to 1.5 dB noise figure, some users have reported system noise figures well below 10 dB. This represents a significant increase in reliable communications range.

*Microwave Associates transistors are available from G. R. Whitehouse, Newbury Drive, Amherst, New Hampshire 03031.

bias circuit design

One factor which is often overlooked in vhf circuit design is the dc bias network. At low frequencies an emitter resistor is often used to provide negative current feedback for dc stability. In low-noise vhf applications, however (and this includes 28 MHz), the emitter bypass capacitor which is an efficient rf bypass at the design frequency often introduces low frequency instability. Furthermore, any series emitter impedance, no matter how small, results in a degradation of noise figure and gain. Therefore, vhf circuits which are designed for lowest noise or maximum gain require that the emitter lead be grounded as close as possible to the transistor package to keep emitter series feedback to an absolute minimum.

The transistor variable which has the most effect on dc stability is collector current. If you study the transistor's parameters, you'll find that gain and noise figure are the most sensitive to changes in bias, and both are stronger functions of collector current than of collector-emitter voltage, V_{CE}. Therefore, the ultimate goal in dc bias design is to stabilize collector current. The temperature-sensitive parameters which affect collector current are the internal base-emitter voltage, V_{BE}; the dc current gain, h_{FE}; and the reverse collector current, I_{CBO}. Normally I_{CBO} is expected to double for each 10°C temperature rise, but because of surface currents in uhf and microwave transistors, I_{CBO} increases at a rate much less than this and can usually be neglected in vhf bias design.

Fig. 2 shows a dc bias circuit which stabilizes V_{BE} and h_{FE} by using voltage feedback through R_B

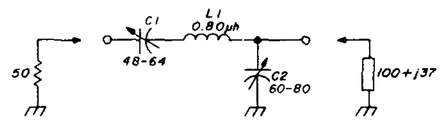

fig. 3. Suggested circuit for matching the low-noise preamplifier stage to 50 ohms.

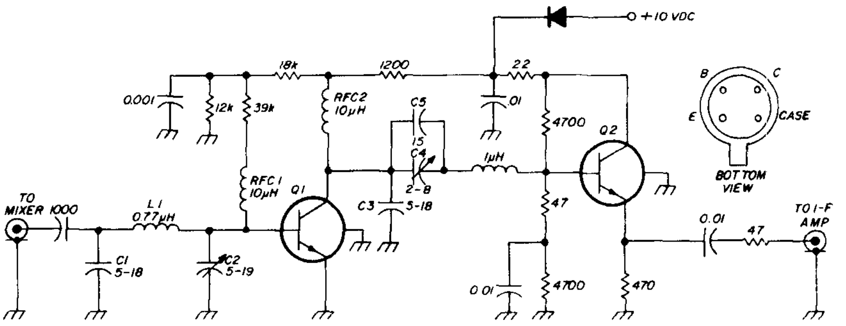

fig. 1. Low-noise preamplifier has a noise figure of 1.1 dB at 30 MHz and 3 dB bandwidth of 10 MHz. Gain is 19 dB. Total current drain with a +10 volt supply is 13 mA. All resistors are 1/4 watt carbon; bypass capacitors are 50-volt ceramics.

and a constant base current source from resistors R_{B1} and R_{B2}.[2] Not shown are the rf chokes which must be placed in series with the base and collector (RFC1 and RFC2 in **fig. 1**).

The design equations for this bias circuit are given in **fig. 2**. First determine the available supply voltage V_{CC}, select the desired transistor operating point (V_{CE} and I_C), and check the transistor data sheet for dc forward gain h_{FE}. If h_{FE} data is unavailable, assume $h_{FE} = 50$; this is a fair assumption for many vhf/uhf transistors. To ensure a constant base current source, the voltage V_{BB} is set at approximately three times the base-emitter voltage, V_{BE}, or about 2 volts for silicon transistors ($V_{BE} = 0.7$ *volt*). The current through R_{B2} is set at five times the base current I_B. Since $I_B = I_C/h_{FE}$, the current through R_{B2} is $5I_C/h_{FE}$. The current flowing through R_{B1} is the

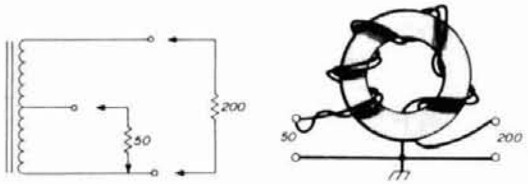

fig. 4. Schematic and construction of a 4:1 rf transformer for matching 200 ohms to 50 ohms. The transformer consists of 18 turns slightly twisted pair number 28 (0.3 mm) enamelled wire on a T-50-6 toroid form.

sum of the current through R_{B2} plus base current or $6I_C/h_{FE}$.

The noise-figure curve at 30 MHz for the Microwave Associates 42001-509 transistor shows a rather broad minimum centered around $I_C = 3$ *mA*; h_{FE} is about 90. With a 10-volt dc power supply, V_{CE} is selected to be 6 volts. Using the design equations of **fig. 2** yields the following bias resistor values: $R_B = 39k$, $R_{B2} = 12k$; $R_{B1} = 20k$; and $R_C = 1250$ ohms. The

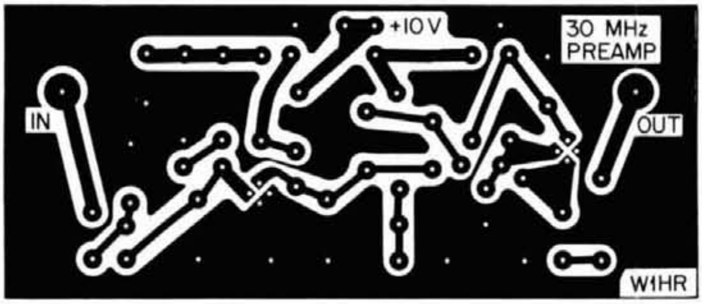

fig. 5. Full-size printed circuit layout for the low-noise preamplifier.

30-MHz preamplifier in **fig. 1** uses the nearest standard resistance values.

In the output emitter follower stage dc stabilization is provided by current feedback produced by the 470-ohm emitter resistor; the input impedance of this stage is approximately 50 ohms. The emitter follower is used to drive a 50-ohm coaxial cable to the first i-f stage or front end rf amplifier. If the preamplifier is located very close to the 28-30 MHz rf stage, the emitter follower may be omitted.

input matching

Another important consideration in low-noise amplifier circuits is the design of the input matching circuit. For the 42001-509 transistor the input impedance for optimum noise figure is 100 + j37 ohms at 30 MHz. The input pi network (C1, L1, C2 in **fig. 1**) transforms this to the 200-ohm source impedance of the Gunnplexer mixer diode. The output of the first stage is matched to the approximately 50-ohm input of Q2 with C3, C4, C5, and L2.

If you wish to use this preamplifier in a 50-ohm system you can either modify the input matching circuit or use a 4:1 rf transformer. A suggested 50-ohm

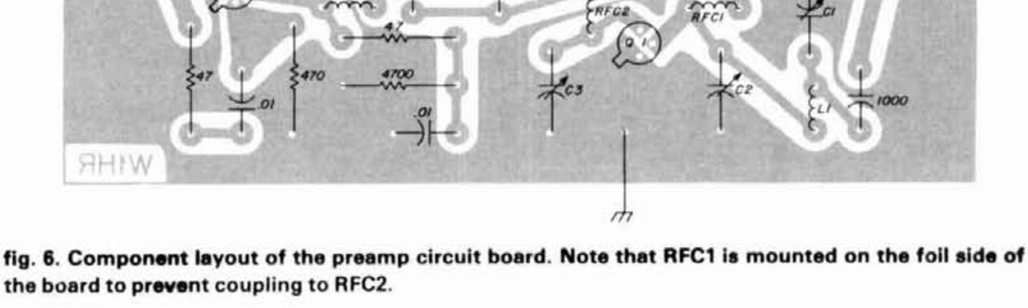

fig. 6. Component layout of the preamp circuit board. Note that RFC1 is mounted on the foil side of the board to prevent coupling to RFC2.

input circuit is illustrated in **fig. 3**. Construction of a simple 4:1 rf transformer which will match 200 ohms to 50 ohms is shown in **fig. 4**.

construction and test

Fig. 5 shows a full-size, printed-circuit layout for the low-noise 30-MHz preamplifier; the component placement is shown in **fig. 6**. Note that the rf choke in the base circuit of Q1 (RFC1) is mounted on the foil side of the board; this is to prevent unwanted coupling to RFC2, which is located nearby. When winding the toroid coils, be sure to spread the windings evenly over the circumference of the form.

With slight modification the circuit board will accommodate the 50-ohm matching circuit of **fig. 3**. L1 and C2 are soldered to the same circuit pads as L1 and C2 in the 200-ohm matching circuit. C1 replaces the 1000-pF blocking capacitor; however, it may be necessary to drill new holes because of the wider spacing of the tabs on the variable capacitor.

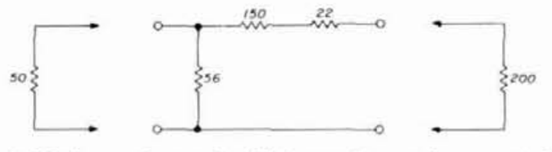

fig. 7. Minimum loss pad which may be used to match 200 ohms to 50 ohms. Pad loss of 11.5 dB must be considered when making noise or gain measurements.

Since most rf signal generators and noise-figure meters are designed for a 50-ohm system, and the preamplifier is designed to match 200 ohms, you must take a 4:1 impedance transformation at the input when tuning the preamp. You can use the 4:1 rf transformer if you wish, or the minimum loss pad* shown in **fig. 7**. This pad has approximately 11.5 dB loss, which must be considered when making gain or noise measurements.

For best operation the preamplifier should be adjusted for minimum noise figure, but this is not possible if you don't have access to noise-measuring equipment. Tuning the preamplifier for maximum gain will degrade noise figure slightly, but noise performance will still be better than that available with most 28-MHz receivers or 30-MHz i-f strips.

*A minimum loss pad is a resistance pad which will provide an impedance match between unequal terminations with the smallest possible attenuation.

references

1. James Fisk, W1DTY, "Receiver Noise Figure, Sensitivity, and Dynamic Range — What the Numbers Mean," *ham radio*, October, 1975, page 8.
2. Kenneth Richter, "Design DC Stability Into Your Transistor Circuits," *Microwaves*, December, 1973, page 40.

ham radio

clean local-oscillator chain for 1296 MHz

An easy-to-build LO for 1296 MHz which can be optimized without a spectrum analyzer. Outputs are also available for 220 and 432 MHz

Development of a local-oscillator chain for 1296 MHz which can be tuned up without a spectrum analyzer, yet has an acceptably clean spectrum, has been a long-time goal. I have built several LO chains which required the use of a spectrum analyzer and several hours of trimming to tune; when they quit in the midst of a contest, there was no recourse. This article describes a 1296-MHz LO chain which can be tuned up in a few minutes with minimal test equipment, including a tripler which needs no tuning. The spectrum analyzer photographs were taken after tuning was completed; it was not used during the tuneup procedure.

What is an "acceptably clean spectrum?" Very simply, it is one which produces no birdies in the operating band. More quantitatively, the following criteria are arbitrarily defined:

1. No spurious (not harmonically related) outputs
2. Undesired harmonics of oscillator suppressed more than 40 dB
3. Undesired harmonics of oscillator well separated (spacing more than 5% of output frequency)
4. No harmonics near the i-f band
5. Low noise content

Now examine **fig. 1A**, the local-oscillator spectrum of a typical 432-MHz converter with a fairly low frequency oscillator, followed by two single-tuned transistor triplers. This converter has enough birdies so the band never sounds dead!

The causes of the poor spectrum of **fig. 1A** are insufficient selectively, excessive multiplication factor, and inefficient multipliers. Increasing the oscillator frequency spaces the harmonics and eases the selectivity, or filtering, problem. However, a single transistor is still an inefficient multiplier; it would much rather amplify than multiply, so the output always has a strong fundamental component. (A tripler also has a strong second harmonic component). One solution is the use of idlers, but they make tuning very critical and usually add a tendency to parametric oscillations. Diode multipliers have the same problems, but with added loss (transistor multipliers often have gain).

A more effective solution is the use of natural multiplying circuits. A push-push doubler was described several years ago.[1,2] This basic circuit was incorporated into both doubler stages of a 432-MHz local-oscillator chain; the resultant spectrum shown in **fig. 1B**. This system has worked beautifully for more than two years.

Recently, a push-pull tripler was added to the out-

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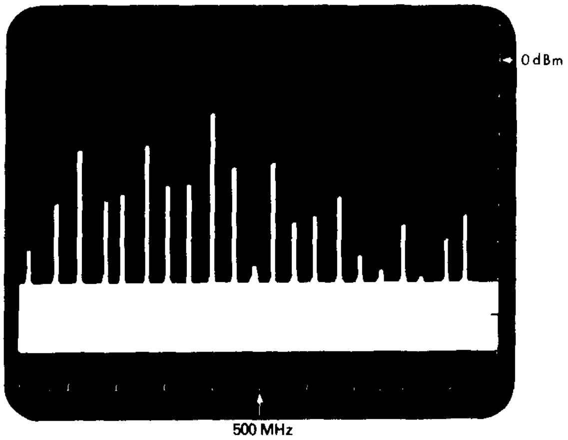
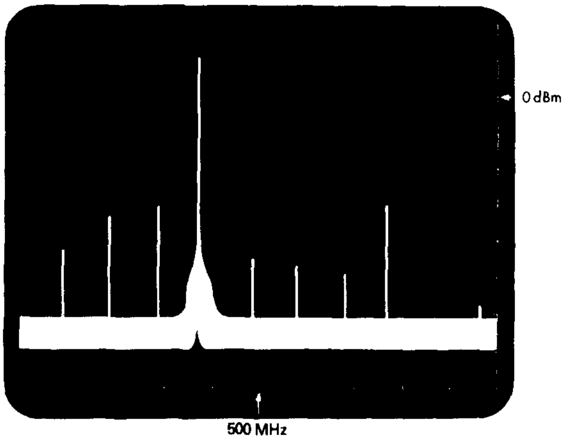
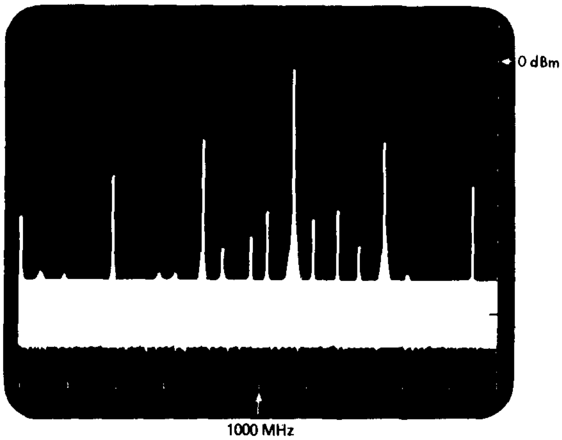
fig. 1. Spectrum analyzer display of the local oscillator chain for a typical 432-MHz converter, A, showing the large number of undesired spurs. Other displays show the output spectrum of the local-oscillator chain described in this article. (Measurements made with Ailtech 727 spectrum analyzer at 10 dB per division.)



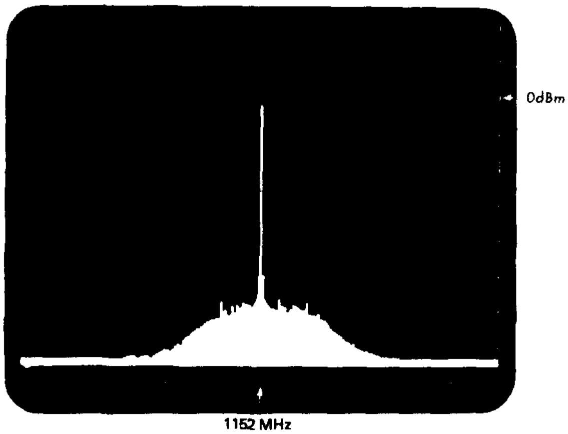
A. Local-oscillator spectrum of a typical 432-MHz converter, after tuning for cleanest output with a spectrum analyzer. Horizontal scale: 100 MHz per division.



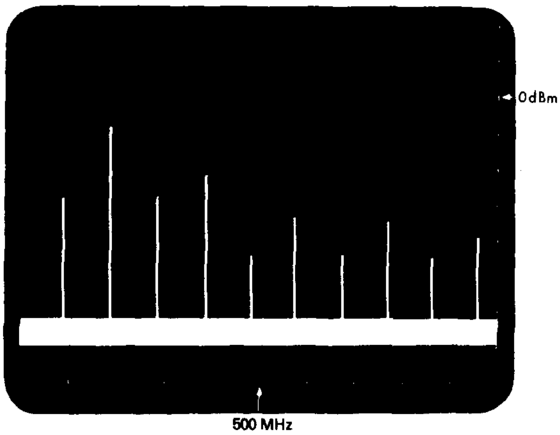
B. Output from the low-frequency multiplier chain described in this article. Output at 384 MHz is approximately 20 mW. Horizontal scale: 100 MHz per division.



C. 1152-MHz output from the frequency tripler for use as a local oscillator on 1296 MHz. Output is approximately 0.5 mW. Horizontal scale: 200 MHz per division.

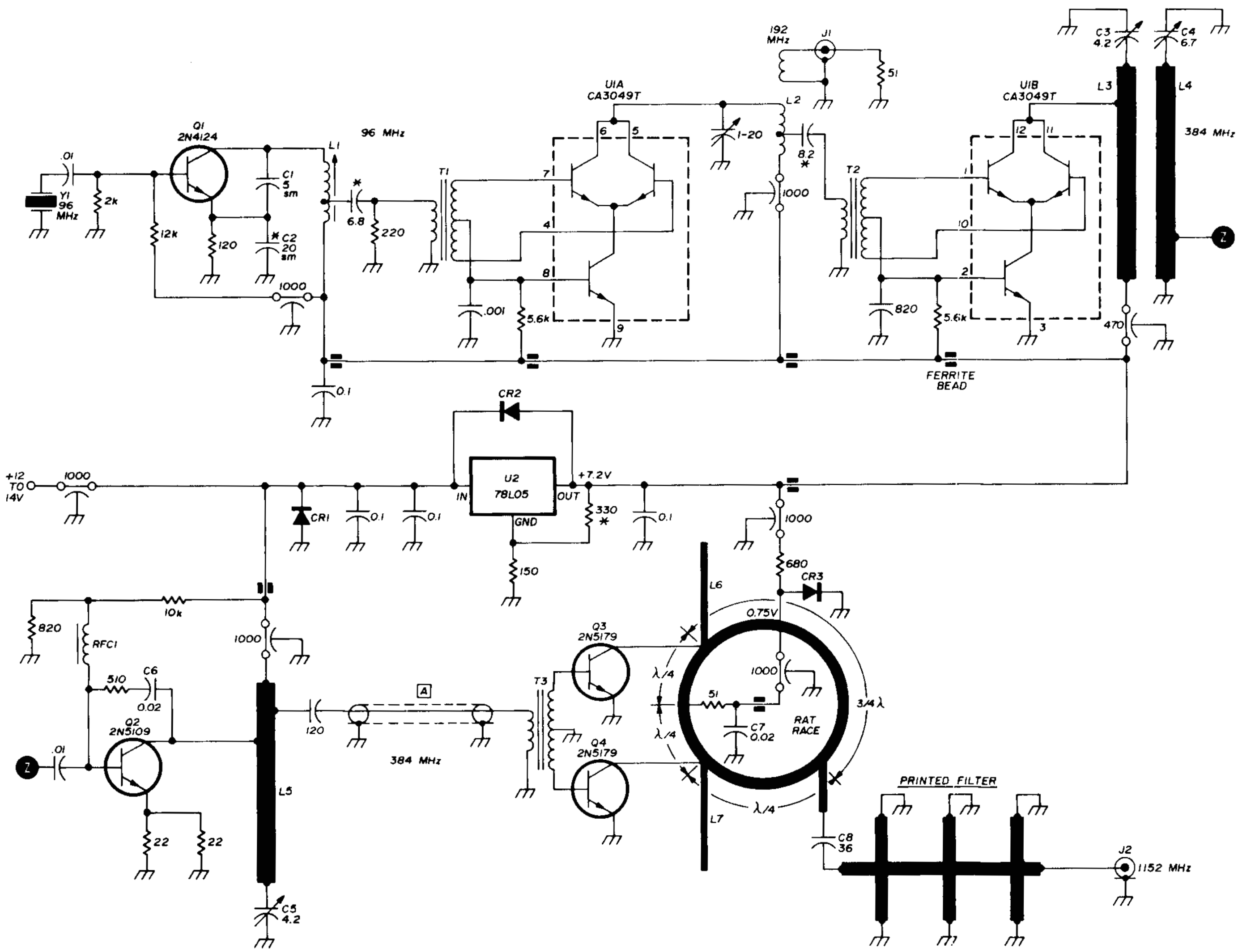


D. Display of the 1152-MHz output showing the noise spectrum. Horizontal scale: 1 MHz per division. Contribution from spectrum analyzer is probably significant.



E. Spectrum of output available from J1 is easily filtered for use as a local oscillator for 220-MHz equipment. Horizontal scale: 100 MHz per division.

put to build a 1296-MHz local oscillator. The push-pull tripler had been a stumbling block because of the requirement that the outputs, as well as the inputs, be 180° out of phase; the push-push doubler requires phase reversal only at the input, which is easily provided by a trifilar wound transformer or a balun. If an untuned transformer was used at the output, it would be a broadband amplifier rather than



C3, C5	1.2-4.2 pF air variable (E.F. Johnson 189-501-5)
C4	1.3-6.7 pF air variable (E. F. Johnson 189-502-5)
C6, C7	0.02 μF (chip capacitor or very short leads)
C8	36 pF (chip capacitor or very short leads)
CR1, CR2, CR3	1N4148 or 1N914 silicon diode
L1	6 turns no. 24 (0.5mm) Formvar close spaced on 3/16" (5mm) coil form with green slug; tap at 1½ turns
L2	4 turns no. 18 (1mm) tinned wire, 3/16" (5mm) ID, 5/8" (16mm) long; tap at 1¾ turns (optional link: 1¾ turns no. 22 (0.6mm) insulated wire, 3/16" (5mm), installed near hot end of L2)
L3	40 nH printed inductance
L4	43 nH printed inductance
L5	60 nH printed inductance
L6, L7	printed transmission lines, $\lambda/4$ long at 768 MHz, Z_0 chosen to provide correct tuning inductance at 1152 MHz
Q1	2N4124
Q2	2N5109, MSC 80099, or equivalent (2N3866 usable with lower gain)
Q3, Q4	2N5179 (case floating — case cut off flush with base)
RFC	ferrite bead with 2 turns no. 30 (0.25mm) Formvar
T1, T2	4 turns no. 30 (0.25mm) trifilar wound on large ferrite bead, μ=900 (Ferronics 11-090-J; suggested replacements: Amidon FT-25, mix 43 or Amidon 801 ferrite bead, mix 43)
T3	3 turns no. 30 (0.25mm) trifilar wound on ferrite bead, μ= 125 (Ferronics 21-030-K; suggested replacements: Amidon FT-25, mix 61 or Amidon 101 ferrite bead)
U1	RCA CA3049T
U2	78L05
Y1	96 MHz third-overtone crystal

fig. 2. Schematic diagram of the 1152-MHz local-oscillator chain. Outputs are available for use on 432 MHz (cable A) and 220 MHz (J1). Cable A is short piece of semi-rigid coax. Components marked with an asterisk may be critical or may need adjustment for optimum performance. Printed-circuit boards are shown in figs. 3 and 4.

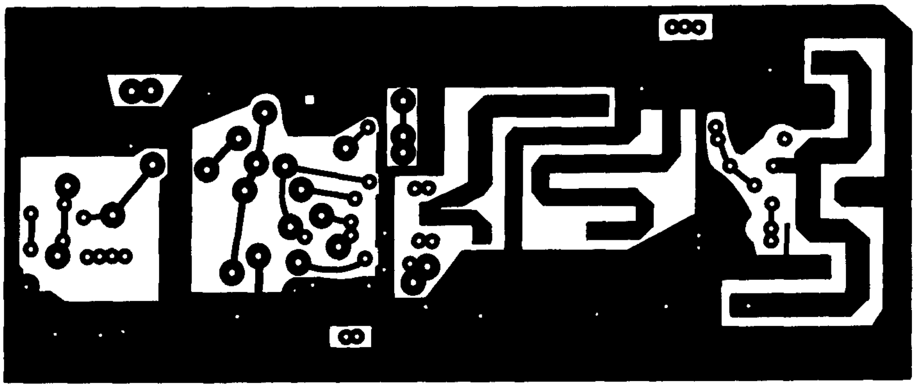


fig. 3. Printed-circuit layout for the single-sided low frequency board (output at 384 MHz). Component placement is shown in fig. 6. Material is single-clad 1.5 mm (1/16") G10 circuit board.

a tripler. The brainstorm which made this work was to use a *rat-race* coupler at the output. As used in this circuit, two input signals which are 180° out-of-phase at the resonant frequency combine at the output port, while other phases and frequencies combine in the terminating resistor. The rat-race is not critical so *no* tuning is required.

The frequency tripler is followed by a printed-circuit filter made up of quarter-wavelength stubs; the output spectrum is shown in **fig. 1C**. For detailed descriptions of the rat-race and printed-circuit filter, see Howe's excellent book, *Stripline Circuit Design.*[3] **Fig. 1D** shows the noise spectrum of the 1152-MHz output — none is evident down to the noise floor of the analyzer (on this non-optimum range for this measurement); **fig. 1E** is the output of the first doubler, at J1, for use as a 220 MHz local oscillator, or for connection to a frequency counter.

The major objection to push-push doublers and push-pull triplers is that two transistors are required. However, prices for usable vhf/uhf transistors have dropped to the point that they are available in the 15 cents to $1 range. As an alternative, several matched transistors on the same chips can be bought as an integrated circuit. The single CA3049T used for both doublers costs $1.13 from a local dealer. While the prices of semiconductors have been steadily decreasing, prices for capacitors and coil forms, needed for idlers and filters in conventional multipliers, have increased.

Construction is straightforward and requires no machining; all frequency-determining elements

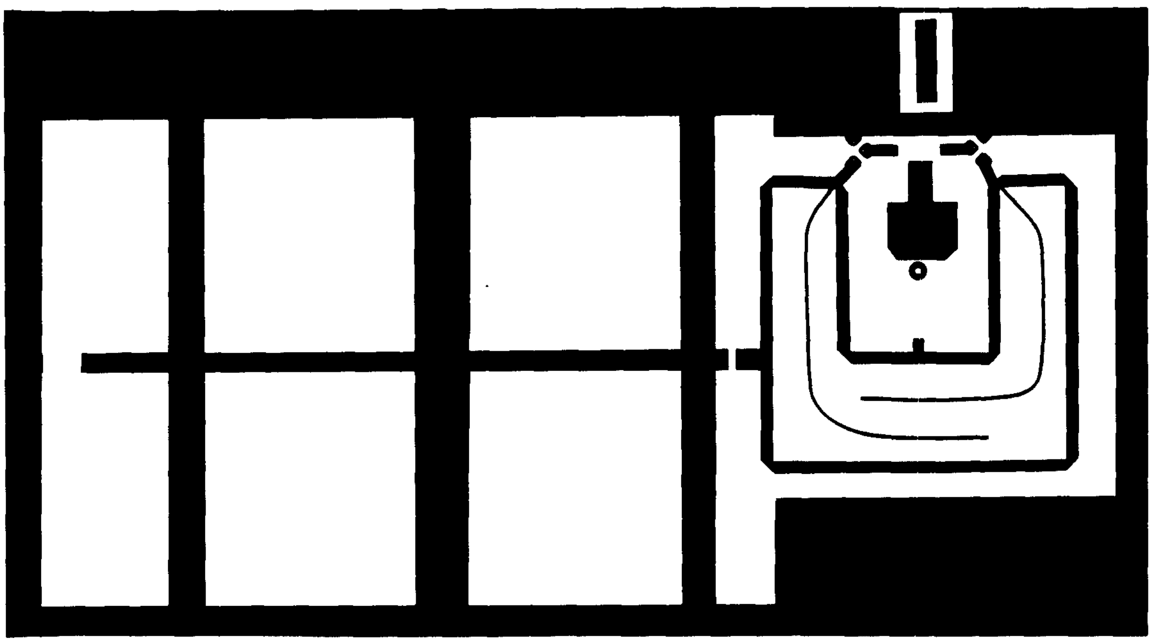


fig. 4. Active side of the double-sided 1152-MHz tripler/filter circuit board. Unetched reverse side serves as the ground plane. Material is 1.5 mm (1/16") double-clad G10 circuit board.

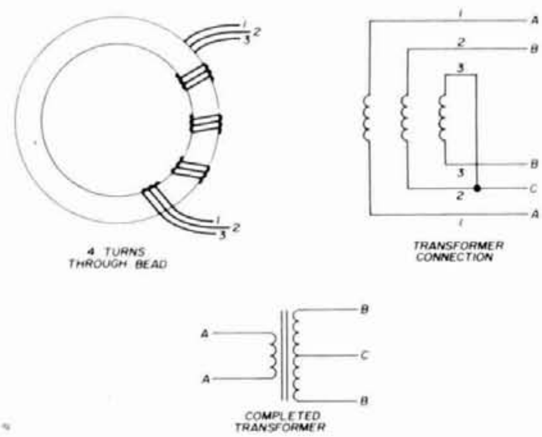
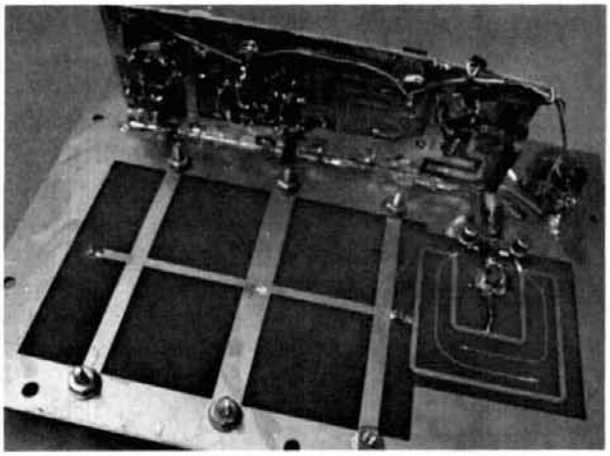


fig. 5. Construction of transformers T1 and T2. Turns are counted by the number of times the wire passes through the center. Wire is no. 30 to no. 34 (0.25-0.16mm) Formvar coated. T3 is similar except it consists of three turns.

above grid-dipper range are printed. Printed-circuit layouts are shown in **fig. 3** and **4**. Component placement may be seen in the photographs. Transistors Q3 and Q4 have their leads bent outward along the printed lines about 0.1 inch (2.5mm) below the case (the fourth lead, case, is cut off flush so the case is not connected) as shown in the close-up photograph.

The trickiest part in building the LO chain is winding transformers T1, T2, and T3. T1 and T2 consist of four trifilar turns on a large ferrite bead such as a Ferronics 11-090-J ($\mu = 900$) as shown in **fig. 5**. All windings are wound at the same time for a total of 12 turns. The four turns are counted from the inside — *not* the outside. Also shown in **fig. 5** are the transformer connections and how they are installed in the circuit. Construction of transformer T3 is similar except it consists of 3 trifilar turns on a Ferronics 21-030-K ferrite bead ($\mu = 125$). Transformer T2 is installed on the copper foil side of the board to minimize lead length. Cable **A** in **fig. 2** is not installed yet; a temporary jack is installed instead.

Tuneup is also straightforward. Monitoring the power at J1, tune L1 and the trimmer capacitor across L2 for maximum output. Check the frequency with a frequency counter, wavemeter, or grid-dipper. It may be necessary to reduce the value of C2 for maximum output with different crystals. Roughly 0.1 milliwatt is available from J1 with the small link.



Stripline filter for the low-noise amplifier. The bolts through the board ensure an adequate ground connection at the end of the quarter-wavelength filters.

Next, terminate J1 with 50 ohms and monitor the output at L5 through the 120 pF capacitor (cable **A** not yet installed). Adjust C3, C4, C5, and L2 trimmer for maximum output (approximately 20 mW). If convenient, check frequency — if not, note capacitor

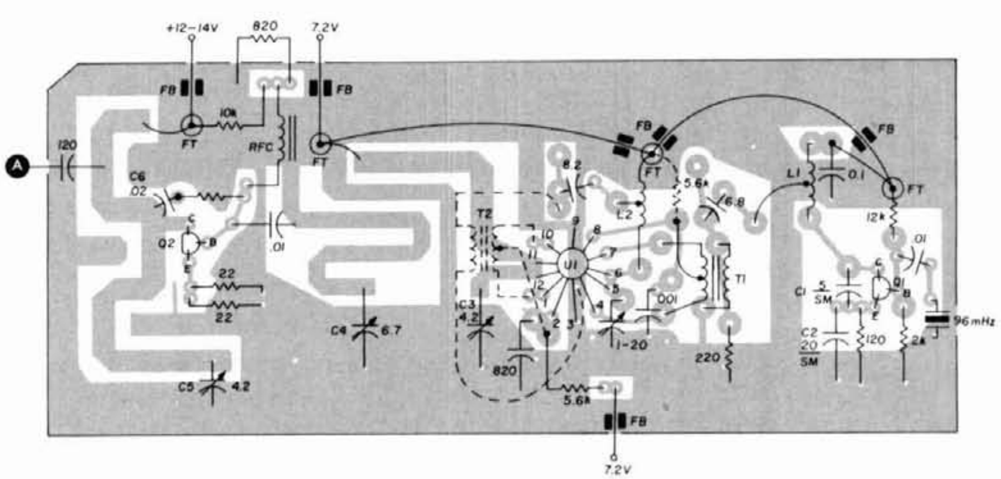


fig. 6. Component placement for the low-frequency multiplier board. Foil side is shown in fig. 3.

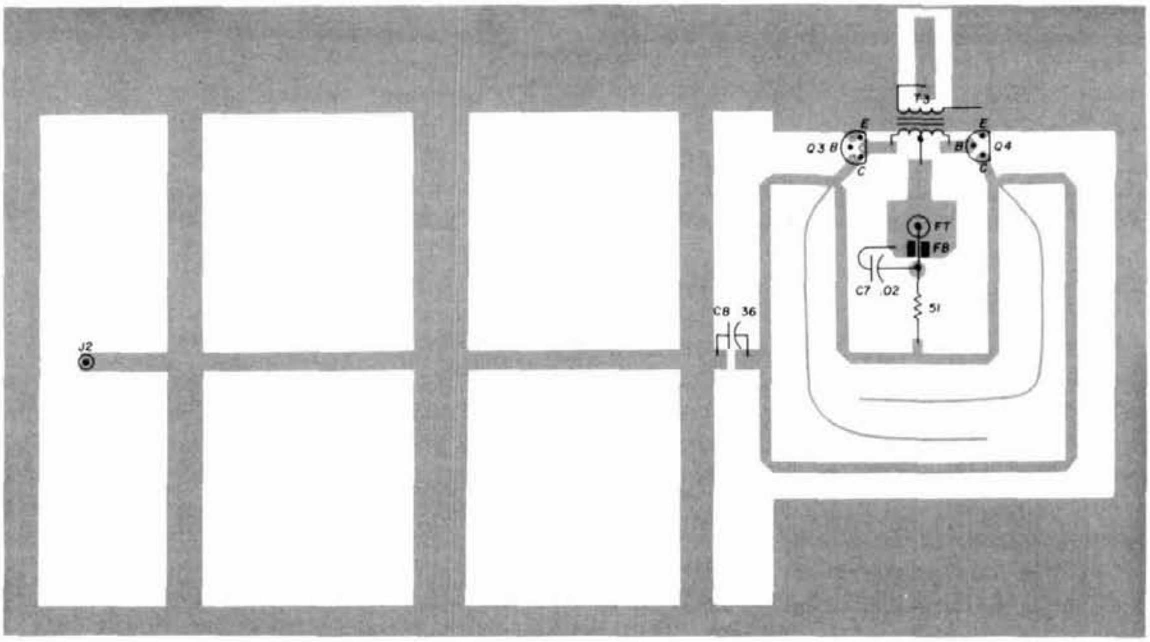


fig. 7. Component placement on the 1152-MHz tripler/filter board. Foil layout is shown in fig. 4. Contrary to other component placement diagrams, *this diagram is shown from the etched side of the board.*

rotor settings in the photographs. Of the three units I have built, the only one which did not proceed smoothly to this point had a defective IC section.

Now either proceed to the tripler or (optional) go back and fine tune the doublers. Small adjustments to the coupling capacitors tapped off L1 and L2, to the output of the voltage regulator (vary the 330-ohm resistor), and to all previous tuning points may help.

Finally, install cable **A**. Approximately 0.5 mW (−3 dBm) should be available at J2. If the output is low, varying the voltage on Q3 and Q4 (set by CR3) may help. If the output is still low, wrap a small square of aluminum foil over the bare end of a *Q-tip* and poke around the stripline circuitry while monitoring the output; if it is working normally, nothing should produce a significant *increase* in output. However, if a balanced mixer[4] with variable dc bias is used, local oscillator powers as low as −10 dBm will probably not significantly degrade the noise figure, if the mixer bias is set at the LO level to be used. This combination was the lowest noise-figure 1296-MHz converter measured at the 1977 Eastern VHF/UHF Conference.

The printed stripline (technically *microstrip*) elements, the rat-race and filter, were chosen partly for their non-critical nature. They have moderate inherent bandwidth, so the output frequency could be shifted ±5% with no changes. Conversely, the dimensions are not too critical. For example, the characteristic impedance of the rat-race is approximately 66 ohms vs the design value of 70 ohms because it was laid out using standard printed-circuit tape. For greater frequency changes, it is primarily the length of the lines, rather than the width, which changes. If major changes are contemplated, further research[3,5] is suggested.

This local-oscillator chain, together with a simple balanced mixer, provides a relatively easy way to listen on 1296 MHz; a varactor tripler would complete a basic station. Since there are no critical adjustments, it can be confidently built with moderate test equipment and skill, yet it has the stability and spectral purity required for an advanced station. Portions of the chain are usable for 220 and 432 MHz as they stand; the addition of a doubler to 2304 MHz is contemplated.

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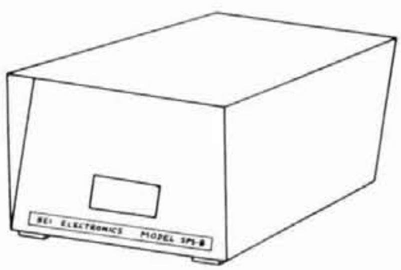

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evaluating noise sideband performance in oscillators

Receiver local oscillators produce noise sidebands which degrade receiver performance — how good is the oscillator in your receiver? Here are some ways to find out

All oscillators produce noise sidebands which can degrade the performance of the equipment in which the oscillator is used. In receivers, noisy local-oscillator stages produce reciprocal mixing products which cause blocking.[1,2] For oscillators used in high-frequency or vhf/uhf equipment, it's sufficient to measure the noise sideband performance in dB/Hz between 500 Hz and 100 kHz from the carrier; a way of doing this is described in this article. Also discussed is a method for determining local-oscillator noise sideband levels by measuring the receiver's blocking performance.

Fig. 1 shows the test setup recommended by the National Bureau of Standards for measuring noise sideband performance between 500 Hz and 100 kHz from the oscillator carrier.* A signal generator with extremely high spectral purity such as a crystal oscillator is used as a reference oscillator for a high-level, double-balanced mixer. The output of the oscillator under test is fed into the rf port of the mixer through a variable attenuator. The output of the double-balanced mixer is then connected to a waveform analyzer.

A simplified block diagram of a waveform analyzer is shown in **fig. 2**. The signal to be analyzed is mixed

*For a small modulation index, as in a hard-limiting oscillator there is no difference between fm and a-m noise.

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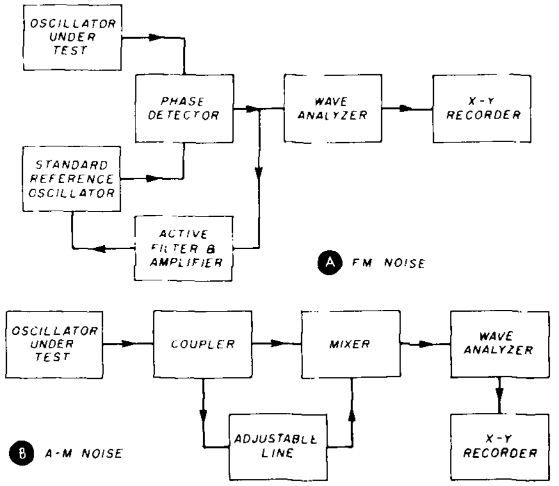


fig. 1. Test setups recommended by NBS for measuring fm noise (phase fluctuations), and a-m noise (amplitude fluctuations). For a low modulation index, as is the case of a hard-limiting oscillator, there is no difference between a-m and fm noise.

with a tunable local oscillator and the i-f output is applied to a narrow bandpass filter. In operation the frequency of the VFO is adjusted so the desired component of the input waveform is equal to the center frequency of the selective filter. Thus the component to be measured is fixed to a predetermined frequency, is amplified, and measured at this fixed frequency; other frequency components in the input waveform are rejected by the narrow bandpass filter. **Fig. 3** shows the filter response curve of the Rohde & Schwartz FAT3 waveform analyzer which has an i-f at 80 kHz; the bandwidth is 4 Hz at the 3 dB points.

To calibrate the test setup, the wave analyzer is set to full sensitivity, where zero equals 10 mV and −80 dB equals 1 μV. The oscillator to be analyzed is then set to a frequency 30 kHz from the crystal oscillator, and the variable attenuator is adjusted so that the double-balanced mixer output gives a full-scale reading equivalent to 10 mV.

If an instrument with high dynamic range and a linear dB scale is used, the attenuator can remain in the same position during the entire test procedure. Where limited dynamic range is available, the attenuator is set so at full sensitivity the instrument will indicate 1 μV on a quasi-linear meter, and the logarithmic display will be simulated later by taking readings from the attenuator.

Now connect a frequency counter to the oscillator, and change the frequency either 1 kHz up or down from the original setting, which was offset 30 kHz from the crystal oscillator. Then reconnect the signal generator to the double-balanced mixer. The waveform analyzer, which typically has three or four bandwidths available, is set at 100-Hz bandwidth. Assume that the logarithmic display on the analyzer indicated −60 dB when the original frequency was selected. This means that, in a 100-Hz bandwidth, the noise sideband voltage is 60 dB below the carrier. Since the noise sidebands are expressed in dB per 1 Hz (dB/Hz), a 20 dB correction factor must be added to the measured 60 dB because of the 100-Hz bandwidth; in this case the noise sideband performance is 80 dB/Hz.

Practical experience indicates that an oscillator which exhibits 80 dB/Hz noise performance is not very good, so the crystal oscillator can be considered much better and its noise contribution may be neglected.

sideband noise evaluator

Commercial waveform analyzers with a built-in, double-balanced mixer, linear logarithmic displays, and wide choice of bandwidths are extremely expensive — as much as $30,000, — so few are in amateur hands. However, you can build an instrument with similar performance for less than $300 if you're willing to give up certain features:

1. Choice of multiple bandwidth
2. Extremely linear logarithmic scale
3. Small i-f bandwidth

The resulting instrument, which I call the Sideband Noise Evaluator, is no longer in the class of the waveform analyzer which permits measurements from a few Hz to 60 kHz or more, but it's ideal for measuring the noise sideband performance of oscillators.

A block diagram of the device is shown in **fig. 4.** The local oscillator must deliver +17 dBm drive and

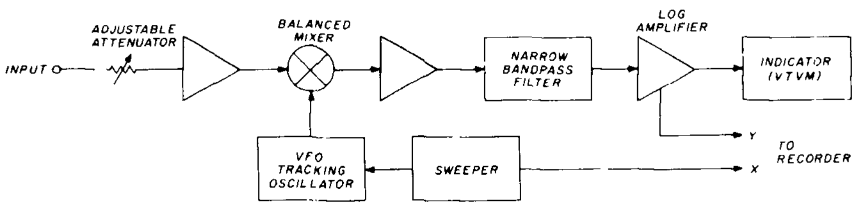


fig. 2. Simplified block diagram of a waveform analyzer. The frequency of the VFO is adjusted so the desired component of the input signal is mixed to the center frequency of the narrow bandwidth filter where it is amplified and measured.

the mixer must accept any frequency combination between 70 kHz and 200 MHz. This frequency range is sufficient to evaluate most oscillators used in radio communications systems.

circuit description

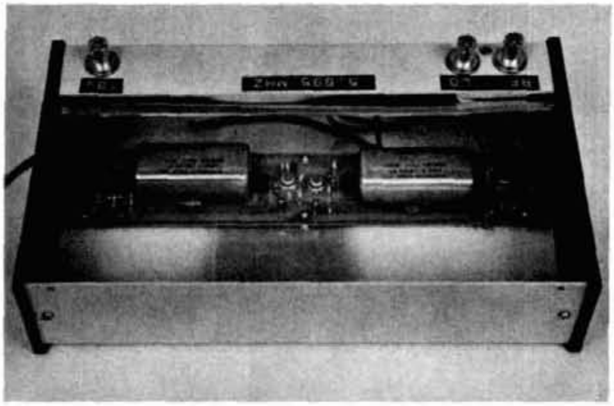
A feedback amplifier is used as a wideband termination for the mixer and as a low-noise preamplifier. It also compensates for losses of the following crystal filter. While any i-f between 1 MHz and 10 MHz can be used for which suitably selective filters can be purchased, the chosen frequency of 5.695 MHz was based on the use of 125-Hz wide filters built by Sherwood Engineering.*

Two high-gain wideband amplifiers boost the signal by about 80 dB and feed a second crystal filter. A post-amplifier is used to compensate for the second filter losses; a detection circuit drives an operational amplifier which in turn drives a meter and the agc input for the wideband amplifiers.

Two wideband amplifiers are used in the circuit, so there is more than 120 dB of agc range available. Thus, the instrument should provide at least 100 dB dynamic range with the 125-Hz bandwidth filters. If you consider a sensitivity of 5 μV relative to 1 volt emf, this results in 120 dB/Hz resolution.

Resolution could be increased 10 dB by using an i-f filter 1 kHz to 2 kHz wide, but it would not be possible to make noise measurements closer than 4 kHz from the carrier. Therefore, the user must decide whether more than 120 dB dynamic range is required, or if it's more important to measure noise close to the carrier. In my opinion, close-in noise is very important because of the number of CW stations which can be heard when receiver selectivity is set to 500 Hz; that's why I used the narrow 125-Hz filters.

A complete schematic for the Sideband Noise



Construction of the Sideband Noise Evaluator showing the location of the printed-circuit board in the Ten-Tec enclosure. The meter is connected to the circuit with two lengths of miniature coaxial cable.

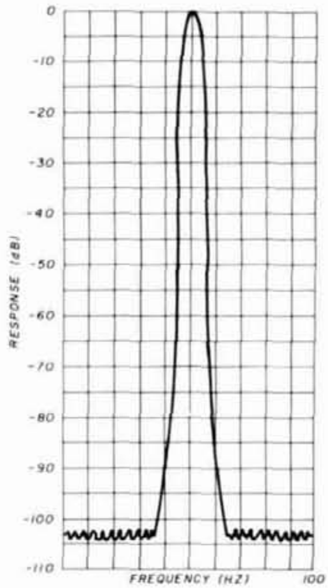


fig. 3. Response curve for the 4-Hz wide bandpass filter used in the Rohde & Schwarz FAT3 waveform analyzer. The center frequency is at 80 kHz.

Evaluator is shown in **fig. 5**. The Mini-Circuit Laboratories† SRA3H was found to be the ideal choice for the input mixer; a 2N5109 CATV transistor with voltage and current feedback both amplifies and provides proper termination to the mixer's i-f port. The 5.695-MHz filters from Sherwood Engineering are designed for 50-ohm input and output impedance; the Mini-Circuits 4:1 and 16:1 transformers provide correct circuit matching.

The Motorola MC1590s which are used in the wideband amplifier are the perfect choice for this application.[3] The gain of the stages is set by resistors R7 and R10, as discussed in the Motorola data sheet.

The Texas Instruments 733 wideband amplifier IC compensates for the losses of the second narrowband filter. The three diodes at the output of the 733 act as a voltage doubler and provide suitable time constants and thresholds to feed an LM301 operational amplifier. The 50-μA meter is shunted with 470 ohms to limit the reading to 45 μA; for a full-scale reading this resistance value must be increased slightly. The two 10k resistors equalize agc distribution to the MC1590 ICs.

After this circuit was built, I found two problems with the Sherwood filters. First, the center frequency was off by 47 Hz; relative to a 6-dB bandwidth of 125 Hz, the discrepancy expressed in per cent is unreas-

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

†Mini-Circuits Laboratory, 2625 East 14th Street, Brooklyn, New York 11235.

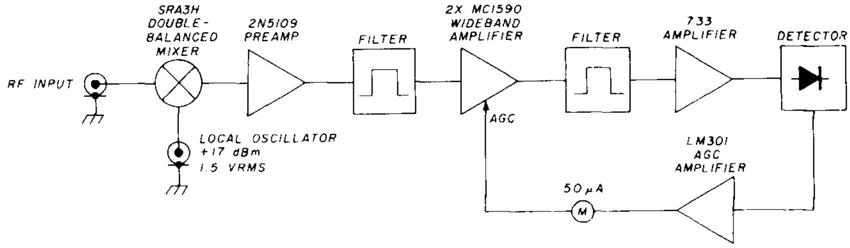


fig. 4. Block diagram of the Sideband Noise Evaluator which may be used to measure oscillator sideband noise. A schematic diagram for the instrument is shown in fig. 5.

onably high. Second, the insertion loss of the Sherwood filters was substantially higher than expected. When this was discovered, the 2N5109 preamplifier circuit between the double-balanced mixer and the first filter was redesigned. The circuit of the new amplifier is somewhat more complex (see **fig. 6**) and costs an additional $5, but has nearly 20 dB gain. The gain of the circuit can be adjusted by changing the value of the unbypassed 50-ohm resistor in the emitter circuit of the first 2N5109.

Unfortunately, the circuit board in **fig. 7** was designed for the single 2N5109 preamplifier (**fig. 5**), but it shouldn't be too difficult to lay out a new PC board if you wish to use the improved preamplifier. This may be worthwhile because the improved preamp expands the instrument's dynamic range by almost 20 dB due to a substantial decrease in noise figure.

calibration

The Sideband Noise Evaluator requires *no* adjustments! It should be possible for amateurs who build this test setup to use it immediately without difficulty. **Fig. 8** is a graph of output readings as a function of the rf input voltage (measured with a + 12.0 volt power supply). While the scale is not as linear as that available with a commercial waveform analyzer, the curve permits adequate resolution for most noise sideband measurements.

If the two-stage preamplifier is used, however,

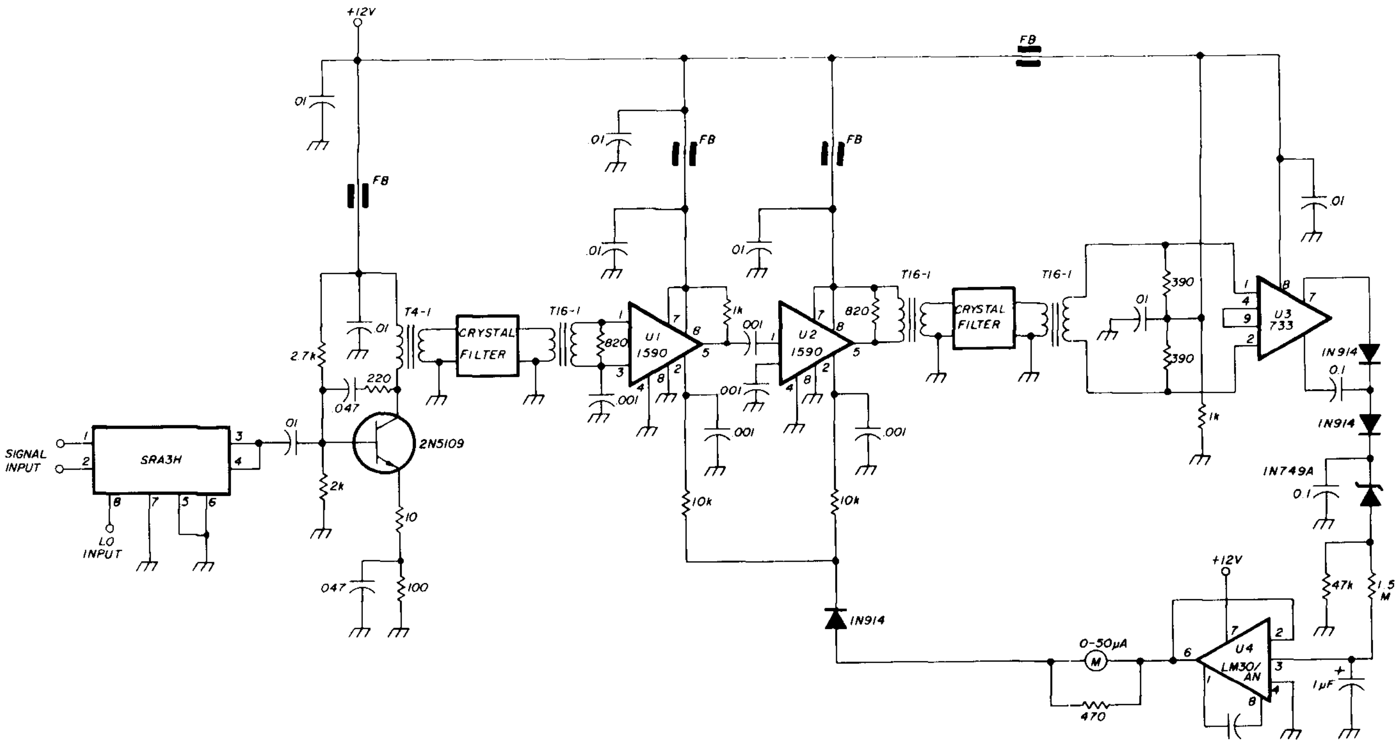


fig. 5. Schematic diagram of the Sideband Noise Evaluator. Circuit operation is discussed in the text. The filters are Sherwood Engineering 5.695-MHz crystal filters with 125-Hz bandwidth. The 4:1 and 16:1 transformers are from Mini-Circuits Labs.

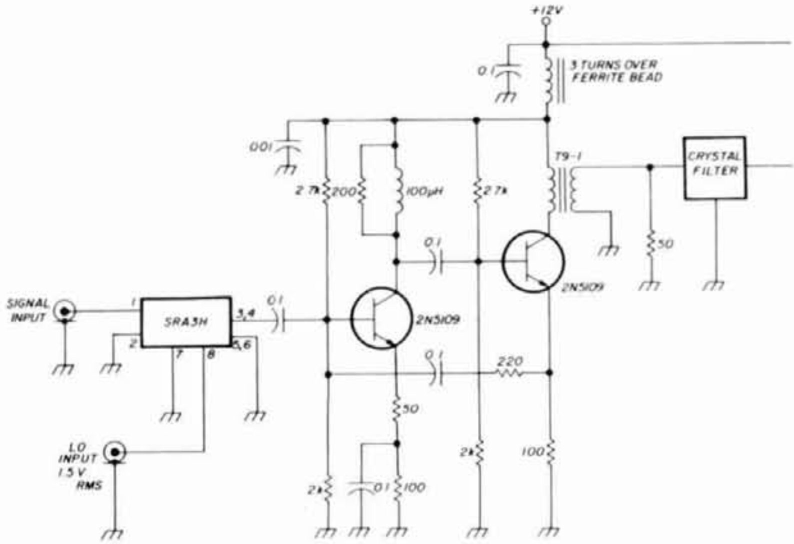


fig. 6. Improved input preamplifier for the Sideband Noise Evaluator which has nearly 20 dB gain. Gain can be adjusted by changing the unbypassed 50-ohm resistor in the emitter of the first 2N5109.

recalibration is necessary because of the higher gain. I recommend changing resistor R15 across the meter (**fig. 5**) to compensate for the increased sensitivity of the instrument with the improved preamp.

crystal oscillators

To obtain full use of the Sideband Noise Evaluator, I suggest you build a set of crystal oscillators. Fundamental-frequency crystals can be purchased that operate between 400 kHz and 30 MHz in what is frequently called the parallel-resonant mode (which should more accurately be called the inductive mode). **Fig. 9** shows a suitable low-noise crystal oscillator circuit with a wideband postamplifier that delivers the required + 17 dBm output level or slightly more. Any inductive-mode crystal between 400 kHz and 30 MHz can be plugged into this circuit and give useful output without any adjustments.

The frequency range between 30 MHz and 100 MHz can be covered by a crystal oscillator which uses either a third- or fifth-overtone crystal. However, the oscillators must be tuned. Various overtone

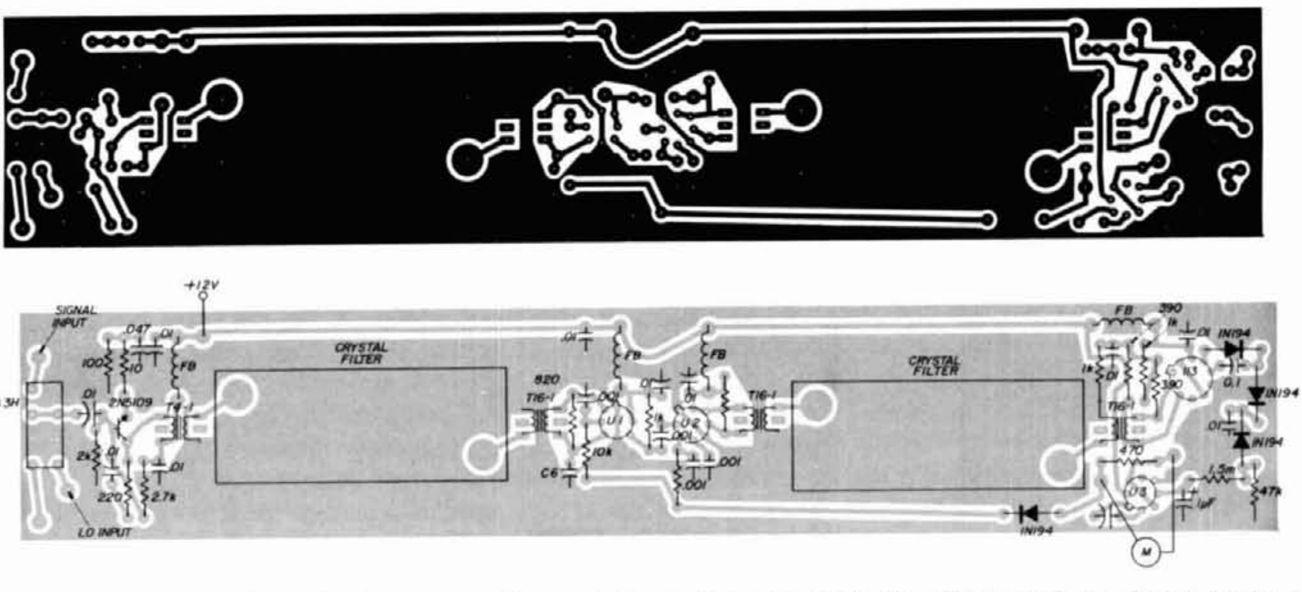


fig. 7. Printed-circuit layout (above) and component placement diagram (below) for the Sideband Noise Evaluator. Note that this artwork is approximately 67% of full size — a full-size PC layout is available from *ham radio* upon receipt of a self-addressed, stamped envelope.

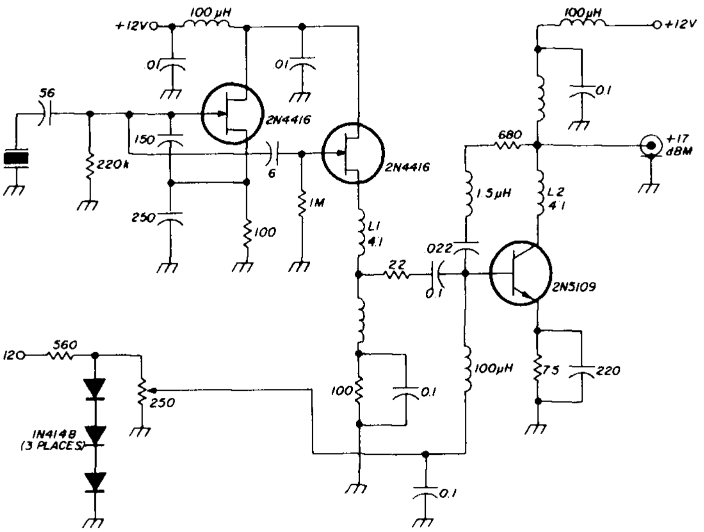


fig. 9. Low-noise crystal oscillator for operation from 400 kHz to 30 MHz; output is +17 dBm. No adjustments are required. Inductors L1 and L2 are 6 turns, center tapped, on 1/4-inch (6-mm) TC9 core.

oscillator circuits have been described in the literature, but the authors have not discussed either short-term stability or sideband noise performance. Probably the worst and noisiest of all oscillator circuits places the crystal between the transistor base and ground (**fig. 10**). The reason for the high noise contribution is that this circuit severely degrades the *Q* of *the crystal. The noise sideband performance* is partially dependent on the circuit but is determined primarily by the *Q* of the resonator: an LC circuit, a high-*Q* cavity, or a quartz crystal, and the latter has the highest *Q* of all known resonators.

Fig. 11 shows a crystal oscillator circuit that can be used for third- and fifth-overtone crystals in the frequency range from 30 to 100 MHz, and delivers +17 dBm to the double-balanced mixer. This circuit combines the best possible noise performance with high output power and excellent stability.[4]

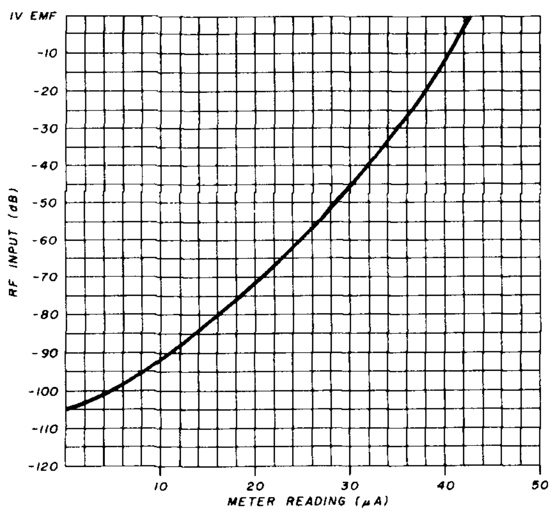
The noise performance of the crystal oscillator circuits of **figs. 9** and **11** is better than 120 dB/Hz at 1 kHz from the carrier, and 150 dB/Hz or more at 20 kHz from the carrier. Because of their excellent noise performance, these circuits can be used as local oscillators without degrading receiver performance; very few oscillators and practically no frequency synthesizers achieve their low-noise sideband levels.

measuring your receiver's oscillator sideband noise

The easiest way to measure the noise sideband performance of the local oscillator in your receiver is to measure the "blocking" or reciprocal mixing. First, accurately calibrate the receiver's S-meter between S1 and S9 + 40 dB using a signal generator and an accurate attenuator (a suitable band is 14 MHz). For the signal generator I recommend a crystal oscillator which uses a 14-MHz crystal.

Tune the receiver to the frequency of the signal generator and increase the input to the receiver so the S-meter reads S9 + 40 dB. Move the tuning dial 10 kHz and note the S-meter reading; more than likely it will be S6. Assuming S9 is 100 µV emf (50 µV terminated), and each S-unit is exactly 6 dB, then S6 is approximately 6 µV. S9 + 40 dB is 5 mV, so the difference between the two signals, for *all* practical purposes, is 60 dB. Since the measurement will probably be made with a 2.7 to 3 kHz ssb filter in the receiver, the conversion factor from 3 kHz to 1 Hz is about 35 dB (10 log BW_{Hz}). Therefore, the noise sideband performance of the internal oscillator is 95 dB/Hz

To fully utilize the low intermodulation capabilities available with high-level, double-balanced mixers in



rf input	current reading	volts emf (unterminated)
1 volt	42 µA	1 volt
−10 dB	40 µA	316 mV
−20 dB	38 µA	100 mV
−30 dB	34 µA	32 mV
−40 dB	31 µA	10 mV
−50 dB	28 µA	3.2 mV
−60 dB	24 µA	1 mV
−70 dB	21 µA	316 µV
−80 dB	16 µA	100 µV
−90 dB	11 µA	32 µV
−100 dB	5 µA	10 µV
−106 dB	1 µA	5 µV

fig. 8. Calibration curve for the Sideband Noise Evaluator shows rf input in dB relative to 1 volt emf vs indicated output current. To convert to dB/Hz add 20 dB (because of 100 Hz filter bandwidth).

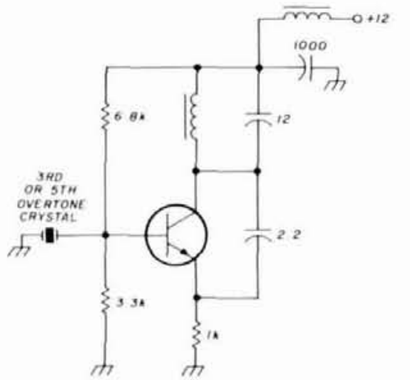


fig. 10. Popular crystal oscillator circuit that is very noisy because the *Q* of the crystal is severely degraded. This circuit is not recommended for any application.

modern communications receivers, the noise sideband performance must be at least 120 dB/Hz at 10 kHz from the carrier. To meet this criterion, the noise sideband performance of the receiver LO in the above example must be improved by 25 dB. I recommend that every designer of high-frequency of vhf/uhf receivers or local oscillators build a Sideband Noise Evaluator to ensure noise performance of at least 120 dB/Hz at 10 kHz from the carrier.[5]

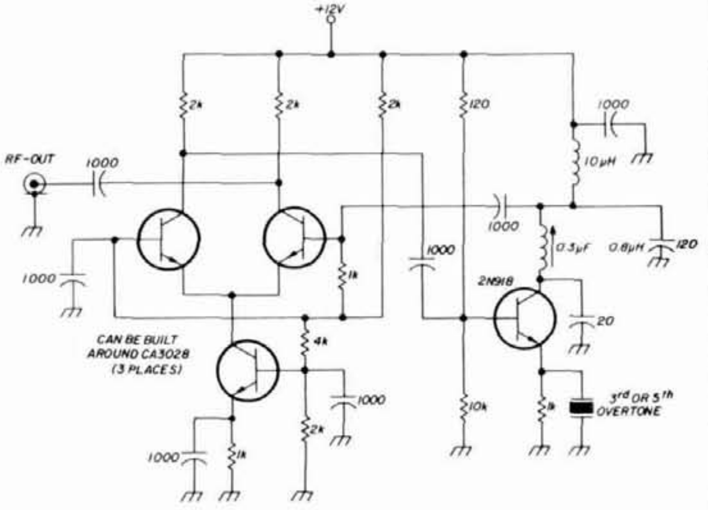


fig. 11. Ultra low-noise crystal oscillator circuit which can be used with third- and fifth-overtone crystals in the frequency range from 30 to 100 MHz. Output is + 17 dBm.

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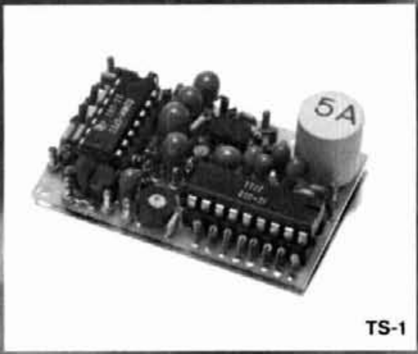
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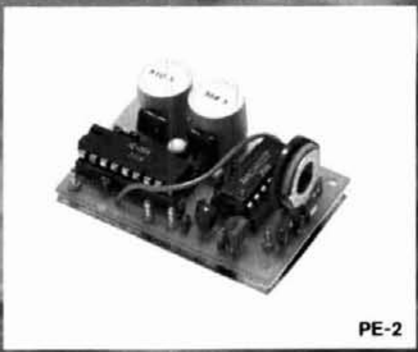
TS-1JR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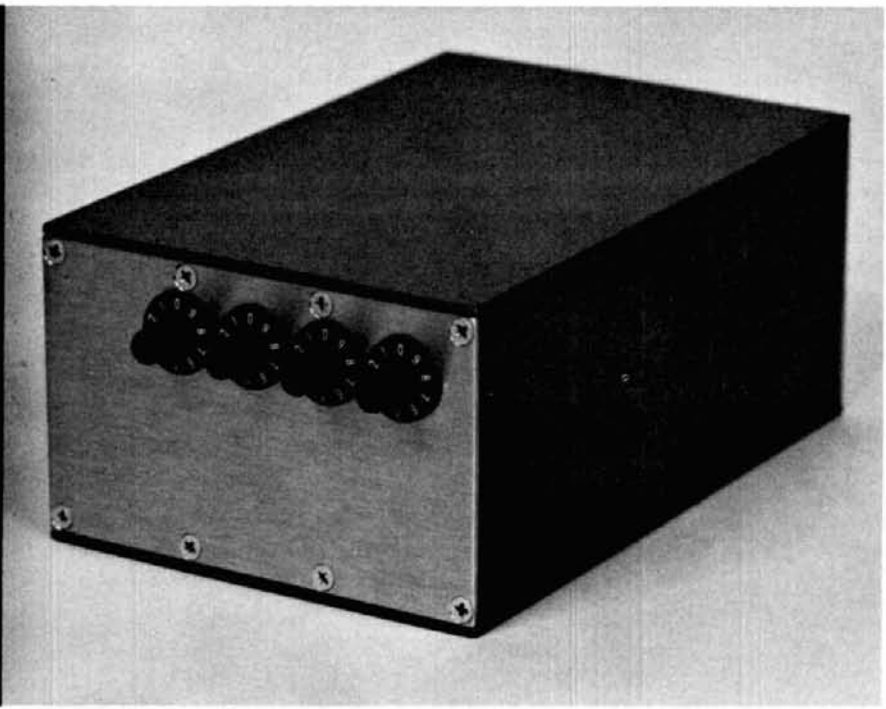

frequency synthesized local-oscillator system

for the high-frequency amateur bands

Design of a versatile frequency synthesizer system for high-frequency use which features exceptional spectral purity, 10-Hz resolution, and low power-consumption

This article is the first of a series which will describe a complete phase-locked local-oscillator system for amateur equipment with a single-conversion, 9-MHz i-f. The system offers performance which will meet the requirements of the advanced communication techniques of the 1980s — techniques which require frequency accuracy, stability, and calibration typically 100 times better than the best commercial equipment being sold today in the amateur market. This article describes the basic VFO synthesizer. Future articles will describe the phase-locked, 9-MHz BFO system, a universal phase-locked up-converter permitting operation on all bands, 160 through 10 meters, and a variety of "acrobatic" tuning methods possible only with frequency synthesizers.

frequency synthesizer accuracy

With routine care, a home builder can put together a frequency standard which will hold to 1 part in 10^7 under room-temperature conditions. I believe that

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every amateur station should have such a standard. Our rigs today usually have several free-tuning oscillators, each one subject to calibration errors, drift, and other inaccuracies, and the usual use of the standard is to ensure that these inaccuracies are held within acceptable limits.

Most operators are satisfied if they are within a few hundred hertz. The typical digital readout which is used in place of a VFO dial resolves the frequency only to the nearest 100 Hz. Thus, even if the counter timebase and all the oscillators not being measured are perfectly on frequency, a ±50 Hz uncertainty remains. To gain the very real 20-dB signal-to-noise level improvement possible with coherent CW and similar methods of radio communications now in development, the total frequency error must be held *within 1 or 2 Hz.*

The only workable method of achieving this precision in a variable-frequency system is to phase-lock every oscillator to the frequency standard. With this arrangement, proper setting of the standard automatically imparts the same accuracy to every other oscillator.

The phase-locked frequency synthesizer is well established in the vhf-fm business, where its agility saves buying hundreds of crystals. When the channel spacing is 10 kHz and the width of the output frequency band is only a few per cent of its center frequency, the synthesizer is easy to build. The requirements for a high-frequency synthesizer, however, are an entirely different story. Outputs must cover a wide range and 10 or 100 Hz "channel spacing" is required, but switching speed must still be very fast. Spurious outputs must be exceptionally low to prevent degrading the performance of wide-dynamic-range receiver designs. In addition, the synthesizer must be, like a VFO, easy to tune. Before going into the details of a practical circuit which meets these requirements, I'd like to discuss one problem which plagues most synthesizer designs.

synthesizer noise

All oscillators generate noise in addition to the desired signal. Such noise is usually classified into three categories: harmonics, nonharmonic discrete spurs, and phase noise. Harmonics are the easiest to eliminate. In applications where good harmonic suppression is required, a bandpass or lowpass filter at the output is usually sufficient.

Nonharmonic spurs (parasitics) in oscillators are extraneous, unstable outputs at unpredictable frequencies. These are stopped by proper bypassing, shielding, and filtering in the design and construction

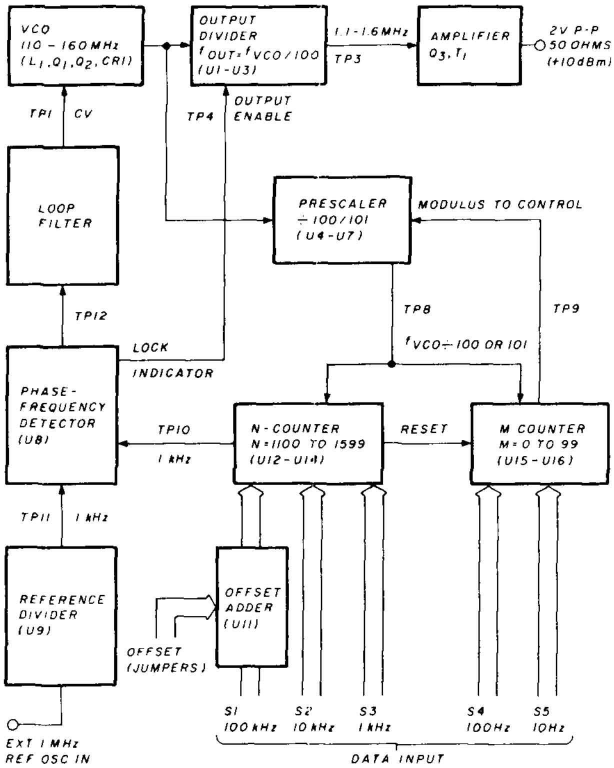
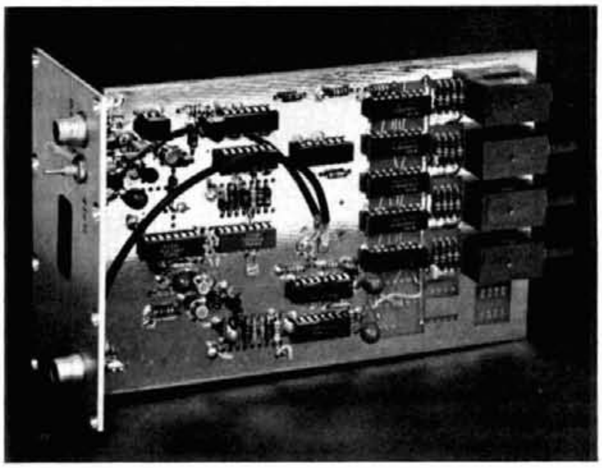


fig. 1. Block diagram of the basic synthesizer for high-frequency use. The VCO operates from 110 to 160 MHz; this same basic scheme may be used for a number of applications in the range from 500 kHz to 16 MHz. Specifications for two hf synthesizers based on this system are given in table 1.

of the circuit. In synthesizers there are several additional sources of spurs: digital counter noise, mixer intermodulation products, feedthrough of an intermediate oscillator's signal to the output, modulation of a VCO by tiny amounts of the loop reference signal from the phase detector -- these all serve to tremendously complicate designs. The presence of spurs in a receiver synthesizer's output produces birdies and the appearance of phantom signals in the i-f passband; the effect is very similar to that of inadequate image suppression.

An ideal oscillator would concentrate all its output energy at just one frequency, the desired carrier frequency. Real oscillators behave as if they were modulated by a broadband hissing noise. The result is that the oscillator's output energy is not perfectly concentrated at the carrier, but is smeared out above and below the carrier for many tens of kilohertz. The power level of this sideband noise typically diminishes in direct proportion to its frequency offset from the carrier.

If you had a super-selective bandpass filter with a



Construction of the frequency synthesizer showing the data input switches (right) and output (top left). The 1 MHz TTL reference signal is connected to the circuitry through the coaxial cable.

1-Hz passband and used this filter to measure the levels of this noise compared to the carrier, you would find that a well-designed vhf oscillator would produce sideband noise about 80 dB below the carrier for a frequency 100 Hz separated from the carrier, 100 dB below at 1 kHz, and 120 dB at 10 kHz. The total power level of uniformly distributed noise which reaches the detector of a receiver is directly proportional to the i-f bandwidth. Thus, if instead of a 1-Hz filter you used a 2-kHz filter, the noise reading would be two thousand times, or 33 dB, worse: only 88 dB down at 10 kHz offset.

It has been known for many years that oscillators produce this phase noise, but until recently oscillator noise was among the least of the designer's worries. The bad effects of this noise were nothing compared to the effects of mixer intermodulation and front-end overload! But today the situation has reversed. During the past few years we have seen spectacular improvements in the design of receiver front ends combined with an increasing interest in frequency synthesizers. But a phase-locked synthesizer requires a voltage-tuned oscillator, and to replace a variable capacitor with a varactor diode instantly increases the oscillator noise level by as much as 20 dB. So now instead of a −88 dB noise level 10 kHz from the carrier, you have only −68 dB.

What does this mean in terms of performance? Assume you have a state-of-the-art i-f filter for ssb which guarantees suppression of at least 120 dB for signals 10 kHz away from the center. Theoretically you could listen to a 1 microvolt DX station without interference from a carrier 10 kHz away which was 100 dB stronger, or 100 millivolts. But instead, the 68 dB noise from the oscillator can mix with a carrier only 70 dB stronger 10 kHz away and produce a hissing noise which would mask the weak DX signal. This effect is called reciprocal mixing, and if a synthesizer signal is not free of noise, it can negate all the high-performance features of the high-intercept mixer and super i-f filters.

Fortunately, there is a way to improve the noise performance of a wide-tuning range VCO, and this method also gives us a synthesizer which can switch fast between frequencies while having very fine narrow channel spacing.

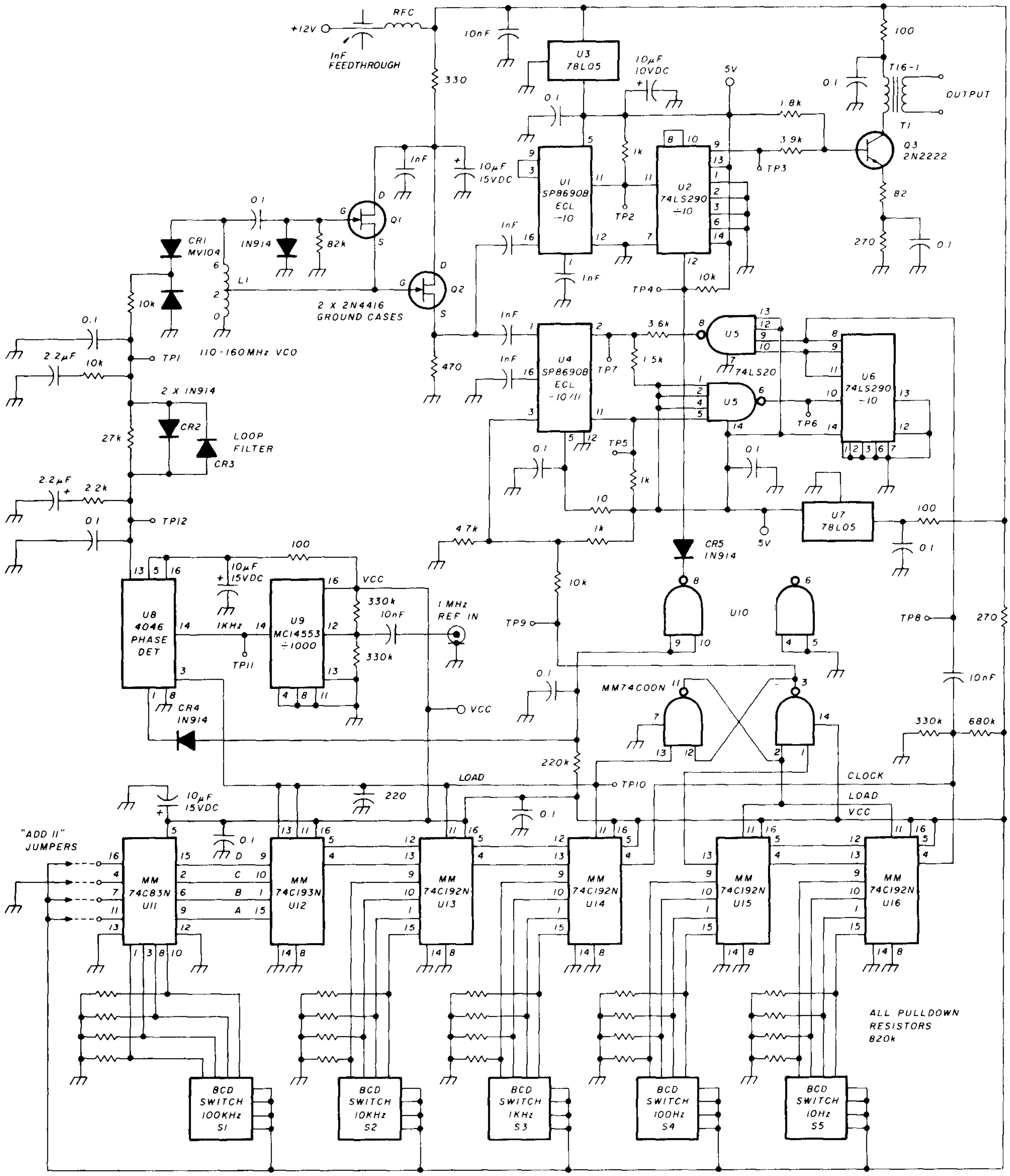
VFO synthesizer

Two versions of the synthesizer can be built on the same circuit board. The first, or A version, covers 5 to 6 MHz in 100-Hz steps. It provides coverage of the 80 and 20 meter bands when used with a single-conversion i-f at 9 MHz. Version B covers a 500-kHz range in 10 Hz steps. It forms the basis for a three-loop local-oscillator system which covers all bands, 160 through 10 meters, with the same 9-MHz i-f. You can build an A-version synthesizer and convert it later to version B simply by adding 3 components and changing the VCO coil. **Table 1** gives specifications for the two versions.

Fig. 1 is a block diagram of the version B synthesizer. The design features low power consumption, very high spectral purity, fast switching, and rf output suitable for driving a diode double-balanced mixer. The VCO covers 110-160 MHz in 1-kHz steps, operating in a loop having a 1-kHz reference frequency. The VCO signal is divided by 100 to yield an output from 1100 to 1600 kHz in 10 Hz steps; phase noise performance is 40 dB better than the VCO, and frequency switching seems instantaneous to the ear. A novel feature is the use of the divide-by-100/101

table 1. Specifications for the hf frequency synthesizers. Figures for phase noise are minimums; typical measured values are approximately 10 dB better.

	version A	version B
Output frequency range	5.0-6.0 MHz	1.1-1.6 MHz
Resolution	100 Hz steps	10 Hz steps
Output level	+10 dBm	+10 dBm
Suppression of nonharmonic discrete spurs	70 dB	80 dB
Phase noise, below carrier in 1-Hz bandwidth, for stated offsets from carrier	80 dB, 100 Hz 100 dB, 1 kHz 120 dB, 10 kHz	90 dB, 100 Hz 110 dB, 1 kHz 130 dB, 10 kHz
Switching speed	30 ms	30 ms
Data input requirement	Parallel BCD, 10 volt CMOS levels, 4 digits for version A, 5 digits for version B.	
Reference requirements	1-MHz sine or square wave, at least 1 volts p-p; input impedance of reference input approximately 50 pF to ground.	
Power	12 volts dc, approximately 75 mA.	



CR1	Dual varactor diode (Motorola MV104)
L1	6 turns no. 22 (0.6 mm), 3 mm (1/8) ID, tapped at 2 turns
RFC	6 turns no. 28 (0.3 mm) on F754-1-06 ferrite bead
S1-S5	Miniature BCD 10-position rotary switches
T1	Broadband rf transformer (Mini-Circuits Lab T16-1)

*A complete kit of parts for this synthesizer including the double-sided PC board, data input switches, and enclosure is available for $210 from Petit Logic Systems, Post Office Box 51, Oak Harbor, Washington 98277.

fig. 2. Schematic diagram of a frequency synthesizer that provides outputs from 1.1 to 1.6 MHz in 10 Hz steps. Total power consumption is only 75 mA. All resistors are 1/4 watt carbon film or composition; all polarized capacitors are dipped tantalum; non-polarized capacitors are ceramics. A kit of parts is available from the author.*

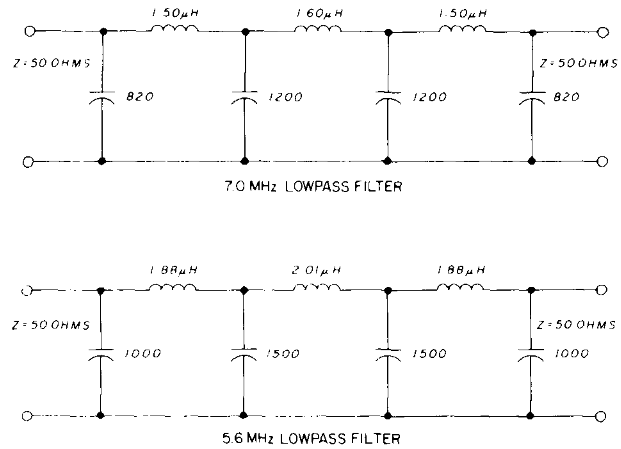


fig. 3. Seven-pole Chebychev lowpass filters for use with the version A synthesizers. Both provide about 60 dB attenuation at twice the cutoff frequency.

prescaler in the programmable counter system. With this prescaler, comparatively slow and low power-consuming CMOS programmable counters operating at frequencies below 2 MHz can control a VCO operating at 160 MHz without sacrificing the 1-kHz spacing between adjacent steps.

The 100/101 prescaler works as follows. Suppose the *N* counter is set to 1500 and the *M* counter to zero. The VCO locks at 150 MHz so the output of the prescaler is 150 ÷ 100 or 1.500 MHz. Since *N* is 1500, the *N* counter output is 1 kHz as intended. Thus the *N* counter output represents 100 times 1500 or 150,000 cycles of the VCO signal. At the end of each complete cycle of the *N* counter, both the *N* and the *M* counter are preset to the values given by switches S1 through S5.

Suppose the *M* counter is now set to 01. The prescaler goes through 1500 complete divides as before, except that the first divide is by 101 instead of 100. The result is that the *N* counter output to the phase detector represents not 150,000 cycles of the VCO output, but 150,001 cycles. Thus for any setting *S* given to the *M* counter, the first *S* prescaler divides are by 101, and the remainder are by 100.

The most significant digit of the *N* counter is a 4-bit binary counter instead of a decimal counter. Thus, the maximum count available is 1599 instead of 999. The offset adder, U11, is programmed by jumpers to cause the synthesizer to deliver output frequencies which correspond in an appropriate way to the settings of S1.

To build a version-A synthesizer, the inductance of the VCO coil is increased to resonate with the varactor at 55 MHz; U2 is omitted, and the base of Q3 is connected directly to pin 8 of U1; U11 is omitted; U12 preset is set to binary 5, and S1 is omitted. S2 then becomes the 100 kHz digit, S3 the 10 kHz digit, and so on.

assembly and checkout

If a fault develops in a PLL system, it is often difficult to locate because many faults all exhibit one symptom, the loop goes out of lock. By assembling the system one section at a time and then checking that section before continuing assembly, trouble spots can be quickly located. Here is the procedure I use:

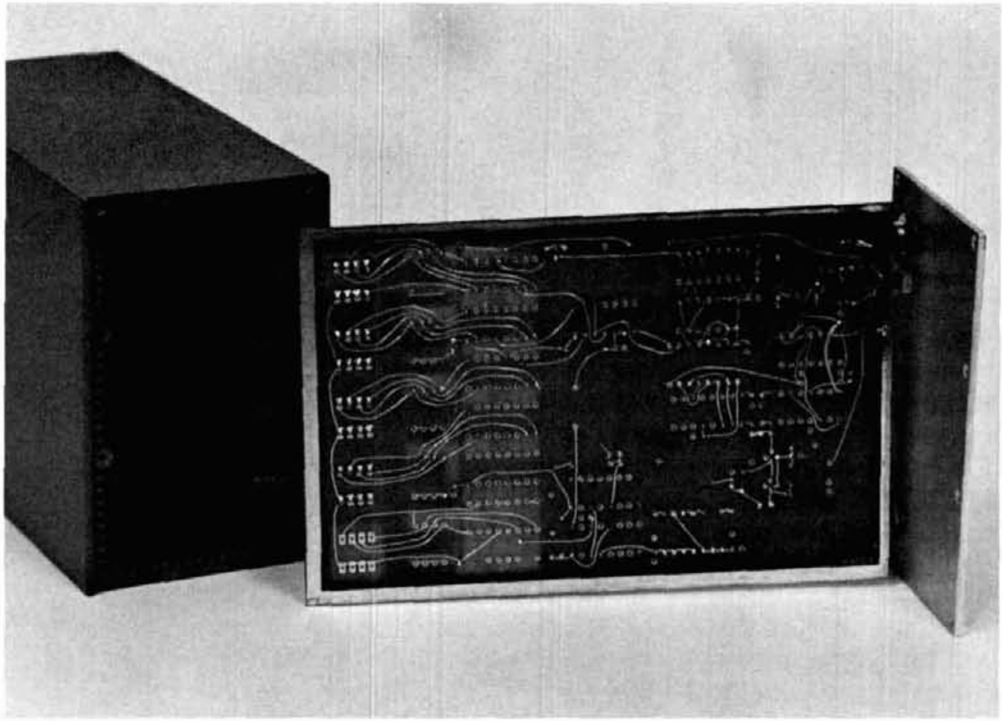
1. VCO and divide-by-100. Assemble the circuits starting at TP1, going through Q1, Q2, Q3, U1, U2, and U3. Take a potentiometer of any convenient value from 1k to 500k and connect the slider to TP1, one end to ground, and the other end to the 12-volt supply. With this you will be able to manually set the VCO control voltage to any value over its entire range. Apply power and check for at least 9 volts dc at the drain of Q1 and 5 volts at pin 5 of U1 and pin 14 of U2. Check for a TTL-level signal at a frequency below 20 MHz at TP2. Check for a TTL signal below 2 MHz at TP3, holding TP4 at ground. Connect a 51-ohm resistor across the output terminals and check for an approximately 2-volt p-p squarewave across this resistor. Leave TP4 grounded.

The following discussion assumes that the synthesizer under construction is the 1.1 to 1.6 MHz version. Set the control voltage to 2 volts and adjust L1 by slightly compressing or stretching the coil until the output frequency is 1.1 MHz. Then bring up the control voltage to 10 volts and check that the output frequency is approximately 1.6 MHz. (If you don't have a frequency counter, a standard a-m broadcast radio will work.) If the frequency isn't quite 1.6 MHz, stretch the coil slightly.

If you are building the 5-6 MHz version, set the control voltage to about 6 volts and adjust the coil L1 until the output frequency is 5.5 MHz.

2. 100/101. Assemble the circuits U4 through U7, including the resistor network leading to TP9. Apply power and check for 5 volts at pin 5 of U4 and pin 14 of U5 and U6. Connect TP9 to 12 volts. Observe a signal at TP5 which is identical to that at TP2. Check for the same signal inverted at TP6. Observe an ECL-level signal at TP7 which has the same frequency as the signal at TP3 except that it is at the lower logic level for one-tenth of the time and at the higher for nine-tenths of the time. Measure the signal at TP8. Then remove the connection from the power supply to TP9 and ground TP9; this should cause the frequency at TP8 to go down by 1 per cent.

3. Programmable counter. Assemble all the circuits of U10 through U16, the BCD switches, and the jumper-wire programming for U11. Apply power and check for at least 10 volts on pin 14 of U10, pin 5 of U11, and pins 16 of U12 through U16. Set the control voltage for 6 volts and set the data input switches to values representing the center of the synthesizer frequency range. TP10 should show an extremely narrow negative pulse at a frequency of approximately 1 kHz. Adjust the control voltage over its entire range and check that this pulse signal frequency rises and drops smoothly with your adjustments, not making sudden jumps or disappearing. With a scope connected to TP9, change the settings of S4 and S5 and observe a 1-kHz negative pulse, the width of which is proportional to the settings from 00 to 99.



Bottom view of the double-sided PC board used for the high-frequency synthesizer.

4. Reference divider and phase detector. Assemble the circuits of U8 and U9, including the loop filter except for the 27k resistor and the two diodes CR2 and CR3. Apply power and check for at least 10 volts at pins 16 of U8 and U9. Connect the output of a 1-MHz frequency standard to the *reference input* terminal. Observe a 1-kHz pulse waveform at TP11. With S1 through S5 at the mid-frequency setting, observe the voltage at TP12 while varying the control voltage. When the control voltage is above some critical value, the TP12 voltage should be near zero, and vice versa. As you move the potentiometer back and forth over this critical value, the TP12 signal should abruptly jump from near zero volts to near the positive supply voltage and back.

5. Closing the loop. Remove the potentiometer from TP1. Solder in the 27k resistor and the diodes CR2 and CR3. Apply power and observe the voltage at TP12 with a VTVM or scope using a 10-megohm probe. The voltage should be at an intermediate value. Touch the VCO components with your finger and observe the voltage rise slightly. The output frequency of the synthesizer should now correspond exactly to the settings of S1 through S5, allowing for possible inaccuracies between the counter timebase and the reference oscillator used for the synthesizer.

Remove the ground from TP4. When making large frequency changes, the output of the synthesizer will be interrupted for perhaps 0.1 second. Remove the signal from the reference oscillator and the synthesizer output should disappear and remain off until the reference oscillator is reconnected. When making frequency changes of 1 kHz or less, the output is not interrupted at all. The disable function insures that, if, for any reason, the synthesizer is not locked and stable, it will have no output.

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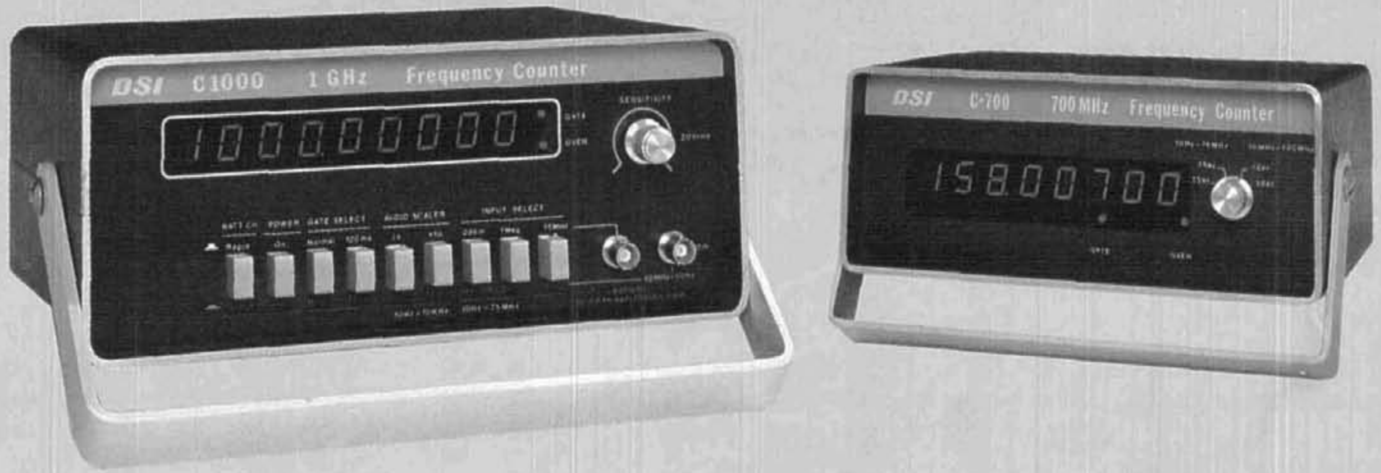

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second generation reciprocating detector

An updated version of the reciprocating detector, which can be used in solid-state receivers with high-frequency i-f strips

During the past three years I've had many requests for revisions to the reciprocating detector circuit[1] so that it can be used directly at high frequencies. Here's an updated IC design that can be used at frequencies up to 20 MHz.

background

Early attempts to directly use the RD above 5 MHz required very careful circuit layout to reduce or eliminate inter-circuit coupling, and in particular to maintain the correct phase relationship required in the feedback loop. Also, the detector portion operated as a half-wave rectifier. A current-regulating source had to be adjusted to cause the signal diode to operate at a level just below conduction, so that at frequencies above 5 MHz the diode and its circuitry ceased to perform uniformly. Result — a badly distorted detected signal.

Despite the distortion, in some cases the circuit performed well enough for signal identification. But much was to be desired. A cure for individual cases was to adjust the bias level for the current-source diode until it just conducted on noise. In most cases, with a tube receiver that produced i-f signals to the RD input exceeding the saturation level of the complete circuit, a clipped response occurred. Single sideband signals then became unmanageable because of widely varying signal levels that couldn't be controlled by the agc systems in older tube receivers.

The original circuit was designed to be used in receivers such as the Collins 51S1 and Drake R4A, which have highly selective dual or adjustable filters in the receiver i-f passband. In the 51S1 receiver the i-f output was fed to the RD through a cathode follower; the maximum output level could not exceed 3 volts. The application using the Drake R4A employed enough attenuation through the coupling to the original product detector output transformer to preclude saturation.

An updated design, which uses ICs, allows the RD to be incorporated into more modern receivers. Models of the new circuit have been made for 10.7, 16, and 20 MHz. Test models were constructed using point-to-point wiring. Later models used PC boards.

circuit description

The circuit consists of two amplifier chips, IC1, and IC2 (**fig. 1**). These are monolithic wideband amplifiers with frequency response between 10 kHz and 20 MHz. These chips are 10-lead devices in TO-5 cans. A third rf amplifier, IC3 is a balanced differential amplifier using an internal constant-current source, which eliminates the orignal problem caused

The "reciprocating detector" was designed by R.S. Badessa at M.I.T. The RD features a carrier-synthesized reference signal and requires no external bfo. The circuit offers advantages over conventional detectors in that it adjusts its bfo level automatically in proportion to the average signal level received. First introduced in *ham radio* in March, 1972, the RD has gone through several metamorphoses. The version presented here uses ICs and can be used in modern receivers using semiconductors. Also included is a design for a 10.7-MHZ ssb filter for single-passband receivers. **Editor.**

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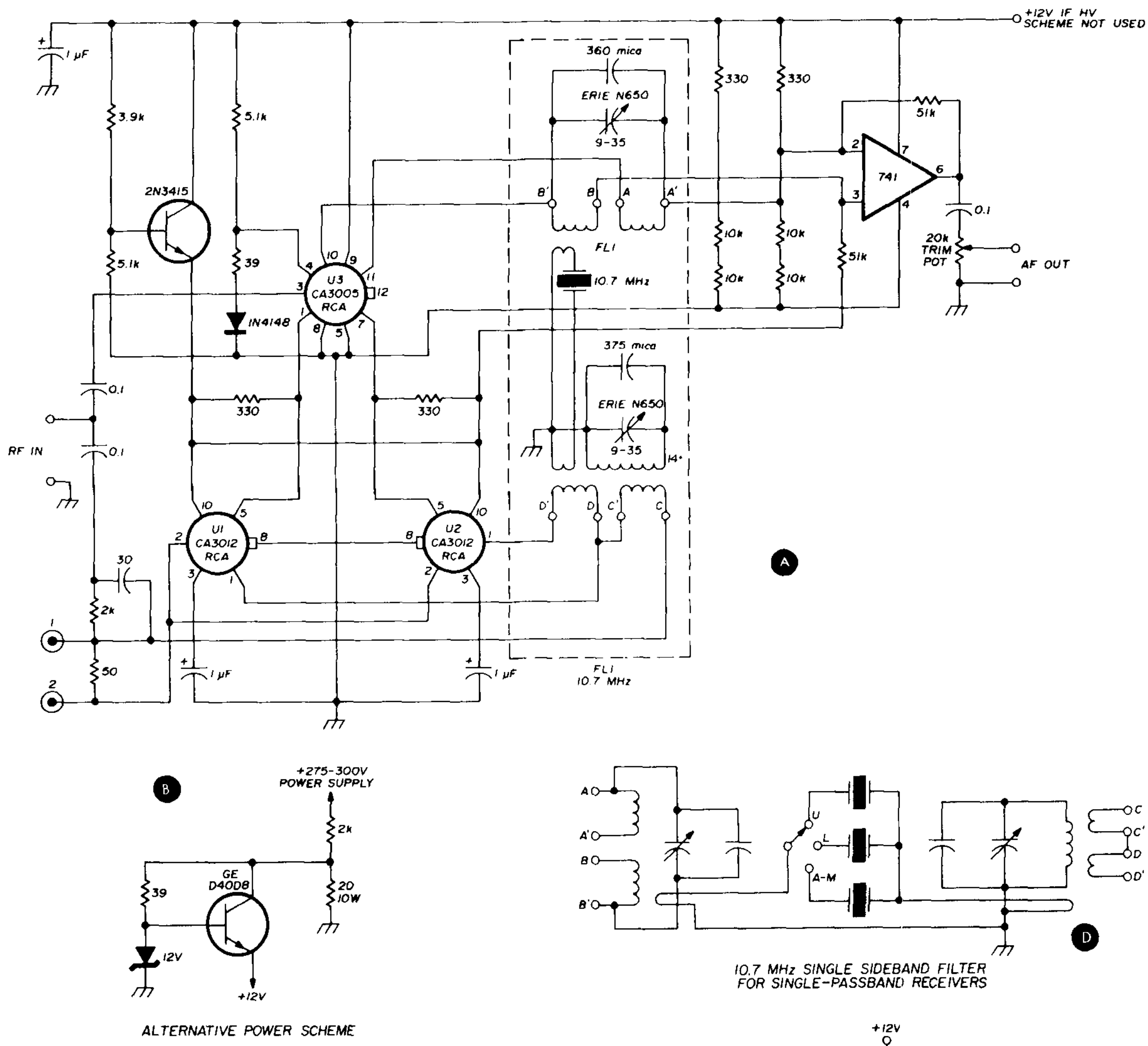

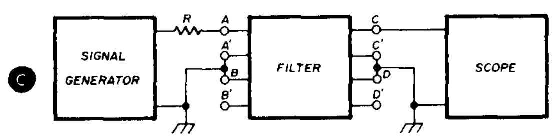

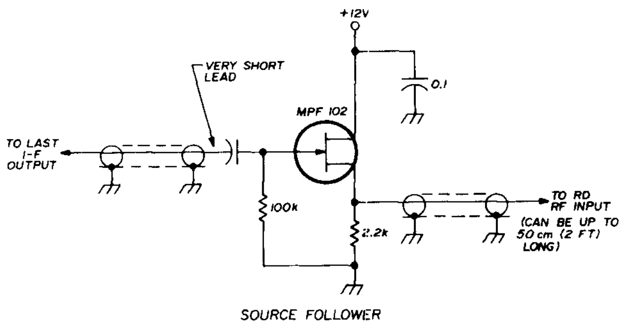

fig. 1. Schematic of the reciprocating detector MKII using wideband-amplifier ICs, (A). Also shown are an alternative power scheme for receivers using a high-voltage dc supply, (B), a test setup for adjusting the filter, (C), and a schematic for a 10.7-MHz ssb filter, (D).

by the biased half-wave rectifier. This amplifier operates from 0-100 MHz. This wideband response allows the circuit to work in the same manner as the original current source for the detector and as the reciprocating switch. These two functions are improvements over the old circuit. The dynamic range improvement alone is worth the effort.

Tracing the signal through the circuit, we see that a capacitive input circuit couples the rf signal into IC3 input. The capacitive coupling isolates any direct current that might be superimposed on the rf signal from the i-f ouput circuit. The input signal is then applied to a phase-shift network, then to one set of inputs of IC1 and IC2. These three inputs are then provided with a signal path that's essentially in series with the reference signal, or beat frequency similar to a conventional product detector.

The reference frequency is generated by filtering a portion of the received signal through a narrowband crystal filter, FL1. The push-pull output of this filter

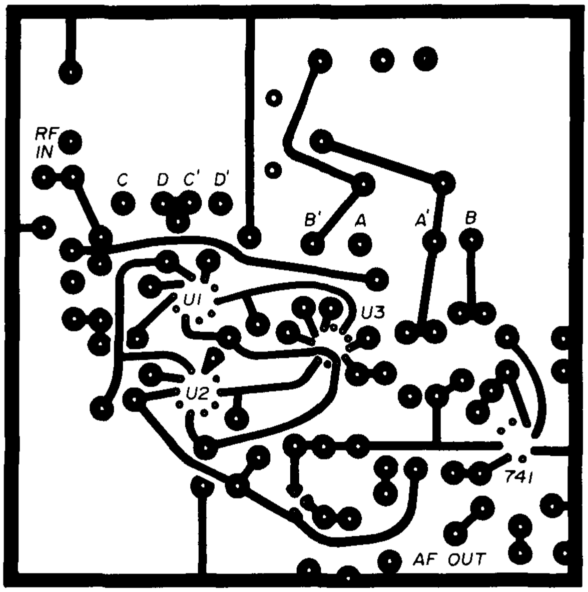

fig. 2. PC-board layout for the updated reciprocating detector.

feeds this signal into the balanced inputs of IC2 and IC3. We now have the circuit reference by virtue of a carrier-controlled feedback loop. The feedback loop response time is determined by the slow recovery time of the narrowband filter; therefore, noise pulses are reduced or eliminated.

IC4, an op amp, is an audio amplifier. The received-signal audio envelope is taken from the input filter, FL1, then applied through two lowpass filters to the 741 IC. This amplifier audio gain is established by the feedback resistor (51k). This value may be changed for a higher or lower level; however the gain will overdrive the first audio inputs of most communications receivers. A 20k trimpot at the audio amplifier output allows gain control.

The RD will work in any receiver if its i-f matches that of the RD reference filter. For best results, narrowband i-f filters offering high selectivity are a major requirement. Many receivers use dual filters offset by the correct dispersion to allow either sideband to be selected by switching in the appropriate filter section. Others use adjustable filters to obtain the same effect. I-f passband circuits using single filters, as in many of the older tube receivers, also perform nicely. With 1.8-kHz filters, however, only one side can be used unless the reference or beat frequency is displaced to center the signal in the filter. The older reciprocating detector circuits didn't provide for this problem.

Circuitry to show how a filter can be constructed for upper/lower sideband selection for use in receivers with single passband filters, incorporating three crystals, is shown in the filter circuit. If your receiver uses filters to select sidebands, then only one crystal, set in the i-f passband center, will be required.

construction of a 10.7-MHz RD filter

The filter used in an RD is not a complicated device. It has a shape factor similar to that of the old variable crystal filters. It's not a wide passband filter because it's used to select the beat-frequency signal.

On a Micro Metals T50-2 toroid core wind 14 turns of 0.2 mm (no. 32) enameled wire. Secure the wire so it won't become loose. Use tape or nylon string. Make sure the coil leads are at least 30 mm (1-1/2 inches) long and are scraped clean of insulation. Next, fold a 30-cm (12-inch) length of the same type wire in half. Twist this pair of wires until you have at least eight twists per 25 mm, or 8 twists per inch. (This is called a bifilar pair.)

Now, using the bifilar pair, wind on the same form a three-turn winding and secure it. This coil should be wound in the area not used by the previous winding, but it isn't important that it be exactly placed or spaced in this area. Next clean off each of the wire ends; then, with an ohmmeter, identify each coil separately. They will be used to complete the connections identified in the filter drawings as C, C′, D, and D′.

After each winding has been identified, the ends opposite each other can be connected to provide the center-tapped winding signified as C′ and D in **fig. 1.** This toroid will contain three coils. Now, on a second core of the same type, wind two eight-turn windings of the bifilar pair and scrape the four ends. Again, with an ohmmeter, identify each coil. These ends can be designated A, A′, B, and B′.

Mount each coil on the PC board as shown (**fig. 2**). If you're not using the board, mount the coils about 30 mm (1-1/2 inches) apart. Don't tighten the coils yet. Next, mount the crystal between each toroid, then wind a single turn of 0.2 mm (no. 32) enameled wire on each toroid, terminating one end of each coil on a crystal terminal or the switch, whichever the case may be. The other two coil ends should be connected together.

If your receiver has selectable sideband filters, a single crystal will be required. If not, then wire it as shown in the filter diagram (**fig. 1**) and include all three crystals. In this case, a selector switch must be used at the filter location. If this is the case, connect the link ends from the two coils to the appropriate switch contacts. Reed switches can be used and provide excellent low-loss control.

The filter components are for 10.7 MHz but will work at 9 MHz with different crystals. The components can be juggled to work around that frequency range. Lower frequencies will, of course, have a higher inductance value.

The crystals can be purchased from any of the manufacturers currently advertising in most of the amateur magazines. It's best to use fundamental-frequency crystals mounted in an HC6/U holder, with wire leads to make soldering easy. This doesn't preclude other types of holders or pin-mounted crystals; however, some of the alignment procedures will be a little more difficult, particularly if pressure-type holders are used.

checkout and test

Connect a signal generator and scope or rf voltmeter through terminating resistor R as shown in the test setup. Set the indicator to a high sensitivity and the signal generator to a high output level. Carefully tune the signal generator across 10.7 MHz. An indication with a very sharp upswing in level will occur when passing through crystal resonance. Carefully adjust the signal generator to the peak of the upswing. Then, with an insulated screwdriver, adjust the 9-35 pF capacitors for future increase. The scope sensitivity and the signal-generator output level will have to be reduced as the resonance of each coil is reached. Frequent readjustment of the signal generator will be required to keep it centered at crystal resonance. As the adjustments proceed you'll notice that the sharp increase at the crystal frequency will become easier to adjust.

If a three-crystal unit is to be constructed, make these adjustments at the passband center or with the a-m crystal in the circuit. The other two crystal frequencies are as sharp as that of the a-m resonant frequency and will be within the inductor resonant frequency.

To use the RD in the ssb mode with a single i-f passband filter, it won't be necessary to offset tune the receiver. Simply use it as you normally would. It's like having a crystal controlled bfo — simply switch in the appropriate crystal.

RD construction

The reciprocating detector is simplicity itself to construct. Attention to component placement is similar to that of any high-frequency device. The filter leads can be connected, after filter adjustment, to those points shown (**fig. 1**) that are alphabetically marked. Use leads as short as possible.

A slight tweaking of the filter might be required after it's installed in the receiver. Use care as to the length of the lead to the RD rf input. This is a two-way street: if the lead is too long, external pickup can cause interference to the hf i-f stages; if the lead is

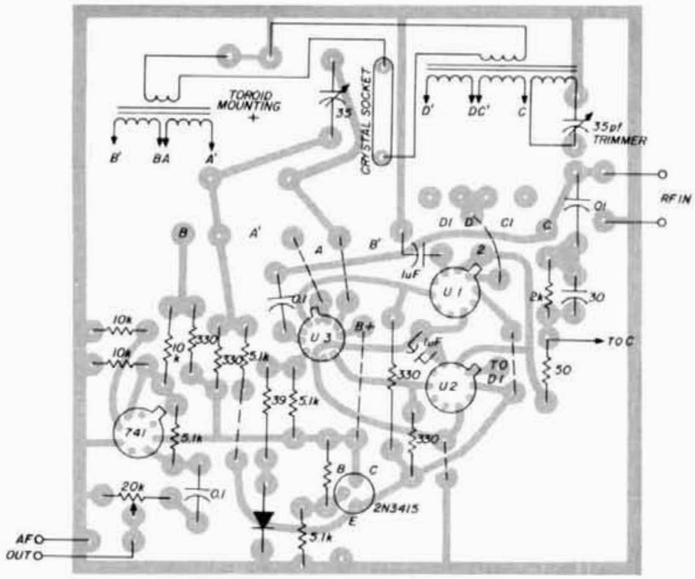

fig. 3. Component placements for the reciprocating detector circuit board.

shielded and too long, it can detune the i-f stage to which it is connected. So a short signal path is required, or an emitter or source follower will be required to reduce these effects.

The choice of how the audio is routed is up to you. It can be connected as shown in the references or used with an external amplifier. Since the output level is in the 100-millivolt level, it can easily drive an external amplifier.

power supply

Power requirements for the new RD are further simplified. The older unit required a dual balanced supply source, but this unit requires a single 12-volt supply at 40 milliamperes. The RD supply can be taken from the receiver supply filter output if it's 12 volts. Or you can use a higher-voltage supply, such as found in a tube reciver, if you use the alternative supply scheme shown in **fig. 1**.

I will be pleased to hear from anyone who has used the RD and will communicate with all who write. Please include a self-addressed, stamped envelope with your letter.

reference

1. Stirling Olberg, W1SNN, "Reciprocating Detector," *ham radio*, March, 1972, pages 32-35.

bibliography

Olberg, Stirling, W1SNN, "5-MHz WWV Receiver," *ham radio*, November, 1972, pages 44-49.
Olberg, Stirling, W1SNN, "Reciprocating Detector Converter," *ham radio*, September, 1974, pages 58-63.

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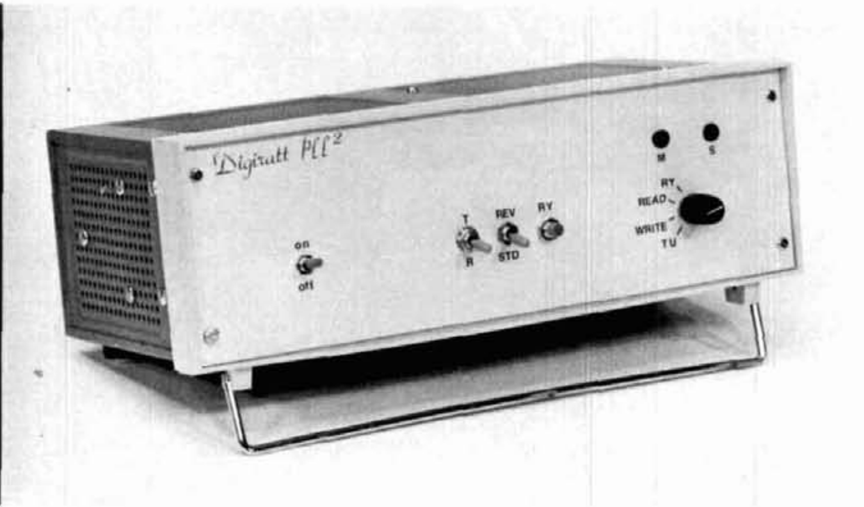


digiratt PLL²

— dual demodulator terminal unit

Continuing in the digiratt series, the PLL² demodulator uses dual phase-locked loops to help eliminate loss of signal due to fading

Phase-lock loop terminal units have been around in various forms for several years. As a means of receiving RTTY signals inexpensively, they are certainly worth considering. The one common drawback to PLL terminal units is that they decode only half the available information present in the RTTY signal. Many comments have been made by amateurs over the years that this fact really isn't a drawback at all, because there are only two possible states that the RTTY signals can be in at any one time and the absence of one condition indicates the existence of the other.

There is, however, an occurrence known as selective fading which can completely eliminate one half of the RTTY signal while leaving the other intact. If you happen to be tuned to the particular tone which fades out, you'll find your printer ceasing to operate until the tone returns.

PLL terminal units have several good points. Among these are:

1. They will follow a drifting signal until it leaves their passband.

2. Since they are inherently frequency selective, they do not require passive input filters.

3. They are not expensive.

It follows then that if a means could be found to use PLL circuits to decode both RTTY signals (mark and space), the overall usefulness of the terminal unit would be improved.

The original Digiratt PLL terminal was an attempt to design a low-cost vhf terminal unit and AFSK generator.[1] I received letters from all over the world which led me to believe that there is a large amount of interest in a simple means of decoding RTTY signals. The Digiratt PLL² is the direct result of those letters and is presented here as one possible approach to the need for such a unit.

The Digiratt PLL² is composed of two identical tone demodulators, using the 567 phased-lock loop.

By John Loughmiller, KB9AT, Route 1, Box 480C, Borden, Indiana 47106

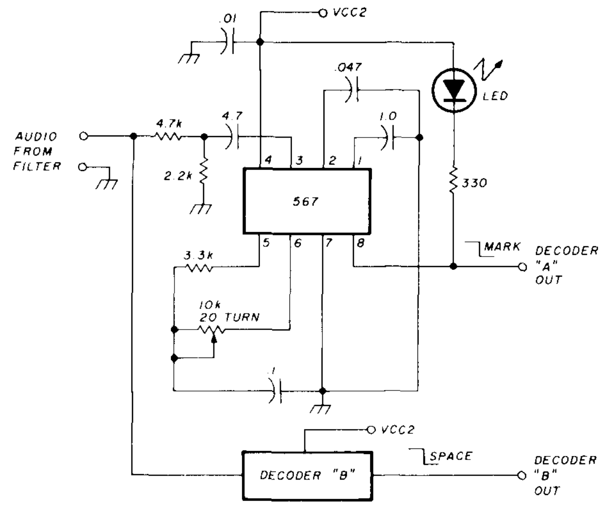


fig. 1. Schematic diagram of the 567 phase-locked loop used as a demodulator. In this case, two 567s are used; the first will demodulate the space signals while the second is for the mark frequency.

A two-stage active filter is included, and may be switched into the circuit ahead of the PLL inputs in order to detect low-level signals. Additionally, a very novel logic circuit, which has the capability to regenerate missing information during selective fading, follows the PLL portion of the terminal unit.

circuit description

The description of the PLL circuits themselves can be found in the original Digiratt article; this unit uses the same decoder circuitry. Decoder A (see **fig. 1**) is tuned to the mark frequency of 2125 Hz, with decoder B tuned to the space frequency (2295 Hz). The A decoder capture frequency is slewed low and the B decoder slewed high so that there exists an area from approximately 2190 Hz to 2220 Hz where neither PLL will conduct. The exact procedure for accomplishing this operation will be covered in the alignment section.

SELCOMP is a name the author has given to the logic responsible for the *SEL*ective fading *COMP*ensation which is part of the Digiratt PLL.[2] The schematic diagram is shown in **fig. 2**. In order to understand its operation, two separate conditions will be explained, normal full-signal operation and operation under "no space" conditions.

Normal. A low-going mark signal is buffered by U5A and then sets the RS latch, U6. The output from the latch causes the multiplexer, U8, to select its pin 5 input. The mark signal then appears at the base of Q1, the selector magnet driver circuit. A subsequent low-going space signal will reset the RS latch, causing the multiplexer to select the pin 6 input. Since the space input from the decoder is a low true signal, it is inverted by U4. Therefore, the output from the

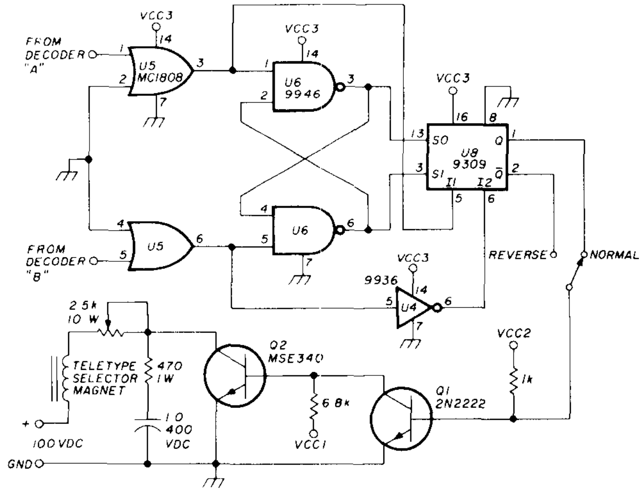


fig. 2. Under normal conditions, the multiplexer switches between the mark and $\overline{\text{space}}$ signals. If either signal is lost, the logic will continue to use the remaining signal as the drive for the selector magnets. The Q and $\overline{\text{Q}}$ outputs from U8 allow you to use either normal or inverted data. The 100 Vdc for the selector magnets comes from the power supply shown in fig. 5.

multiplexer will be low for mark and high for space. Also, the complement signal from the multiplexer can be used to provide for normal and inverted signals.

No Space. Now, assume that space data is lost. When this occurs, the RS latch will never reset and the magnet driver will receive only mark data. If mark data is lost, the selector magnet is driven by $\overline{\text{space}}$ data because the RS latch is never set and therefore

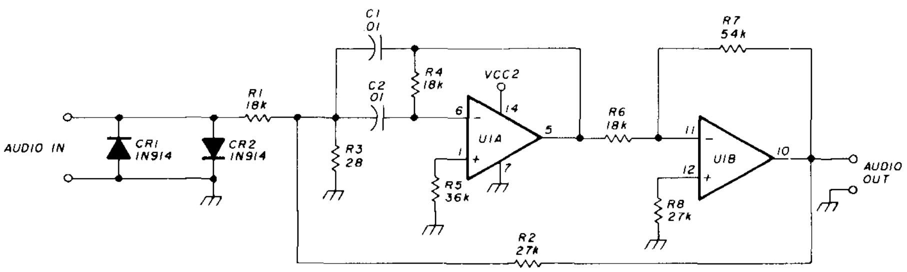


fig. 3. Schematic diagram of the 2-stage active filter. Using an LM3900, this filter provides 23-dB gain, with a designed center frequency of 2210 Hz. All resistors are 1/4 watt, 1 per cent tolerance. C1 and C2 are polystyrene capacitors.

the multiplexer always looks at the incoming inverted space data.

Because of this logic scheme, selective fading is greatly minimized. There is one condition which the logic will not correct, and that is a selective fade during a 22 ms character element. In other words, if a single bit is lost there is very little any simple system such as SELCOMP will do for the problem.

Fortunately, such rapid selective fading is relatively rare. A much more common occurrence is printer noise or external impulse type noise, which your receiver's noise blanker will generally handle.

The active filter, as shown in **fig. 3**, is a two-stage device with a gain of approximately 23 dB and a Q of 25. The center frequency is 2210 Hz. For the reader who wishes to design his own filter, the formulas are included in the **appendix**.

Unfortunately, the use of an active filter, and its added performance, is offset due to the rather large additional cost of 1 per cent resistors.

The back-to-back diodes, across the audio input, are required to prevent front-end overloads. These diodes should be used if the active filter is not built. In that case, the diodes are connected across the common feed point.

construction

Construction of the PLL² is straightforward, with

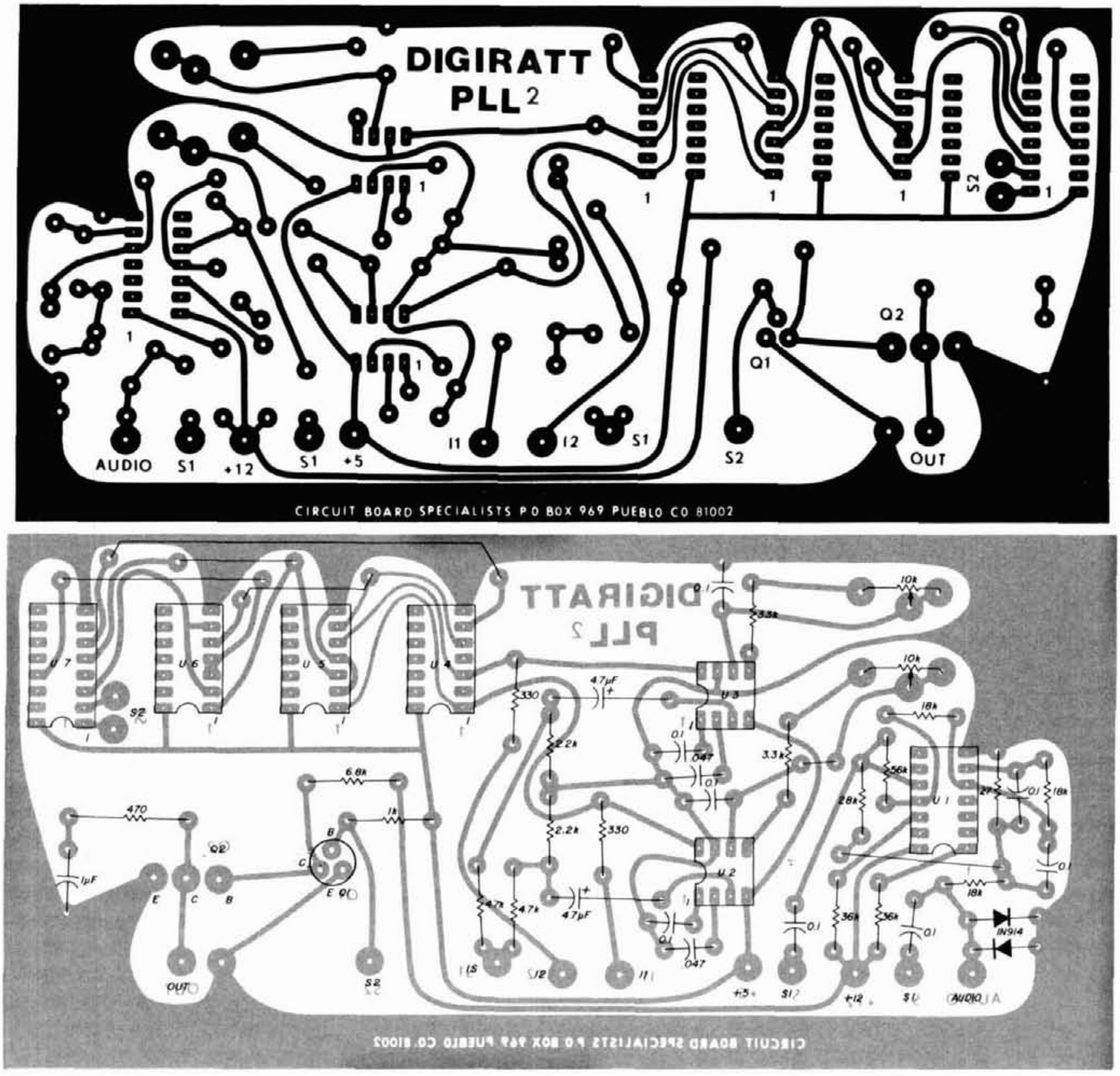


fig. 4. The circuit board for the PLL² is shown in A, with the parts placement diagram shown in B.

the entire circuit mounted on a single printed circuit board.* A copy of the printed circuit board and its parts placement are shown in **fig. 4**. The cabinet, which housed the prototype, measured 15 × 29 × 9 cm (5-7/8 × 11-3/8 × 3-1/2 inches).† Use shielded audio cable from the input connectors to the circuit board. For difficult RFI problems, you can apply 0.01-μF capacitors at the V_{cc} pin of each IC.

alignment

The Digiratt PLL[2] should be aligned as follows:

1. Apply a 2125 Hz, 1-volt, p-p sinewave into the audio input of the unit.

2. Adjust the decoder A until the mark-indicating LED illuminates.

3. Change the audio input to 2295 Hz and repeat the procedure for decoder B.

4. Reset the audio source for 2190 Hz and adjust decoder A until the mark LED goes out.

5. Reset the source for 2125 Hz and verify that the mark LED illuminates.

6. Again reset the source to 2190 Hz and verify that the mark LED goes out.

When the above conditions can be met, the mark portion of the circuit is aligned.

Decoder B is adjusted in the same manner, using 2220 and 2295 Hz. If you now sweep the frequency from 2050 to 2350 Hz, the mark LED should illuminate at 2125 Hz, remaining on until 2190 Hz. At 2220 Hz the space LED should light until you reach 2295 Hz. Between 2190 and 2220 Hz, neither LED should be be on.

*A complete kit of parts is available from Circuit Board Specialists, Box 969, Pueblo, Colorado 81002 for $31.10. The circuit board alone costs $7.50.

†Available from Radio Shack — RS270-282.

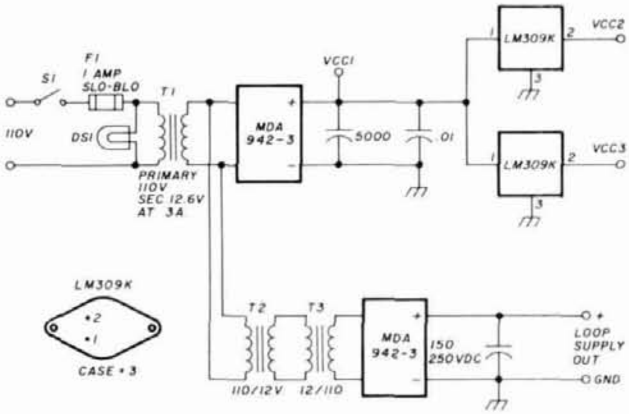
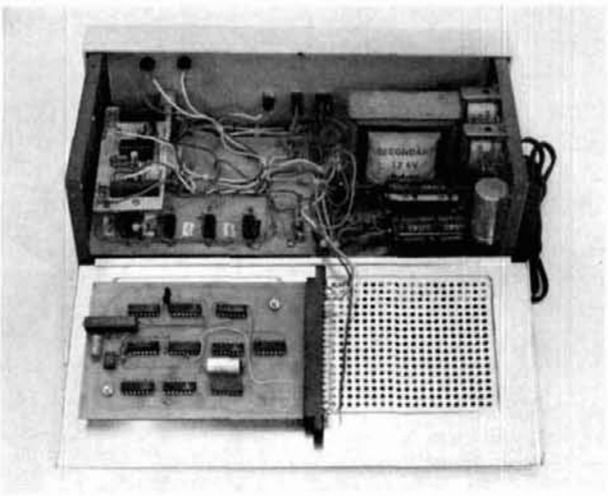


fig. 5. Power supply for the Digiratt PLL.[2] When mounting the LM309Ks, ensure that an adequate heat sink is provided.



In the foreground is the RY generator board (*ham radio*, January, 1978). From left to right, Prototype PLL twin decoders and Sel-comp logic, low voltage supply, 100 VDC, 100 mA loop supply is on the extreme right.

summary

The Digiratt PLL[2] is not the ultimate terminal unit, however the builder can expect very satisfactory results from it in all but the most adverse conditions.

The following persons have been most helpful to me over the past year or so during which I have been designing the various Digiratt projects: Don Smith, W9EPT; Bernie Holtman, W4GO; and Gus Bezy, K9FUI.

appendix

Design equations for a two-stage, bandpass filter using an LM3900 Norton amplifier.[2]

$$R1,\ R4,\ and\ R6 = \frac{25}{6.283 \times F_o \times C_1}$$

where F_o = center frequency in Hertz

C_1 = chosen value in Farads

$$R2 = R1 \times 1.5306$$

$$R3 = \frac{R1}{623.34}$$

$$R5 = 2R1$$

$$R7 = 3R1$$

$$R8 = \frac{R1 \times R7}{R1 + R7}$$

$C1,\ C2$ = Any convenient value

These equations will yield a two-stage filter with a gain of 23 dB and a Q of 25. Use 1 per cent resistors.

references

1. John Loughmiller, WB9ATW, "Digiratt — RTTY AFSK Generator and Demodulator," *ham radio*, September, 1977, page 26.
2. "The LM3900," *Linear Applications*, Volume I, National Semiconductor Corporation, Santa Clara, California, 1976.

ham radio

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8 Combination "S" and SWR METER; switches automatically.
9 Combination ALC control and NOISE BLANKER on/off switch.
10 DRIVE control for final stage.
11 *SQUELCH combination on/off switch and control.*
12 4-Position SELECTIVITY switch for SSB and CW.
13 4-Position MODE switch, automatic SSB Normal, Reverse, CW, and Lock (key down).
14 Combination push-pull POWER switch and AUDIO LEVEL control.
15 Combination RF ATTENUATOR on/off switch and control.
16 VOX GAIN control.
17 VOX DELAY control.
18 VOX ANTI-TRIP control.
19 *11-Position BAND SWITCH.*
20 MICROPHONE jack, hi-z input.
21 HEADPHONES jack.
22 RECEIVER OFF-SET TUNING SWITCH: 3-position: Max-Min-Off.
23 VOX-PTT SWITCH.
24 QSK (full break-in) SWITCH: 2-position: Fast-Slow.

More Details? CHECK—OFF Page 142

high-sensitivity preamplifier for frequency counters

Discussion of the design requirements for a counter preamplifier, which in addition to high sensitivity and input impedance also exhibits frontend overload protection

Frequency counter design has been greatly simplified since the introduction of the Intersil 7207/7208 and the recent 7216/26 integrated circuits. Several designs have appeared in *ham radio*[1,2] which make use of the 7207/7208 chip set with simple preamplifiers. Since a frequency counter's performance is largely limited by the preamplifier used to condition its input signal, this stage should receive significant attention during the design phase.

Such a preamplifier should have high input impedance, much like that of an oscilloscope vertical amplifier. It should also have enough sensitivity to permit the use of a X10 oscilloscope probe for minimum circuit loading, even at high frequencies. The preamplifier should be able to handle large input signals without overload, necessitating some form of input attenuator. Since the 7208 is a 5-MHz counter, the preamplifier should have a 50-MHz bandwidth for use with a prescaler. For proper counting of low-frequency signals with slow rise and fall times, the preamplifier should make use of a Schmitt trigger with hysteresis to prevent multiple triggering. A lowpass filter is also useful for counting noisy low-frequency signals.

preamplifier design

One method of achieving the high input impedance is to use an fet input stage followed by a broadband integrated circuit amplifier for high sensitivity.

By Paul Kranz, W1CFI, Mettacomett Path, Harvard, Massachusetts 01451

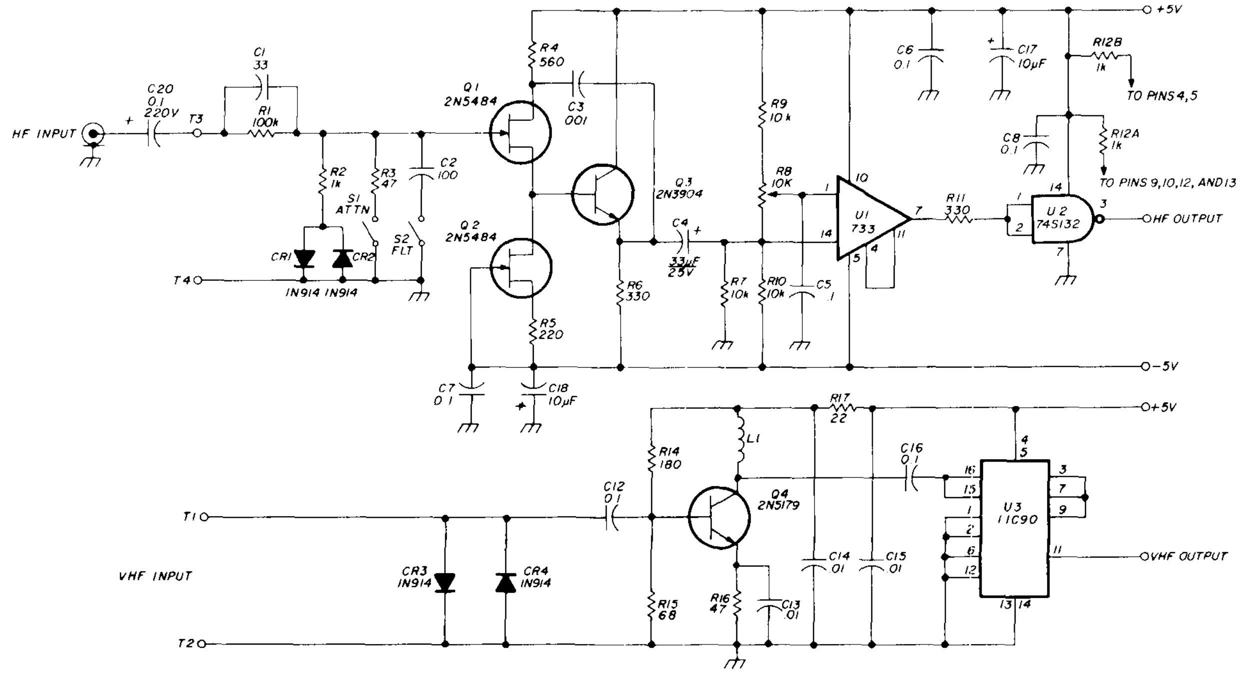

fig. 1. Schematic diagram of the high-sensitivity 0-50 MHz preamp and vhf prescaler. Input sensitivity ranges from less than 5 mV at 1 MHz to about 21 mV rms at 50 MHz. Note that the 733 uses the 14-pin DIP package. The filter (S2) is used to ensure accurate counting while measuring noisy low-frequency signals.

Overloading of the fet input stage can be prevented by a diode limiter and an attenuator. Since there are a variety of TTL integrated circuits available with Schmitt trigger inputs, one of these devices can provide the hysteresis and also the TTL signal conditioning. With the Schmitt trigger operating correctly, enough gain can be added ahead of it to provide sensitivity into the low-millivolt region.

The circuit that resulted from this approach appears in **fig. 1**. I've also included the vhf preamplifier and prescaler discussed by K4JIU.

The hysteresis and TTL signal conditioning are provided by a 74S132, which has a worst-case hysteresis of 0.8 volts. Therefore, I needed a preceding voltage gain of at least 100 to attain a sensitivity of a few millivolts. A 733 broadband amplifier seemed to be just

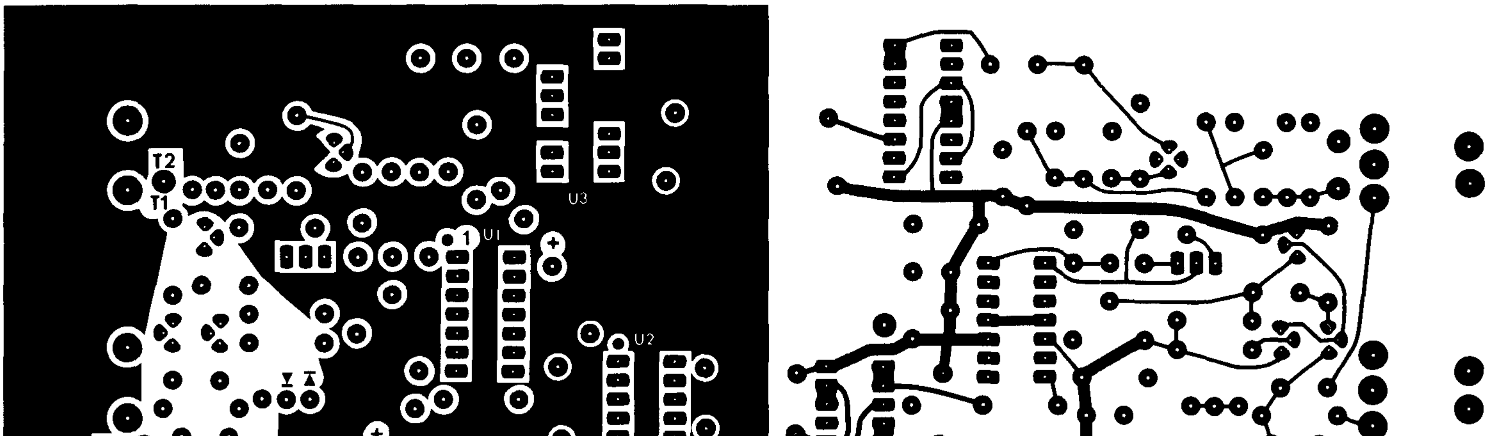

fig. 2. Circuit board layout for the preamp: bottom of board, right; component side, left.

the device, since it will provide a voltage gain of 400 to approximately 40 MHz.

During the breadboard testing, I found that the 733 would break into oscillation whenever I connected a X10 oscilloscope probe to its output. I was able to eliminate the problem by connecting a small resistor in series with the probe, thus reducing the capacitive probe loading on the 733 output. For this same reason, it seemed like a good precautionary measure to include some resistance between the 733 output and the 74S132 input. I chose R11 to be as large as possible and yet provide for proper sinking of the 74S132 input current by the 733 output under worst-case conditions.

The fet buffer amplifier, composed of Q1, Q2, and Q3, has the high-input resistance and low-input capacitance necessary for an oscilloscope-type input. In this stage, Q2 is a current source which offers several important benefits. First, it provides a high-source impedance for Q1 so that its voltage gain is nearly unity. Second, it serves as an active current sink to pull down the base of Q3 on negative-going half cycles of the input signal. Lastly, it provides a measure of temperature compensation so that the maximum signal swing is available over a wide range of operating temperatures.

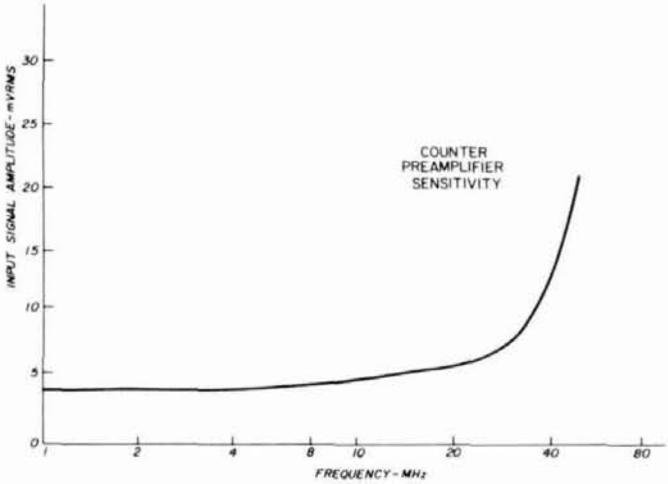

fig. 4. Input sensitivity curve for the counter preamplifier.

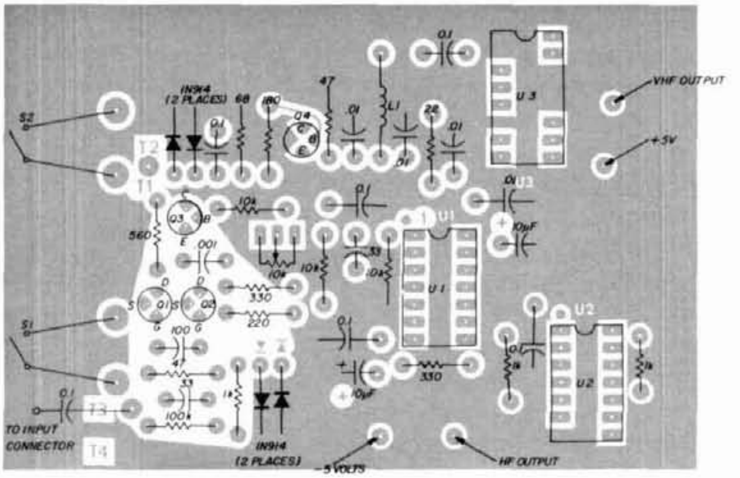

fig. 3. Parts placement diagram for the high-sensitivity preamplifier.

A significant reduction in the input capacitance of an fet preamplifier can be obtained by driving the input transistor's drain in phase with the input signal. This technique, implemented by C3 and R4, virtually eliminates the drain to gate capacitance of Q1, thus reducing its input capacitance by as much as 5 pF.

I spent most of my design time on the components in the gate circuit of Q1. These components control input capacitance, input resistance, overload characteristics, and lowpass filtering. At low frequencies, and for small signal amplitudes, the input circuit consists of only R1. The gate bias current is supplied by CR1, CR2, and R2. As the low-frequency input signal increases in amplitude, CR1 and CR2 begin to conduct and form a 100-to-1 voltage divider between R1 and R2, thus limiting the Q1 gate voltage. Lowpass filtering is provided by R1 and C2, which have a 16-kHz corner frequency. With high-frequency input signals, C1 compensates for the input capacitance of the Q1 gate components and keeps the voltage gain of the stage roughly constant. The price paid for this is the unavoidable lowering of the input impedance of the preamplifier as frequency increases. The R1-R2 attenuator rapidly loses effectiveness as the input frequency increases above 1 MHz. Again, this is due to the shunting effect of C1. An input attenuator, R3 and S1, solves this problem so that it is possible to connect as much as 60 volts rms directly to the counter input at 5 MHz and still obtain correct counting of the input signal. Without this attenuator, the preamplifier would saturate at 3 volts rms input at frequencies above 10 MHz.

construction

A printed circuit layout and component assembly appear in **figs. 2** and **3**. The two-sided printed circuit board contains the high-frequency preamplifier as well as the 500-MHz prescaler. Short conductor lengths and liberal use of bypass capacitors have kept the circuit stable and free of oscillations in the four preamplifiers which have been assembled.

The attenuator switch and filter switch are designed for printed circuit board mounting and may be diffi-

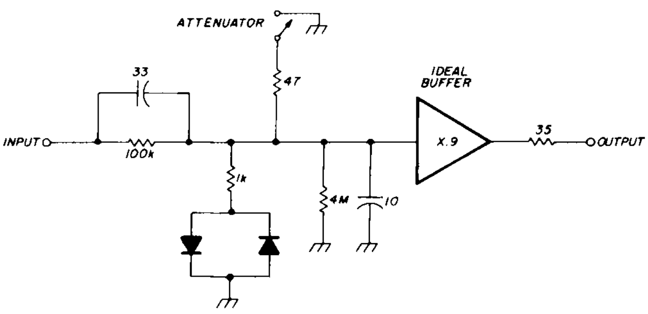

fig. 5. Equivalent circuit for the input of the preamplifier. At signal levels up to 50 mV rms, the amplifier has an equivalent input impedance of 4 megohms shunted by 10 pF.

cult to procure. However, there is no requirement that they be mounted this way. Actually, a decrease in input capacitance may be obtained by mounting them off the printed circuit board.

In order to reduce the input capacitance, C20 was mounted between T3 and the input BNC connector. A 1.5-megohm resistor may be connected between T3 and T4 to standarize the input to one megohm shunted by 10 pF.

The preamplifier will require approximately 70 mA from the +5 volt supply and 40 mA from the −5 volt supply.

performance

The only adjustment needed to get the preamplifier operating properly is the setting of the sensitivity potentiometer, R8. The adjustment is easily accomplished by applying a 30 to 50 MHz signal to the preamplifier input and adjusting R8 for a steady pulse train out of the 74S132. The signal can then be reduced and R8 readjusted. This process should be repeated until the adjustment of R8 produces maximum sensitivity.

A plot of the input sensitivity of the preamplifier appears in **fig. 4**. The sensitivity remains constant from a few Hz to 5 MHz and then begins to increase to 21 mV rms at 50 MHz.

I measured the input impedance from these threshold levels up to 50 mV rms and found it to be equivalent to 4 megohms shunted by 10 pF. An equivalent circuit of this input is shown in **fig. 5**.

As the input amplitude increases above 50 mV rms, the diode attenuator begins to lower the input impedance so that at amplitudes greater than 300 mV rms the input impedance is determined by R1, C1, and R2. It is possible to design an input network such that the input impedance at higher frequencies is still very high, but it would suffer from the lack of protection afforded by this design. **Fig. 6** is a plot of the maximum input signal for proper counting (attenuator off) as a function of frequency. The maximum low-frequency input of 140 volts rms is limited by the 1/4-watt dissipation of R1. At high frequencies, the input buffer will overload when the Q1 gate voltage reaches 10 to 15 volts pk-to-pk. Counting errors will occur when this level is exceeded. The input attenuator, to a point, helps relieve the overloading. However, as 50 MHz is approached, the input impedance due to C1 is only slightly greater than 100 ohms. The maximum input at 50 MHz would therefore be approximately 9 volts rms.

conclusions

This preamplifier, in conjunction with the K4JIU counter, performs admirably as an inexpensive laboratory frequency counter. The input impedance and sensitivity of the preamplifier worked out in practice to be as the design predicted and certainly adequate for most measurements. However, one thing did surprise me, the effect of input capacitance in lowering the input impedance at high frequencies. Although the preamplifier's input impedance is no worse than the typical input impedance of an oscilloscope, it still presents a very low impedance at 50 MHz.

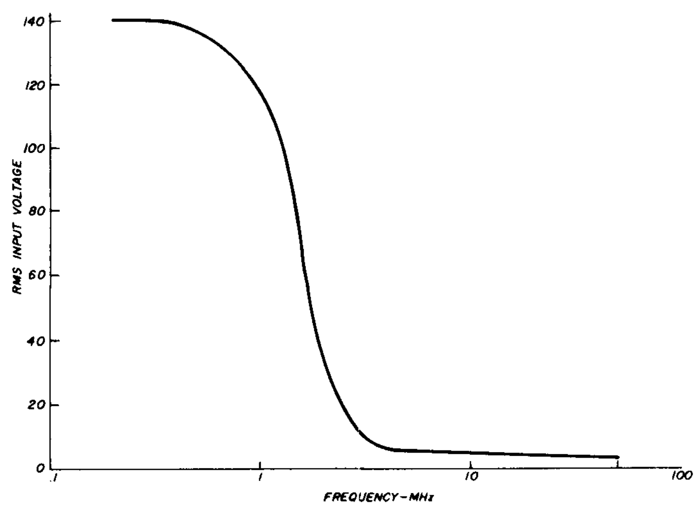

fig. 6. Maximum signal levels for an input impedance of 100k ohms and 33 pF.

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1. John H. Bordelon, K4JIU, "Simple Front Ends for a 500-MHz Frequency Counter," *ham radio*, February 1978, page 30.
2. Holton E. Harris, W1WP, "Simplifying the Digital Frequency Counter," *ham radio*, February 1978, page 22.

ham radio

twin-diode mixer — a new microwave mixer

A new microwave mixer using two diodes and half-wavelength lines yields an approximately 6 dB noise figure

This article describes a new microwave mixer, unique in that it has few parts and does not require boards or complicated metalwork. You can build it in a minimum of time, and with confidence of having a good mixer when you're done. The 1296-MHz model to be described has a 6.4 dB noise figure including a 1.2 dB i-f noise figure. Other features include the following:

1. A very low local-oscillator power requirement of −3 dBm

2. The local oscillator frequency is half that normally used

3. No dc return is necessary

4. There is no tuning

5. There is high isolation between all ports

mixer theory

A diagram of the ideal mixer is shown in **fig. 1**. The ideal filters pass currents only at the rf or i-f frequency, with the switch toggled at the normal LO frequency, $f_{rf}-f_{if}$. Thus, energy from an rf source is converted to the i-f and delivered to a load at the i-f port. There is no energy lost in the mixer, and the receiver's noise figure is that of the i-f.

In a real mixer, the switch takes the form of a diode which is turned on and off by the local oscillator. However, the diode is never a perfect open or short circuit, and as such will absorb some energy. Losses also occur in the circuitry surrounding the diode; the total loss depends in a complicated way upon the mixer circuit, the pump level, and, to a lesser extent, the diode itself. All high-performance mixers attempt to achieve the conditions of the ideal case shown in **fig. 1**.

Mixer performance can be characterized by the following equations:

$$T_{ssb} = (L_c - 1)\, T_0 \qquad (1)$$

$$T_{dsb} = (L_c - 2)\, T_0 \qquad (2)$$

$$L_c = \frac{i\text{-}f\, P_{in}}{rf\, P_{out}} \qquad (3)$$

By Jim Dietrich, WAØRDX, Post Office Box 208, Mulvane, Kansas 67110

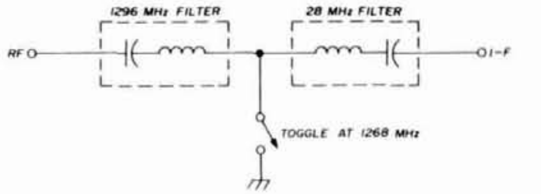


fig. 1. Diagram of the ideal mixer operating at 1296 MHz. The switch will toggle at a 1268-MHz rate, causing the 1296-MHz input to be converted to 28 MHz.

where

> T_{ssb} *is the ssb mixer noise temperature*
> T_0 *is the physical temperature of the mixer in degrees Kelvin.*
> T_{dsb} *is the double sideband mixer noise temperature.*

Note that **eq. 3** requires an input signal at the i-f frequency to directly measure conversion loss. The loss going the other direction is generally different. The loss of a dsb mixer is never less than 2, since i-f energy is equally converted to signal and image frequencies. However, as shown by **eq. 2**, the noise figure is not limited to 3 dB.

Eq. 2 is complicated by the problem of measuring the dsb mixer noise figure. The equivalent ssb noise performance in terms of the indicated noise figure is

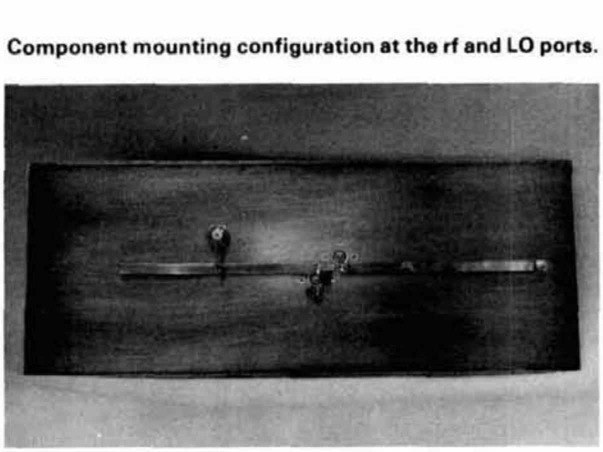
$$F_{ssb} = 2F_m - 1 \qquad \textbf{(4)}$$

where

> F_m *is the measured noise figure.*

For example, if the meter indicates 4 dB for a dsb mixer, the ssb noise figure is actually 6 dB, not the 7 dB obtained by simply adding 3 dB to the indicated noise figure.* This common mistake introduces substantial error for low noise figures. For each type mixer, ssb system noise temperature is given by the following:

Component mounting configuration at the rf and LO ports.



ssb mixer

$$T_{SYS} = (L_c - 1)\,T_0 + L_c T_{i\text{-}f} + T_{ant} \qquad \textbf{(5)}$$

dsb mixer

$$T_{SYS} = (L_c - 2)\,T_0 + L_c T_{i\text{-}f} + 2T_{ant} \qquad \textbf{(6)}$$

twin-diode mixer

An alternative to using a single diode is to use a pair of parallel-connected diodes, of opposite polarity, and pumped by a local oscillator at one-half the normal frequency. Each diode is turned on once during the LO cycle, 180 degrees apart, and both are off when the LO voltage is zero. Thus, a pair of diodes

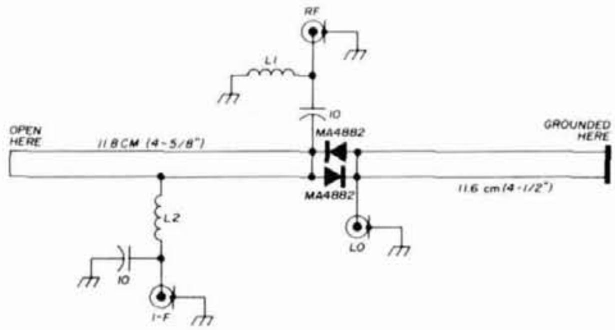


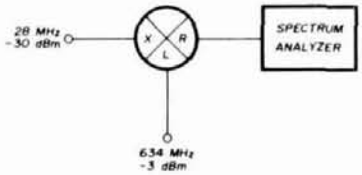
fig. 2. Schematic diagram of the twin-diode mixer. The half-wavelength lines are 5 mm (3/16 inch) wide and mounted 1.5 mm (1/16 inch) above the ground plane. Ensure that the line on the right is connected to the ground plane, while the one on the left remains open. Since the local-oscillator frequency is approximately one-half that of the input rf, the grounded half-wavelength line will look like an open circuit to the LO port and also like a low impedance to the rf port. L1 is 3 cm (1 inch) no. 28 AWG (0.3 mm) wire; L2 is 15 turns no. 32 AWG (0.2 mm) wire wound to a diameter of 1.5 mm (1/16 inch).

pumped at 634 MHz performs identically to a single diode pumped at 1268 MHz.

Fig. 2 shows a full-size mixer circuit for 1296 MHz which takes advantage of the frequency relation in the twin-diode scheme. On each side of the diode pair is a half-wavelength line at the rf frequency, and thus a quarter wavelength at the local oscillator frequency. At each frequency, an open circuit exists at the respective port, with a short on the opposite side of the diode pair. This ideally gives total isolation between the ports. The i-f and local oscillator ports may be dc or ac coupled, but the rf port must be capacitively coupled so that it presents an open circuit to the i-f. The bandwidth of the mixer is about 20 per

*With F_m equal to 4 dB, converting to a ratio will yield 2.5 ($antilog_{10} 4/10 = 2.5$). Using this number in **eq. 4** produces a noise figure of 6 dB [$10 \log_{10} (2 \cdot 2.5 - 1)$].

fig. 3. Test setup for the mixer conversion loss and isolation measurements. All ports are terminated in 50 ohms.



cent, so line lengths as well as component values are not critical.

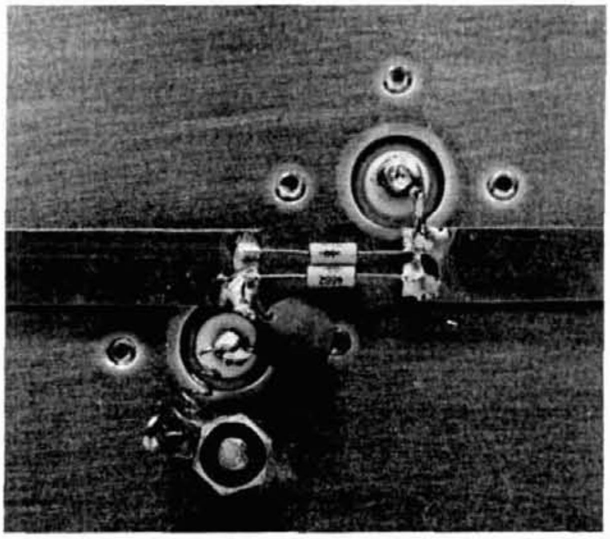
mixer evaluation

The mixer was tested by measuring the noise figure and also by measuring the actual conversion loss. The results were in good agreement. The conversion loss test setup is shown in **fig. 3.** The input level was -30 dBm ± 0.1 dB at 28 MHz. This level was used because more signal causes undesired higher-order products in the output, and less signal is difficult to accurately measure. The LO level is -3 dBm, which was found to be optimum both in the conversion loss and noise figure measurements. The spectrum analyzer was calibrated for absolute level at 1296 MHz so that the overall accuracy of the conversion loss measurement is ± 0.5 dB.

Fig. 4 shows the output as observed on the analyzer. All LO harmonics are 37 dB down (-40 dBm) from the input at 634 MHz. The desired signal and its primary image are both 6 dB (-36 dBm) below the 28-MHz input level. Other responses are down enough that they can be neglected.

As can be seen, the device is indeed a double-sideband mixer, so that ssb receiver noise temperature is found from the following:

$$\begin{aligned} T_e &= (L_c - 2)\, T_0 + L_c T_{i\text{-}f} \\ &= 2(297K) + 4(92K) \\ &= 962K \end{aligned}$$



View of the microwave mixer showing overall layout. Built on 3 mm (1/8 inch) aluminum this mixer used brass shim stock for the lines. BNC connectors were used at all ports.

where

$$L_c = \text{the conversion loss } 6\ dB = 4$$
$$T_0 = \text{the mixer operating temperature } 297K$$
$$T_{i\text{-}f} = \text{a } 1.2\ dB\ i\text{-}f \text{ noise figure } 92K$$

The single sideband noise figure is

$$F_{ssb} = 1 + T_e/290 = 6.4\ dB$$

Addition of a good input filter should lower the conversion loss to 3 to 4 dB and thus give an overall noise figure of 5 dB or less. However, care must be taken to keep filter losses low or else this improvement will not be obtained. In view of the fact that the mixer, as described, is probably better than most in use today, I haven't taken time to build a filter.

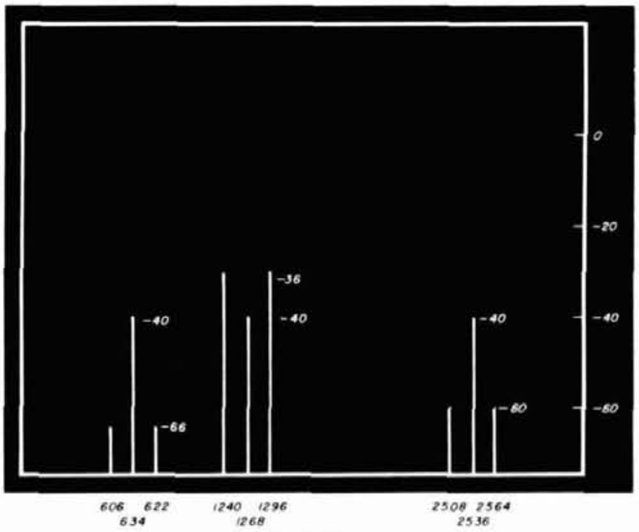


fig. 4. Mixer output as observed on the spectrum analyzer. The local oscillator was 634 MHz at -3 dBm. The 28-MHz i-f input was at -30 dBm. The output levels are read directly in dBm.

summary

In this article I have presented a new mixer configuration for use at 1296 MHz. The circuit can be used at higher microwave frequencies by simply scaling the half-wavelength lines. The device exceeds the performance of most available doubly balanced mixers by producing a 6.4 dB noise figure, nearly 40 dB isolation between all ports, and an LO requirement of only -3 dBm. In addition, the LO frequency is one-half that normally required, a most attractive feature.

A brief review of mixer theory, including noise performance, was presented to give a better understanding of twin-diode mixer operation. The noise relationships can be used to properly characterize receiver system performance using the twin-diode mixer or any other ssb or dsb mixer.

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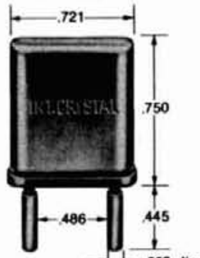

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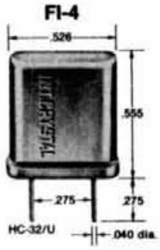

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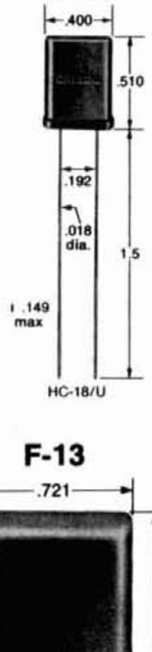

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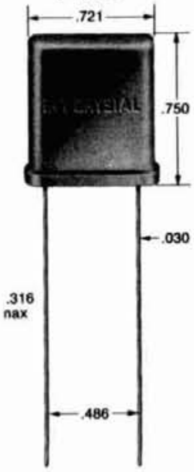

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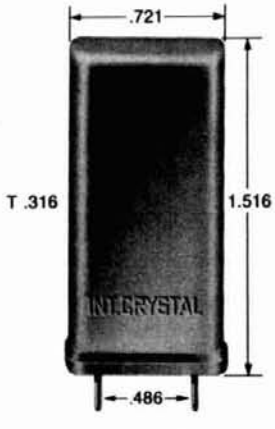

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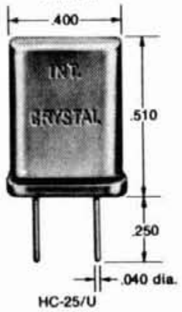

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two-meter preamplifier
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A simple, two-meter, one-transistor preamp is used to overcome the lack of receiver sensitivity

Does your 2-meter HT suffer from the lack of receiver sensitivity? A simple single transistor preamp can do wonders, especially if you use an external power amplifier while operating near the repeater fringes. One problem that frequently arises is that they can hear you, but you can't hear them!

I have a Regency HRT-2 hand-held, which I also use mobile with a Heath HA-201 10-watt amplifier located in the trunk. After a few months of operation, I became annoyed at the received signal drop-out and static near the fringe of the repeater coverage. My friends all told me my transmitted signal was excellent — full quieting into the repeaters; I tried a 1/4-wavelength whip in the center of my car roof, and finally a 5/8-wavelength whip, but to no avail. The noise and breakup persisted.

The HRT-2 has a receiver sensitivity of 0.7 μV for 20 dB of quieting, quite adequate in the city and compatible with the 2-watt output of the transmitter. However, as a mobile unit with 10 watts of output capability, the 0.7 μV sensitivity leaves much to be desired. A preamplifier seemed to be the solution.

simple preamplifier

There are a variety of circuits available for two meter-preamplifiers, but I wanted to keep it simple, low cost, and fairly compact, so it would fit inside the case if possible. Also, I felt it would be desirable to increase the receiver sensitivity from 0.7 μV to about

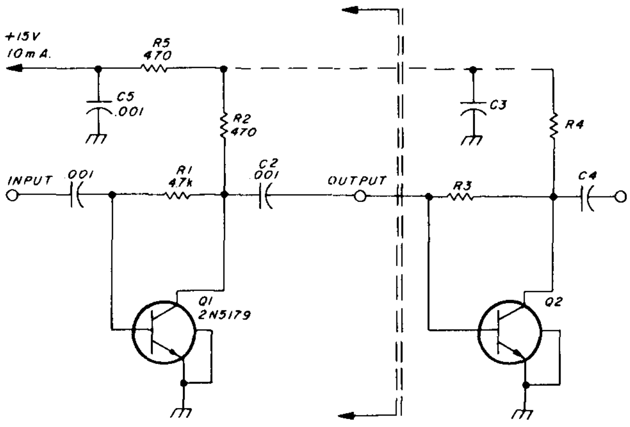

fig. 1. Schematic diagram of the wideband rf amplifier. Only the portion to the left of the dotted line is used. All resistors are 1/4 watt, and all capacitors are disc ceramics.

0.1 or 0.2 μV, typical of most high-quality mobile radios; this meant a gain of about 17 dB. Another requirement was that it be broadband, so that tuning for each repeater would not be necessary. The last requirement was that it have the capability of being switched into the circuit during receive and switched out during transmit without the need for relays.

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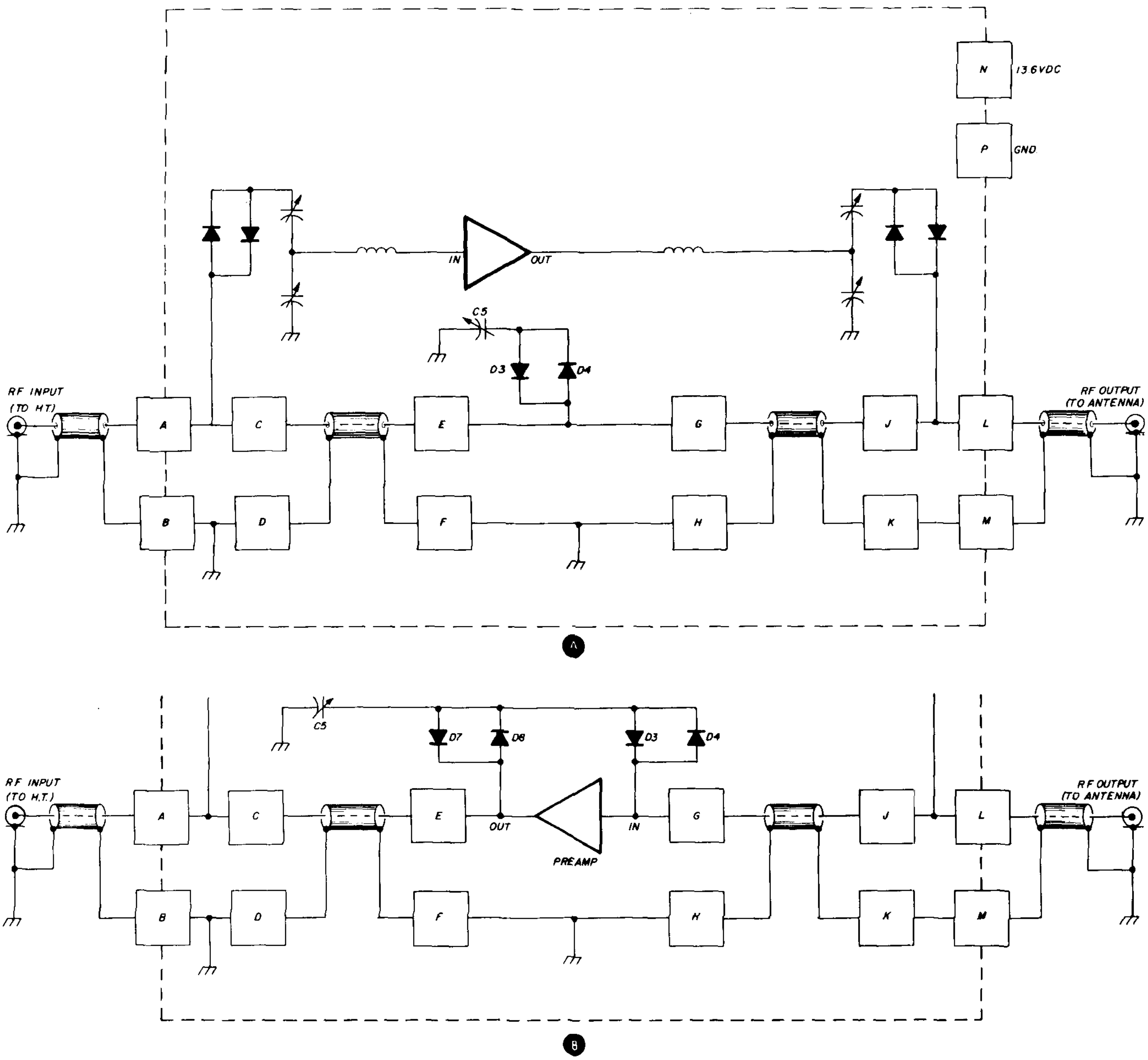


fig. 2. Schematic diagram of the Heath 2-meter amplifier (A) and the changes made to incorporate the receiving preamplifier (B). Power for the preamp can be obtained from the 13.6 volt line. D7 and D8 are also added to the existing circuitry.

A search through the literature revealed that all of these requirements would be tough to meet in one circuit. However, a simple and inexpensive wideband rf amplifier was found[1] which met most of the original objectives (**fig. 1**). Two transistors were used in the original circuit, but in my case only one was used with a gain of 14 dB. Although tuned circuits are recommended for improved selectivity, the amplifier was built up as a broadband unit to see how it would work. When externally connected to the HT the results were excellent. Clean reception was now possible with all the local repeaters, including two that previously were extremely noisy, and there was no evidence of overloading from undesired signals.

construction

The next problem was how to wire it into the HT. After a quick look inside the case, it appeared to be a major task to dismantle the entire set. And the thought of upsetting the rf circuits or breaking other connections was disgruntling.

A search for solid-state switching circuits did not reveal any that would be compact enough to fit inside the case. Suddenly, it occurred to me that the answer was inside the Heath power amplifier. It already contained a solid-state TR switching network, a quite clever one at that. I decided to place the preamp *within* the power amplifier, since all the required connections were right there.

The Heath 10-watt power amplifier is a single, rf-switched transistor. The amplifier is automatically coupled to the circuit as soon as one watt of signal appears at the input. A pair of switching diodes then conduct, routing the signal to the transistor. When the HT is switched to *receive*, the diodes no longer conduct, and the received signal is passed through two 1/4-wave transformers to the receiver. A pair of switching diodes at the junction of the two transformers provide a short circuit to ground during transmission, preventing any rf feedback.

To connect the preamplifier, it was merely necessary to break the center connection between the two 1/4-wave transmission lines, insert the preamp, and add another pair of switching diodes to the input of the preamp, leaving the existing pair at the output. The revised circuit is shown in **fig. 2**. The additional diodes were obtained from Heath, and are 1N4149 or equivalent. The 12-volt supply for the preamplifier was obtained from a convenient tie point in the HA-201. The preamp was mounted by its own leads, as close as possible to the coiled transmission cables, with connecting leads kept as short as possible. Mounting did not seem to be critical. However, I would recommend using plastic tape or other insulating material between the preamp and the case to prevent accidental shorting.

results

The results have been gratifying. Received signals are now clear and free from static and breakup at distances over 40 kilometers (25 miles) from the repeater sites. A slight readjustment of the HA-201 power amplifier was necessary to compensate for the capacitance effects of the added circuit, but there is no evidence of power loss with a wattmeter connected before and after the modification.

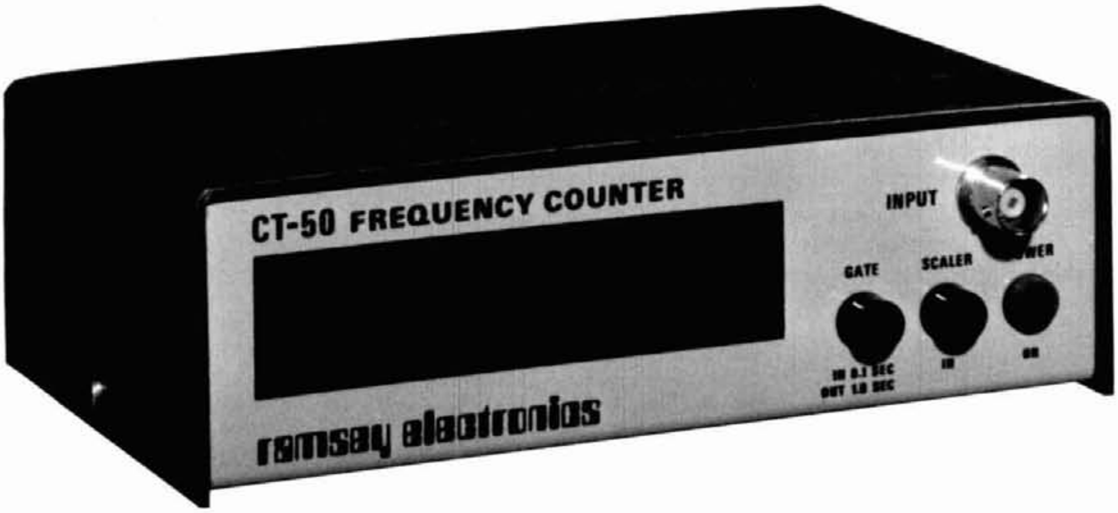
This solution may not work everywhere, particularly if there are strong nearby signals. If this should be a problem on some repeater frequencies, it may be possible to add a switch to short out the preamp when it is not needed. A little experimentation before modification of the HA-201 will probably determine the best arrangement for your own location.

references

1. Randall Rhea, WB4KSS, "General Purpose Wideband RF Amplifier," *ham radio*, April, 1975, page 58.

ham radio

Incredible...



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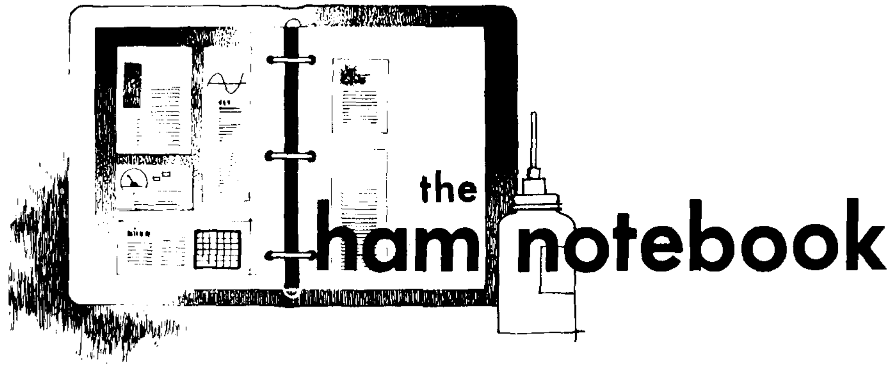

general coverage using the Collins 75S receiver

A recent article in *QST*[1] detailed a relatively simple and inexpensive method for extending the frequency coverage of the 75S-1. However, this particular method did not allow for proper operation of the receiver, especially with regard to transceive operation, since the correct tuned circuits for the preselector, rf, amplifier, and crystal oscillator circuitry are not necessarily selected. The method I've employed for some months does allow for split or transceive operation, is somewhat more flexible, and requires only a slight modification to the receiver.

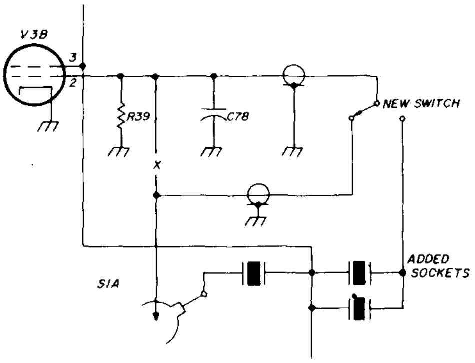

fig. 1. Schematic diagram of the change to the 75S-series receivers to permit general coverage while still maintaining transceiver capabilities. The two additional sockets are mounted on a small metal bracket above the present crystal bank.

An aluminum bracket is drilled to accept two crystal sockets and a *miniature spdt switch. One socket* accepts HC-17/U, and the other, HC-6/U crystals. The bracket is secured to the left side rail over the existing crystal sockets. The dimen*sions of the bracket allow its right* side to rest on the tops of the crystals in sockets 1, 2, and 3E, providing more rigidity.

The switch is mounted with the handle pointing left/right for instant recognition of NORMAL or GENERAL COVERAGE. Two solder lugs are attached under the socket nearest the switch for coax braid connections. The coax lead that normally goes from V3B, pin 2, to the arm of S1A, was broken at the V3 side and passed through the hole near the front panel to the new switch. A new piece of coax is run from the switch to V3B. The rest of the wiring is done with hookup wire. Attachment to the common point of the sockets may be made at crystal socket 3E.

With the switch in the NORMAL position, the 75S-1 operates with the standard compliment of crystals. In the GENERAL COVERAGE position, a properly chosen crystal may be inserted and *with the band switch selecting the proper frequency range*, operation outside the amateur bands (or extended 10-meter coverage, for instance) is accomplished. The bandswitch position is especially important when operating the receiver in transceive with the 32S series transmitters. A similar modification could be made to the transmitter, although this has not as yet been attempted.

Paul Pagel, N1FB

reference

1. Vernon L. Gibbs, W4JTL, "An extended Frequency Range for the Collins 75S-1," *QST*, October, 1977.

new product detector for the R-4C

As mentioned in a previous article,[1] the product detector in the Drake R-4B and R-4C leaves room for improvement. The present design allows the audio to leak back into the last i-f stage, from where it is detected, causing the AGC to vary at an audio rate. To correct this error we developed a reasonably simple product detector which eliminated the problems. Unfortunately, as stated in the article, the main disadvantage of the MC1496 was the high number of external components.

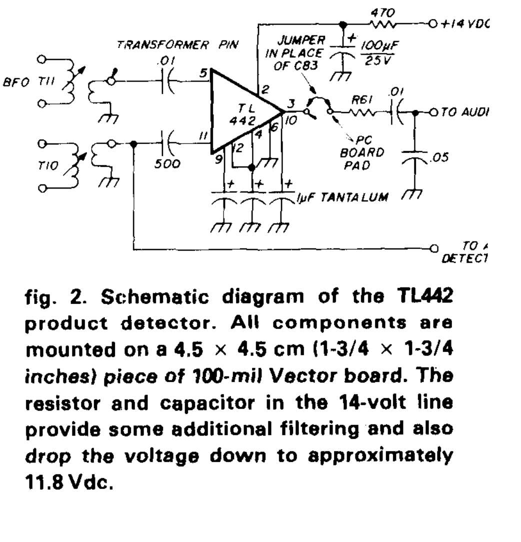

fig. 2. Schematic diagram of the TL442 product detector. All components are mounted on a 4.5 × 4.5 cm (1-3/4 × 1-3/4 inches) piece of 100-mil Vector board. The resistor and capacitor in the 14-volt line provide some additional filtering and also drop the voltage down to approximately 11.8 Vdc.

In recent correspondence with Howard Sartori, W5DA, he suggested another device which has also proved suitable as a product detector, the TL442 from Texas Instruments. As seen in **fig. 1**, the

*A parts package for this product detector is available from G. R. Whitehouse, Newbury Drive, Amherst, New Hampshire 03031.

circuit is extremely simple, yet provides essentially the same performance as the MC1496.*

To begin installation, it is first necessary to remove Drake parts CR2, CR3, C83, C84, and R60. Next, the wires connecting the output of T11 and the printed circuit board are removed. The 0.01 μF coupling capacitor to be installed should connect between the transformer pins and the IC socket. There shouldn't be any connections on the circuit board for either the BFO of i-f inputs. Completing the installation only requires that the IC, socket, and associated components be mounted on a small piece of 100-mil Vector board and mounted in the same location as the MC1496 version. All other connections can be made according to **fig. 1**.

Audio output is slightly higher than a stock R-4C. The combination of R61 and the original 0.05 μF bypass capacitor provide the proper high-frequency rolloff. In this configuration, and also in the original, the product detector will accept a 20 dB increase in signal level before it overloads.

As an addendum, several people have reported an audio oscillation problem after incorporating the 0.0015 μF capacitor referred to in the original article. We've found that this can be cured by inserting a 4700-ohm resistor in series with the added capacitor and also connecting a 0.01 μF capacitor across the headphone jack.

reference

1. J. Robert Sherwood, WB0JGP and George B. Heidelman, K8RRH, "Present-Day Receivers — Some Problems and Cures," *ham radio*, December, 1977, page 10.

Rob Sherwood, WB0JGP
George Heidelman, K8RRH
Sherwood Engineering

preprogramming the Kenwood TR7500

A Kenwood TR7500 was recently obtained for mobile usage, and has proven excellent for that purpose. It became apparent, however, that dialing up the commonly used frequencies could be, for this operator at least, hazardous while driving because of the need to watch the frequency read-out dial while changing channels.

Table 1. Diode programming information for the TR7500.

frequency	P1	P2	P3	P4	P5	P6
146.16	0	0	0	0	0	0
146.19	1	0	0	0	0	0
146.22	0	1	0	0	0	0
146.25	1	1	0	0	0	0
146.28	0	0	1	0	0	0
146.31	1	0	1	0	0	0
146.34	0	1	1	0	0	0
146.37	1	1	1	0	0	0
146.40	0	0	0	1	0	0
146.43	1	0	0	1	0	0
146.46	0	1	0	1	0	0
146.49	1	1	0	1	0	0
146.52	0	0	1	1	0	0
146.55	1	0	1	1	0	0
146.58	0	1	1	1	0	0
146.61	1	1	1	1	0	0
146.64	0	0	0	0	1	0
146.67	1	0	0	0	1	0
146.70	0	1	0	0	1	0
146.73	1	1	0	0	1	0
146.76	0	0	1	0	1	0
146.79	1	0	1	0	1	0
146.82	0	1	1	0	1	0
146.85	1	1	1	0	1	0
146.88	0	0	0	1	1	0
146.91	1	0	0	1	1	0
146.94	0	1	0	1	1	0
146.97	1	1	0	1	1	0
147.00	0	0	1	1	1	0
147.03	1	0	1	1	1	0
147.06	0	1	1	1	1	0
147.09	1	1	1	1	1	0
147.12	0	0	0	0	0	1
147.15	1	0	0	0	0	1
147.18	0	1	0	0	0	1
147.21	1	1	0	0	0	1
147.24	0	0	1	0	0	1
147.27	1	0	1	0	0	1
147.30	0	1	1	0	0	1
147.33	1	1	1	0	0	1
147.36	0	0	0	1	0	1
147.39	1	0	0	1	0	1
147.42	0	1	0	1	0	1
147.45	1	1	0	1	0	1
147.48	0	0	1	1	0	1
147.51	1	0	1	1	0	1
147.54	0	1	1	1	0	1
147.57	1	1	1	1	0	1
147.60	0	0	0	0	1	1
147.63	1	0	0	0	1	1
147.66	0	1	0	0	1	1
147.69	1	1	0	0	1	1
147.72	0	0	1	0	1	1
147.75	1	0	1	0	1	1
147.78	0	1	1	0	1	1
147.81	1	1	1	0	1	1
147.84	0	0	0	1	1	1
147.87	1	0	0	1	1	1
147.90	0	1	0	1	1	1
147.93	1	1	0	1	1	1
147.96	0	0	1	1	1	1
147.99	1	0	1	1	1	1

Users of the TR7500 should be aware that the transceiver has forty-four preprogrammed channels — all ARRL band-plan frequencies between 146 and 148 MHz, including all repeaters, and simplex frequencies. However, the transceiver also offers six blank channels, which are designed to be user programmed, by use of a diode matrix, for frequencies not included in the preprogrammed sequence. These frequencies must be on standard 30 kHz centers. Complete instructions for programming these additional channels, are in the transceiver operating manual.

The thought occurred to me that regular channels could also be programmed into the blank channels, rather than having to dial them out in the regular sequence. A review of the circuit and the programming instructions lead to a simple exercise in binary numbering, and a *complete* programming table was worked out. With this information, the six blank channels were quickly programmed.

The plan has worked out very nicely. The six channels are programmed for three repeaters, and three simplex frequencies, which completely handles local driving requirements. While driving, a quick glance identifies which of the six channels the radio is set on, with subsequent changes made by feel. Of course, any of the other regular channels is immediately available, simply by dialing up the appropriate channel in the normal manner.

Table 1 shows the complete diode programming instruction for all channels from 146.16 MHz to 147.99 MHz. Note that the columns are headed by designators P1 through P6, as used in the diode programming instructions of the operating manual.

Bob Locher, W9KNI

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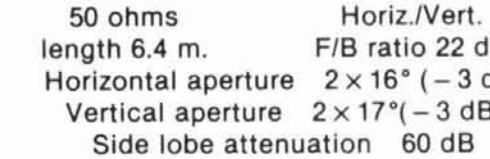

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NEW products

Drake R-7 receiver

The new Drake R-7 receiver is presently in the design and prototype stage, with first shipments scheduled for early 1979. Preliminary specifications are listed in **table 1**. The receiver is 100 per cent solid state, fully synthesized with a permeability tuned oscillator (PTO) for smooth tuning. It has continuous tuning from 0-30 MHz, and offers both a digital readout and an analog dial.

As with the Drake TR-7 transceiver, the R-7 receiver features up-conversion to a first i-f at 48 MHz; a special high-level, double-balanced mixer provides a high intercept point and strong signal handling characteristics. The receiver uses a full set of bandpass "window filters" that operate from 30 MHz, through VLF, to zero MHz. This permits performance in the MF/LF/VLF range that is very similar to that in the high-frequency range. As a result, external VLF preselectors or converters are not required.

The bandswitch selects various groups of window filters and determines the frequency limits of each range. Any 500-kHz segment within these limits is selected by simply depressing the UP or DOWN pushbuttons until the desired segment is reached. Tuning within the segment is accomplished by the PTO, which is connected to the main tuning knob.

A 10-dB, pushbutton-selected preamp can be activated on all ranges above 1.5 MHz. This preamp improves the overall sensitivity from approximately 0.5 μV to approximately 0.2 μV. As with any rf amplifier, however, its use lowers the intercept point by approximately the same amount as its gain. Therefore, preamp use should be limited to weak signal environments for best overall front-end performance.

The second i-f of the R-7 operates at 5645 kHz, and the selectable 8-pole crystal filters operate in this range. A choice of 300 Hz, 500 Hz, 1800 Hz, and 4.0 kHz filters are available, in addition to the 2.3-kHz ssb filter. Any of these filters may be selected from the front panel with a 5-position switch. It should be noted that the MODE switch operates independently of these filters, and can select either a special new synchro-phase a-m detector, or the product detector. Excellent international a-m shortwave and broadcast band reception can be realized with the low-distortion synchro-phase a-m detector.

The third i-f operates at 50-kHz and features a tunable i-f notch filter for heterodyne rejection. The notch depth is approximately 40 dB.

Extremely flexible selectivity combinations may be realized by the proper choice of an 8-pole crystal filter, notch adjustment, and positioning of the passband tuner, which is also employed in the R-7 receiver. The passband tuner is full range and enables the operator to properly set the passband position, in relation to the selectivity filter, for any mode continuously from RTTY to CW or any sideband. Various positions of agc, from OFF to SLOW, are also available from the front panel.

table 1. Preliminary specifications for the new Drake R-7 communications receiver.

Frequency coverage	0-30 MHz with DR-7 digital readout general coverage board; 0.5 MHz
	0.5-2.0, 2.5-3.0, 3.5-4.0, 4.5-5.0, 7.0-7.5, 14.0-14.5, 21.0-21.5, 28.5-29.0 MHz without Aux-7 (Aux-7 adds any eight 500-kHz segments from 0 to 30 MHz)
Frequency stability	Less than 100 Hz drift after warmup
Readout accuracy	Analog dial: better than ± 1 kHz when calibrated to nearest marker
	Digital: 15 ppm $\pm$ 100 Hz
Sensitivity (500 kHz - 30 MHz)	0.5 μV or less for 10 dB (S + N)/N on ssb and CW; 0.2 μV or less with preamp turned on
	2.0 μV or less for 10 dB (S + N)/N on a-m (30% modulation); 1.0 μV or less with preamp turned on
	(Preamp not operational below 1.5 MHz)
(0-500 kHz)	2.0 μV or less for 10 dB (S + N)/N on ssb and CW; 1.0 μV or less for 10 dB (S + N)/N on a-m
Selectivity	Same as TR-7 (ultimate selectivity greater than 90 dB)
Agc	Same as TR-7
Intermodulation	Intercept point at +20 dBm, minimum; two-tone dynamic range, 95 dB
Image and i-f rejection	Greater than 80 dB
Audio output	2.5 watt with less than 10% TGD into 4-ohm load
Power supply	110/220 Vac 50/60 Hz or 11-16 Vdc
Dimensions	Same as TR-7

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The R-7 will transceive with the Drake TR-7, and these functions are pushbutton controlled. The R-7 also has a unique antenna switch/toroidal splitter so that both the R-7 and the TR-7 may be used on the same antenna for simultaneous dual receive. This will be a boon to DXers who wish to monitor an out-of-band DXpedition and the in-band pile-up at the same time. The antenna selector also permits alternate antennas to be used on the receiver and a main antenna on the transceiver, or vice versa. The alternate antenna may also be split between the two units.

The receiver features receiver incremental tuning (RIT), so that the receiver frequency may be varied independently of the transmit frequency when operated in transceive with the TR-7. As with the TR-7, the digital readout in the R-7 may be used as an external counter to 150 MHz.

The receiver's built-in power supply operates from either 12 Vdc or 120/240 Vac. The styling, color, and size of the R-7 matches that of the TR-7, and either the internal speaker or an external MS-7 speaker may be used. Further information and prices will be available from the R.L. Drake Co. by the end of 1978.

Racal RA6700 receiver

The Racal RA6700 is a fully synthesized, tunable, solid-state communications receiver designed for all modes of reception over the frequency range of 15 kHz to 30 MHz. The internal synthesizer provides single-knob frequency control that allows rapid tuning across the complete frequency range with the feel and smoothness of a VFO, while retaining the accuracy and stability of the in-

More Details? CHECK—OFF Page 142

ternal frequency standard. In addition, the operator can select either a 100- or 10-Hz tuning rate, with the separate MHz control knob, permitting rapid changes from one end of the frequency range to the other.

The RA6772 incorporates the very latest techniques in mixer and signal path refinements to produce improved dynamic range and to reduce intermodulation products, reciprocal mixing, cross modulation, blocking, and spurious responses to a degree that exceeds the capabilities of any other general-purpose receiver currently in production. The basic receiver will accommodate six i-f filters; two may be asymmetrical ssb filters with either 3- or 6-kHz nominal bandwidths. Provisions are also available for additional filters if the standard 300-Hz, 3-kHz, and 8-kHz filters are not adequate. During CW reception, the internal BFO provides ± 3 kHz tuning range.

The receiver is ruggedly constructed to permit operation under extreme conditions. Yet the internal layout permits easy access for servicing, all components being accessible without the use of extension leads or adapters. The rear panel contains all input and output connectors, with many internal connections making possible the use of the receiver as the basis for a more sophisticated receiver system.

To enhance receiver versatility and flexibility for communications, surveillance, and direction-finding applications, a number of options and variations are available. In addition to the normal modes of reception provided in the basic RA6772, additional units can be added within the receiver to permit reception of ISB and FSK signals, along with AFC operation. When configured for teletype, one or two machines may be directly connected to the receiver without the need for external power supplies.

The RA6700 receiver series is well suited for single- or multiple-receiver systems, with numerous options and configurations possible. As an ex-

More Details? CHECK—OFF Page 142

ample, the RA6774 can be controlled by a computer, or, as in the case of the RA6780, by either local or remote manual control. For additional information, contact RACAL Communications, Inc., 5 Research Place, Rockville, Maryland 20850.

Sherwood Engineering crystal filters

Sherwood Engineering has recently expanded still further its already extensive line of high-performance crystal filters. To complement their filters for the Drake R4C, they've now added a 2.1-kHz a-m filter (CF-2.1K/8AM) which plugs directly into a normal a-m filter socket. This 8-pole ladder filter, which can be used to replace the normal 4- or 6-kHz filters, exhibits a −6dB bandwidth of 2.1 kHz and is 3.6 kHz wide at the −60 dB point.

To help you take advantage of the extensive filter capability that can be obtained by using the full line of Drake-type filters, Sherwood Engineering is now offering a custom-made, dual function switch for the front panel of the R4C. This switch, which replaces the present AGC switch, makes it possible to switch each filter from the front panel. In addition, the new concentrically mounted AGC switch provides five AGC positions, instead of the original four (off, fast, medium, and slow). This offers the operator the option of incorporating an additional AGC speed for greater time-constant flexibility. The switch itself does not replace the Sherwood Engineering relay kits, but is offered as an alternative to the normal toggle switches.

In addition to filters for the Drake R4C, Sherwood Engineering is also manufacturing the CF-350/8, a 350-Hz CW filter for use with the TR4, TR4C, or the TR4Cw. This 8-pole filter has a shape factor of 2.43:1 (as compared with the 4:1 factor for the normal 500-Hz filter supplied in the TR4Cw), yet it is easily installed in many TR4s in less than two hours.

For the TR4Cw, installation can be accomplished in less than five minutes.

Other new products include 350-Hz CW filters for both the Kenwood TS-820 (CK-350/8) and the Signal One (CS-350/8). Each is a custom-made, 8-pole ladder filter which exhibits a 6/60-dB shape factor of 2.43:1. Both are direct replacement filters which can be installed in minutes. The CS-350/8 has a lower insertion loss, in many cases, than the hard-to-find deluxe factory CW filter.

For the CW operator who wants the best, Sherwood Engineering has developed a 125-Hz CW filter for the new Drake TR-7. This filter is excellent for the high selectivity demanded during DX and contest work. The new filter will mount directly on the TR-7 filter board. Or, for the person with both Drake CW filters already installed, a plug-in filter board can be obtained to add the additional flexibility afforded by incorporating all filters.

For complete details and price information on all products, contact Sherwood Engineering, 1268 South Ogden Street, Denver, Colorado 80210.

Alliance heavy-duty rotator

The HD-73 Heavy-Duty rotator, combining wind- and ice-resistant features plus two-speed rotational control never before incorporated into a unit of its size and performance, is offered by the Alliance Manufacturing Company Inc., of Alliance, Ohio.

Designed especially for the serious amateur who wishes to increase his capability with in-tower or mast-mounting option, the HD-73 features a unique dual-speed control with one five-position switch. It provides a one-minute-per-revolution speed for

More Details? CHECK—OFF Page 142

rotating over an extended arc, and a slower speed permitting pinpoint adjustments for the best signal on receive and transmit.

Improved automatic brake action not only simplifies positioning, but also reduces risk of antenna damage by sudden stops that impose high inertial stress on antenna, tower, and rotator.

Designed to move antennas with a maximum of 1 square meter (10.7 square feet) of load capacity, the HD-73 develops a wind-load bending moment capable of withstanding the most severe prevailing wind conditions. Icing, another weather hazard for rotators, is overcome by the heaviest hardened-steel pitch-gear teeth of any rotator in its size and price range. The consistently high performance of the unit in all weather conditions is enhanced by a factory-installed lubricant that withstands temperature ranges of +49 to −29°C (+120 to −20°F).

The HD-73's 20-volt ac, capacitor-operated, split-phase, reversible motor and its transformer are doubly protected by fuse and thermal protectors against shorts, possible connection error, and prolonged operation. No voltage on the motor or leads exceeds Underwriters Laboratory safety limits.

The meter, a dc, D'Arsonval, taut-band type, is calibrated in bold S-W-N-E-S lettering as well as with a degree-graduated scale for full 360° position recording. The voltage supply for meter indication is solid-state regulated to assure accuracy regardless of wide line-voltage or load variation. The bar switch permits dual-speed rotor control with utmost accuracy and fingertip ease.

The power required is 117 volts ac, 60 hertz. The mast mounting size range is 35 mm (1-3/8 inch) to 63 mm (2-1/2 inch) O.D.; it requires a six-conductor cable. Total shipping weight of the rotator with two pairs of brackets and control box is 7.7 kg (17 lbs).

More Details? CHECK—OFF Page 142

See your nearest Alliance Distributor, or write to Alliance Manufacturing Company, Inc., Alliance, Ohio 44601.

compact amateur handheld from Standard Communications

A compact new 1-watt, 2-meter handheld amateur fm transceiver is now available from Standard Communications Corp. of Carson, California. This transceiver, designated C-118, is approximately the height and width of a dollar bill, and permits the user to transmit up 600 kHz, down 600 kHz, or receive and transmit on the same frequency with just one crystal. This provides 18-channel capability with only six crystals.

The C-118 also incorporates a built-in capacitor microphone and LED status lights for CHANNEL BUSY and TRANSMIT. Also included at no additional charge is a BNC connector with flexible antenna, provisions for an external dc power supply, and earphone. It has a frequency range of 144-148 MHz and comes equipped with one crystal for operation on 146.94 simplex and 146.34/94 MHz.

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To obtain a free copy of the C-118 data sheet, write Standard Communications Corp., P. O. Box 92151, Los Angeles, California 90009.

500-watt rf transformer

Palomar Engineers has a new broadband rf transformer. It matches vertical and mobile antennas to 50-ohm coaxial cable. Impedance values of 8, 12.5, 16, 22, 32, and 50 ohms can be selected by a convenient switch.

The transformer is mounted in a die-cast aluminum case 10 × 12 × 5 mm (4 × 5 × 2 inches) fitted with UHF (SO-239) connectors. The rf ferrite toroid core is wound with teflon-insulated wire and is rated at 500 watts in continuous commercial service. Operating frequency range is 1-30 MHz (1-10 MHz below 20 ohms).

Price is $35 plus $2 for shipping in the United States and Canada. For a free descriptive brochure, write to Palomar Engineers, Post Office Box 455, Escondido, California 92025.

Stolen Equipment

STOLEN: June 30, 1978, Rochester, NY from K2DHA. 2-meter transceiver, KDK model FM144-10SXRII Serial #5670. Scanner adapter attached to above. Amateur Wholesale Electronics, FMSC-1, no serial #. Amateur Wholesale Electronics Touch Tone Pad, Model FMTP-1. If tendered for trade, sale, or service, please notify: K2DHA, A. C. Peed, L66 Monteroy Road, Rochester 14618 or: Brighton Police Dept., 2300 Elmwood Ave., Rochester, N.Y. 14618.

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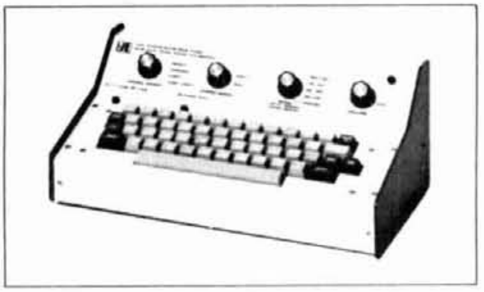

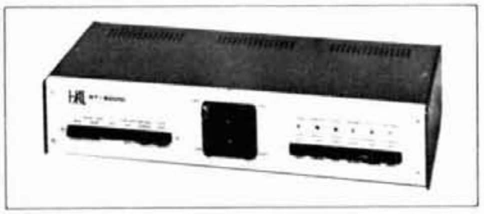

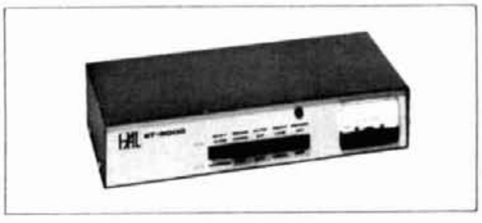

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Write today for HAL'S new catalog and RTTY guide and discover how much fun RTTY can be.

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GIANT KILLER!

ATLAS 210x/215x

5 BAND - 200 WATT - ALL SOLID STATE HF SSB CW TRANSCEIVER

Don't let its small size, and light weight fool you . . . the Atlas 210x/215x is a top notch performer, with all the power and performance that you find in rigs twice as big, and costing twice as much. And none of the others have as many superior features as our little Giant Killer, regardless of their size.

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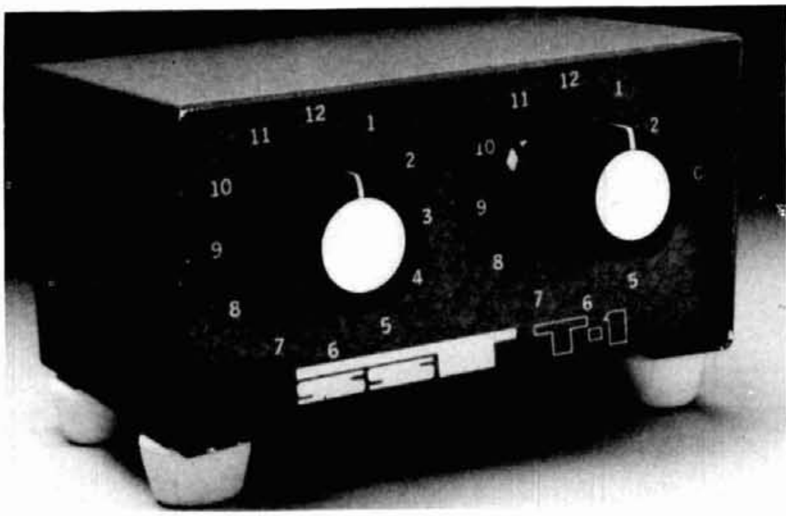
SST T-1 RANDOM WIRE ANTENNA TUNER



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

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

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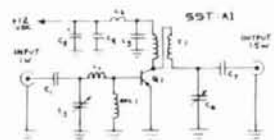

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CALL PINS 3 lines 1-1/4 × 3-1/4 $1.55 each. Call — First Name and Club. Colors black, red or blue with white letters. (Catalog) Arnold Linzner, 2041 Linden Street, Ridgewood, N.Y. 11227.

I WOULD LIKE TO TRADE my Western Data Systems 6502 based computer for an SSB xmtr, SSB xcvr, or 2-meter FM xcvr. WB6WDI, P.O. Box 234, Mt. Shasta, CA 96067.

HAM RADIO MAGAZINES — Vol 1 No. 1 through today (complete set) in HR binders — Mint! $225 or trade for model railroad brass. Steve Hyett, 1440 Royal St. George, Naperville, IL 60540.

Coming Events

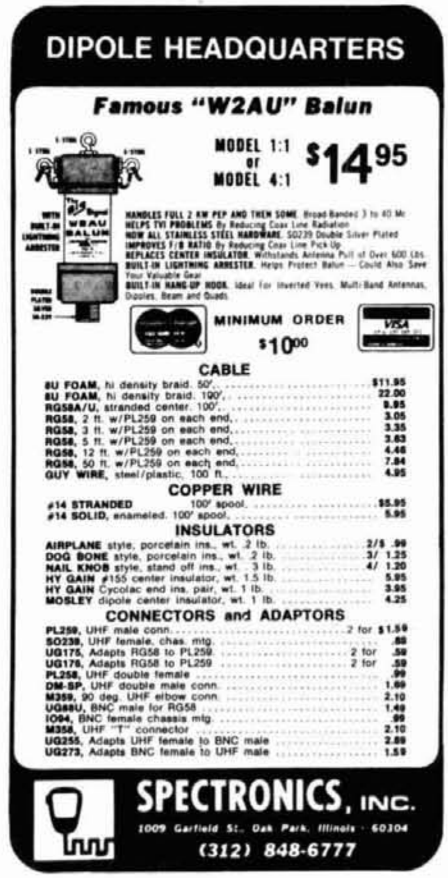
OCTOBER 1 — Talk-in WR4ADH — 146.34/146.94. North West Georgia ARC Hamfest Fairgrounds, Rome, GA WA4IBI Scott Lomax, (404)278-2581.

MICHIGAN: R.A.D.A.R. The Repeater Association of Downriver Amateur Radio is holding its 2nd annual swap & shop on Sunday, Oct. 22, 1978 at Kennedy High School in Taylor, Michigan. Located on Northline Rd. East of Telegraph Rd. (U.S. 24). Admission $2.00. For info write R.A.D.A.R. Inc., P.O. Box 1023, Southgate, Michigan 48195.

MASSACHUSETTS: Hampden County Radio Association's annual Ham and Electronic Equipment auction, Friday, October 6, Feeding Hills Congregational Church, Feeding Hills. Doors open — 7PM — Auction 8PM. For info Larry Soltz, WB1CJH, (413)567-6707.

MARYLAND: The Foundation for Amateur Radio's annual Hamfest, Gaithersburg Fairgrounds, Gaithersburg, Sunday, October 8. Flea market, food, exhibits, ladies and children's programs. Picnic grounds and free parking. Fee: $2.00, flea market space $5.00 each, commercial $15.00 each. Pre-registration required prior to October 4. Talk-in provided. For info Ron Levin, W3GBU, 802 Greenview Court, Reistertown, MD 21136. (301)833-1816.

NEW YORK: LIMARC Hamfest, Islip Speedway, Islip Avenue (Rt. 111), Islip, Long Island, October 15. Gate opens 9:30 AM. Admission: $1.50. Ladies and children under 12 free. Sellers and exhibitor's spaces $3.00 each. Limarc tune-up clinic (bring your own power cord). Talk-in 146.25/85 or 52. For info, Hank Wener, WB2ALW, 53 Sherrard St., East Hills, NY 11577 — (516)484-4322 evenings or Ken Denston, WB2RYC (516)379-6463 evenings.



More Details? CHECK—OFF Page 142

NEW JERSEY: Knight Raiders VHF Club Auction & Flea Market, St. Joseph's Church, East Rutherford, Saturday, October 14. Doors open 10:00 AM. Free admission/parking. Flea market tables — $6.00/full or $3.50/half. Talk-in 146.52 and 144.65/145.25. For info: Bob Kovaleski, (201)473-7113 or Bob Czyzewski (201)791-5651. Evenings only.

INDIANA: Marshall County ARC's 3rd annual Swap and Shop Hamfest, the Armory, 11th and West Madison Sts., Plymouth, October 29, 7:00 to 4:00. Donation $2.00. Free tables. For info and res. tables: Melvin Mahler, P.O. Box 151, Plymouth, IN 46563.

MISSISSIPPI: Gulf Coast Ham/Swap Fest, Sunday, October 22, International Plaza, west end of Biloxi/Ocean Springs bridge, Highway 90, Biloxi. Donation: $1.00. Tables $2.00. Talk-in 146.13/73 and 146.52. For info, advance tickets and tables: Irvin L. Kelly, K5YIN, 116 Wiltshire Blvd., Biloxi, MS 39531. (601)374-3340.

MISSOURI: Mo-Kan Council of ARC's ARRL Convention, Hilton Airport Plaza Inn, Kansas City. October 13, 14, and 15. Exhibits, Ladies' program, luncheon, fashion show, Saturday night banquet ($12.00/person). Pre-registration $3.00. Checks to Mo-Kan Council of ARCs, P.O. Box 704, Kansas City, MO 64141.

NEW YORK: Radio Amateurs of Greater Syracuse 14th annual Hamfest, New York State Fairgrounds, Arts and Home Center, Syracuse, Saturday, October 7. 9AM to 6PM. Talk-in 90/30 — 31/91. Exhibits, indoor/outdoor flea market, ladies programs. Tickets before October 1 — $1.50. $2.00 gate. Under 12 free. Overnight and trailer parking available. For info R.A.G.S., P.O. Box 88, Liverpool, N.Y. 13088.

ONTARIO: London Amateur Radio Club wil hold its 10th annual RSO Convention October 13th, 14th and 15th at the Downtown Holiday Inn City Center Tower, London, Ontario. A Friday night Oktoberfest-type Eyeball is FREE to all registrants and their spouses. Events and programs for the weekend include contesting, antennas, DXing, CW-FM-RTTY-ATV-SSB discussions, technical topics, computers and AMSAT. R.S.O. — CARF — CRRL forums, and DoC discussion. Saturday night banquet, prizes, and dancing to the big band sounds. Sunday flea market, and much, much more. Talk-in (VE3RSO) 75 ssb, 3775 kHz; 2-meter FM 146.46/147.06 (VE3LAC). For more information write London Amateur Radio Club, Inc. Attention: Convention Tickets, P.O. Box 82, Station B, London, Ontario N6A 4V3.

NEW YORK: The annual United States Air Force Military Affiliate Radio System (USAF MARS) Region One Convention will be held in Albany, New York on October 13-15, 1978. For more information write Convention Committee, P.O. Box 1978, Boiceville, N.Y. 12412.

NEW JERSEY: Livingston ARC Annual Fleamarket, Saturday, October 14, 1978 from 10AM until 4PM at the Fairfield United Methodist Church, corner of Plymouth and Horseneck Road, near Route 80 and only one block from Route 46. Registration $4 per car space; buyers and lookers free. Refreshments. For more information, write LARC, 116 Orton Road, W. Caldwell, N.J. or call (201)226-7943.

ISLAND DX AWARD: Sponsored by Radio Amateurs residing on Whidbey Island, the IDX Award is available to all Radio Amateurs and SWLs of the world who can meet some simple requirements: QSL confirmations from 50, 100, 150, or 162 (max possible) islands *including* Whidbey Island. Special band and mode endorsements are added features of this attractive award certificate. since not all islands are qualified contacts, each Amateur should have the IDX special Island Listing and a copy of the rules. As an added incentive, a special IDX wall plaque will be awarded to the first Radio Amateur who confirms the maximum possible number of recognized islands. Please send large business-size SASE to Bill Gosney, WB7BFK, 2665 North, 1250 East, Whidbey Island, Oak Harbor, Washington 98227. Foreign amateurs please include 5 IRCs.

JAMBOREE ON THE AIR: annual gathering of Scouts, former Scouts and interested hams on the Amateur Radio bands reaches a milestone in October — it's twenty-first birthday! The jamboree-on-the-air will be held over the weekend of 21st and 22nd October starting at 0001 LOCAL time on Saturday, October 21, 1978 and terminating 2359 LOCAL time on Sunday, October 22, 1978. However, each station is permitted to select its own operating schedule, including Friday evening, if desired. Frequencies suggested: Phone: 3940, 7190, 14290, 21360, 28990 and 52500 kHz. CW: 3590, 7030, 14070, 21140, 28090 kHz. SSTV: normal SSTV frequencies. The World Scout Bureau plans to operate from a special camp about 15 km. from Geneva, Switzerland, and will use the callsign HB9S/portable. They plan to be on all bands simultaneously and all modes including SSTV, RTTY, and OSCAR. For more information, write: Harry A. Harchar, W2GND/K2BSA, Boy Scouts of America, North Brunswick, N.J. 08902; telephone (201)249-6000.

More Details? CHECK—OFF Page 142

YOUR BEST BUY IN KITS FREQUENCY COUNTERS

6 GOOD REASONS FOR BUYING A HAL-TRONIX FREQUENCY COUNTER

(1) 100% COMPLETE KIT, (2) EASY ASSEMBLY, (3) COMPLETELY ENCLOSED IN METAL CABINET, (4) IC SOCKETS USED THROUGHOUT FOR EASY TTL REPLACEMENT (5) EASY ON YOUR POCKET BOOK, AND (6) NO EXPENSIVE CHIPS TO REPLACE (EXAMPLE — IF YOU LOSE A DECODER, LATCH OR DRIVER IN A HAL-TRONIX COUNTER, THE AVERAGE COST OF REPLACEMENT OF THE LOW-COST TTLS IS LESS THAN $1.00 EXCLUDING THE PRE-SCALE CHIP. IN SOME OF THE NEWER COUNTERS NOW BEING MARKETED BY MY COMPETITION, THEY ARE USING THE EXOTIC SINGLE CHIP AND WOULD COST YOU CLOSE TO $30.00 TO REPLACE). THIS IS SOMETHING YOU SHOULD CONSIDER.

ANALOG-DIGILAB KIT $139.50

DESIGNED BY HAL-TRONIX AND MIKE GOLDEN OF R.E.T.S. ELECTRONICS SCHOOL OF DETROIT. FOR RUGGED CLASSROOM USE.

FOR THE RADIO AMATEUR, STUDENT, EXPERIMENTER OR DESIGNER
SPECIFICATIONS: OUTPUT VOLTAGES: +5V, +12V, −12V; USABLE CURRENT: 750mA; % Regulation at 500mA: 0.2%; Short-circuit limited at 1.0 amp; Thermal overload protected. Power requirements: 117VAC, 60HZ, 40 Watts.
Function Generator: Frequency range: 1HZ to 100 HZ in 5 bands. Amplitude adjustable from 0 to 10 VPP. DC offset adjustable from 0 to ± 10V. Waveforms: Sine, square, triangular and TTL Clock. TTL Clock 0 to +5V level, 200 ns rise and fall time. Frequency determined by Function Generator. Output impedance 1.2K ohm.
Most of all, it's easy to construct and service. PC boards are predrilled, plated thru and solder flowed. Over 1000 units sold to schools.

NEW FROM HAL-TRONIX FIRST TIME OFFER

SIX-DIGIT ALARM CLOCK KIT for home, camper, RV, or field-day use. Operates on 12-volt AC or DC, and has its own 60-Hz time base on the board. Complete with all electronic components and two-piece, pre-drilled PC boards. Board size 4" x 3". Complete with speaker and switches. If operated on DC, there is nothing more to buy.*
PRICED AT $16.95
Twelve-volt AC line cord for those who wish to operate the clock from 110-volt AC. $2.50
*Fits clock case advertised below.

6-DIGIT CLOCK 12/24 HOUR

COMPLETE KIT CONSISTING OF 2 PC G10 PRE-DRILLED PC BOARDS, 1 CLOCK CHIP, 6 FND 359 READ-OUTS, 13 TRANSISTORS, 3 CAPS, 9 RESISTORS, 5 DIODES, 3 PUSH-BUTTON SWITCHES, POWER TRANSFORMER AND INSTRUCTIONS INCLUDED. DON'T BE FOOLED BY PARTIAL KITS WHERE YOU HAVE TO BUY EVERYTHING EXTRA.
PRICED AT $12.95

ATTENTION RADIO CLUBS

For club or group projects, request FREE information about our DISCOUNTS on any of the HAL-TRONIX kits. Discounts range from 10-25%, depending upon the quantity needed.
We are experienced in supplying kits in volume quantities to schools, laboratories, clubs, and common-interest groups. Nobody beats HAL-TRONIX quality and price. Just try us and see for yourself.

CLOCK CASE

Available and will fit any one of the above clocks
Regular Price . . . $6.50 **But Only $4.50 when bought with clock**

60-HZ TIME BASE

CRYSTAL TIME BASE KIT. WILL ENABLE MOST ALL DIGITAL CLOCKS TO OPERATE FROM 12 VDC. LOW PROFILE UNIT. EASY 3-WIRE HOOKUP. ACC 2PPM. ADJUSTABLE.
COST ONLY $4.95 EACH OR 2 FOR $9.00 — OR ONLY $4.50 WITH CLOCK PURCHASE.

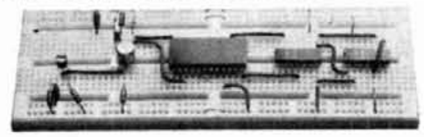
10-MHz CRYSTALS

HI-QUALITY CRYSTALS, DESIGNED FOR FREQUENCY CONTROL AND ELECTRONIC TIME PIECES; AGING FACTOR 5PPM. MEETS OR EXCEEDS MIL-C-3098 SPECS. MADE ESPECIALLY FOR HAL-TRONIX BY SENTRY.
PRICE $5.95 OR 2 FOR $10.00

WATCH FOR FUTURE ANNOUNCEMENTS OF NEW HAL-TRONIX KITS — ON THE WAY:

- POCKET-SIZE FREQUENCY COUNTERS
- CAPACITANCE METERS
- FUNCTION GENERATOR

DISTRIBUTOR FOR **A P PRODUCTS, INCORPORATED**



SUPER STRIP
SS-2 #923252
PRICE $17.00

BY POPULAR DEMAND — we are continuing to offer with any purchase of $99 or more from ad or flyer, a Fairchild **clock module** FCS-8100A (suggested retail price $20).

Look at these Summer Specials

COMPLETE KITS: CONSISTING OF EVERY ESSENTIAL PART NEEDED TO MAKE YOUR COUNTER COMPLETE. **HAL-600A** 7-DIGIT COUNTER WITH FREQUENCY RANGE OF ZERO TO 600 MHz. FEATURES TWO INPUTS: ONE FOR LOW FREQUENCY AND ONE FOR HIGH FREQUENCY; AUTOMATIC ZERO SUPPRESSION. TIME BASE IS 1.0 SEC OR .1 SEC GATE WITH OPTIONAL 10 SEC GATE AVAILABLE. ACCURACY ± .001%, UTILIZES 10-MHz CRYSTAL 5 PPM.

COMPLETE KIT . . . ~~$149~~ . . . $129

HAL-300A 7-DIGIT COUNTER WITH FREQUENCY RANGE OF ZERO TO 300 MHz. FEATURES TWO INPUTS: ONE FOR LOW FREQUENCY AND ONE FOR HIGH FREQUENCY; AUTOMATIC ZERO SUPPRESSION. TIME BASE IS 1.0 SEC OR .1 SEC GATE WITH OPTIONAL 10 SEC GATE AVAILABLE. ACCURACY ± .001%, UTILIZES 10-MHz CRYSTAL 5 PPM.

COMPLETE KIT . . . ~~$124~~ . . . $109

HAL-50A 8-DIGIT COUNTER WITH FREQUENCY RANGE OF ZERO TO 50 MHz OR BETTER. AUTOMATIC DECIMAL POINT, ZERO SUPPRESSION UPON DEMAND. FEATURES TWO INPUTS: ONE FOR LOW FREQUENCY INPUT, AND ONE ON PANEL FOR USE WITH ANY INTERNALLY MOUNTED HALTRONIX PRE-SCALER FOR WHICH PROVISIONS HAVE ALREADY BEEN MADE. 1.0 SEC AND .1 SEC TIME GATES. ACCURACY ± .001%. UTILIZES 10-MHz CRYSTAL 5 PPM.

COMPLETE KIT . . . ~~$124~~ . . . $109

ATTENTION RADIO CLUBS

For club or group projects, request FREE information about our DISCOUNTS on any of the HAL-TRONIX kits. Discounts range from 10-25%, depending upon the quantity needed.
We are experienced in supplying kits in volume quantities to schools, laboratories, clubs, and common-interest groups. Nobody beats HAL-TRONIX quality and price. Just try us and see for yourself.

NEW FROM HAL-TRONIX

DELUXE 12-BUTTON TOUCH-TONE ENCODER KIT utilizing the new ICM 7206 chip. Provides both VISUAL AND AUDIO indications! Comes with its own two-tone anodized aluminum cabinet. Measures only 2¾ x 3¾". Complete with Touch-Tone pad, board, crystal, chip and all necessary components to finish the kit.
PRICED AT $29.95
For those who wish to mount the encoder in a hand-held unit, the PC board measures ony 9/16" x 1¾". This partial kit with PC board, crystal, chip and components.
PRICED AT $14.95

PRE-SCALER KITS

HAL 300 PRE $19.95
(Pre-drilled G10 board and all components)
HAL 300 A/PRE $24.95
(Same as above but with preamp)
HAL 600 PRE $34.95
(Pre-drilled G10 board and all components)
HAL 600 A/PRE $39.95
(Same as above but with preamp)

SPECIAL — due to OVERSTOCK (while they last!)

FAIRCHILD FND-70
common cathode readouts (can replace FND-359 . . . same pin-out)

Qty.	Price each	Amount
10	40c	$4.00
100	35c	35.00
500	30c	150.00
1000	25c	250.00

"HAL"
HAROLD C. NOWLAND
W8ZXH

HAL-TRONIX
PO BOX 1101, SOUTHGATE, MI 48195
PHONE (313) 285-1782

SHIPPING INFORMATION:
ORDERS OVER $15.00 WILL BE SHIPPED POSTPAID EXCEPT ON ITEMS WHERE ADDITIONAL CHARGES ARE REQUESTED. ON ORDERS LESS THAN $15.00 PLEASE INCLUDE ADDITIONAL $1.00 FOR HANDLING AND MAILING CHARGES. SEND **SASE** FOR FREE FLYER.

More Details? CHECK—OFF Page 142

More Details? CHECK—OFF Page 142

7400 TTL

SN7400N	.16	SN7470N	.29	SN74160N	.89
SN7401N	.18	SN7472N	.29	SN74161N	.89
SN7402N	.18	SN7473N	.35	SN74162N	1.95
SN7403N	.18	SN7474N	.35	SN74163N	.89
SN7404N	.18	SN7475N	.49	SN74164N	.89
SN7405N	.20	SN7476N	.35	SN74165N	.89
SN7406N	.29	SN7479N	5.00	SN74166N	1.25
SN7407N	.29	SN7480N	.50	SN74167N	1.95
SN7408N	.20	SN7482N	.99	SN74170N	1.59
SN7409N	.20	SN7483N	.59	SN74172N	6.00
SN7410N	.18	SN7485N	.79	SN74173N	1.25
SN7411N	.25	SN7486N	.35	SN74174N	.89
SN7412N	.25	SN7489N	1.75	SN74175N	.79
SN7413N	.40	SN7490N	.45	SN74176N	.79
SN7414N	.70	SN7491N	.59	SN74177N	.79
SN7416N	.25	SN7492N	.43	SN74179N	1.95
SN7417N	.25	SN7493N	.43	SN74180N	.79
SN7420N	.20	SN7494N	.65	SN74181N	1.95
SN7421N	.29	SN7495N	.65	SN74182N	.79
SN7422N	.39	SN7496N	.65	SN74184N	1.95
SN7423N	.25	SN7497N	3.00	SN74185N	1.95
SN7425N	.29	SN74100N	.89	SN74186N	9.95
SN7426N	.29	SN74107N	.35	SN74188N	3.95
SN7427N	.25	SN74109N	.59	SN74190N	1.25
SN7429N	.39	SN74116N	1.95	SN74191N	1.25
SN7430N	.20	SN74121N	.35	SN74192N	.79
SN7432N	.25	SN74122N	.39	SN74193N	.79
SN7437N	.25	SN74123N	.49	SN74194N	.89
SN7438N	.25	SN74125N	.49	SN74195N	.69
SN7439N	.25	SN74126N	.49	SN74196N	.89
SN7440N	.20	SN74132N	.75	SN74197N	.89
SN7441N	.89	SN74136N	.75	SN74198N	1.49
SN7442N	.49	SN74141N	.79	SN74199N	1.49
SN7443N	.75	SN74142N	2.95	SN74200N	5.59
SN7444N	.75	SN74143N	2.95	SN74251N	1.79
SN7445N	.75	SN74144N	2.95	SN74279N	.79
SN7446N	.69	SN74145N	.79	SN74283N	2.25
SN7447N	.59	SN74147N	1.95	SN74284N	3.95
SN7448N	.79	SN74148N	1.29	SN74285N	3.95
SN7450N	.20	SN74150N	.89	SN74365N	.69
SN7451N	.20	SN74151N	.59	SN74366N	.69
SN7453N	.20	SN74152N	.59	SN74367N	.69
SN7454N	.20	SN74153N	.59	SN74368N	.69
SN7459A	.25	SN74154N	.99	SN74390N	1.95
SN7460N	.20	SN74155N	.79	SN74393N	1.95
		SN74156N	.79		
		SN74157N	.65		

20% Discount 100 pcs combined order 25% -1000 pcs combined order.

C/MOS

CD4000	.23	CD4028	.89	CD4070	.55
CD4001	.23	CD4029	1.19	CD4071	.23
CD4002	.23	CD4030	.49	CD4072	.49
CD4006	1.19	CD4035	.99	CD4076	1.39
CD4007	.25	CD4040	1.19	CD4081	.23
CD4009	.49	CD4041	1.25	CD4082	.23
CD4010	.49	CD4042	.99	CD4093	.99
CD4011	.23	CD4043	.89	CD4098	2.49
CD4012	.25	CD4044	.89	MC14409	14.95
CD4013	.39	CD4046	1.79	MC14410	14.95
CD4014	1.39	CD4047	2.50	MC14411	14.95
CD4015	1.19	CD4048	1.35	MC14419	4.95
CD4016	.49	CD4049	.49	MC14433	19.95
CD4017	1.19	CD4050	.49	MC14506	.75
CD4018	.99	CD4051	1.19	MC14507	.99
CD4019	.49	CD4053	1.19	MC14562	14.50
CD4020	1.19	CD4056	2.95	MC14583	3.50
CD4021	1.39	CD4059	9.95	CD4508	3.95
CD4022	1.19	CD4060	1.49	CD4510	1.39
CD4023	.23	CD4066	.79	CD4511	1.29
CD4024	.79	CD4068	.39	CD4515	2.95
CD4025	.23	CD4069	.45	CD4518	1.29
CD4026	2.25			CD4520	1.29
CD4027	.69			CD4566	2.25

74C00

74C00	.39	74C89	6.49	74C163	3.00
74C02	.55	74C90	3.00	74C164	3.25
74C04	.75	74C93	2.00	74C173	2.60
74C08	.75	74C95	2.00	74C192	3.49
74C10	.65	74C107	1.25	74C193	2.75
74C14	3.00	74C151	2.90	74C195	2.75
74C20	.65	74C154	3.00	74C922	9.95
74C30	.65	74C157	2.15	74C923	8.95
74C42	2.15	74C160	3.25	74C925	14.95
74C48	4.75	74C161	3.25	74C926	11.95
74C73	1.50			80C95	1.50
74C74	1.15			80C97	1.50

LINEAR

78MG	1.75	LM340T-8	1.25	LM733N	1.00
LM300H	.80	LM340T-12	1.25	LM739N	1.19
LM301H	.35	LM340T-15	1.25	LM741CH	.35
LM301CN	.35	LM340T-18	1.25	LM741CN	.35
LM302H	.75	LM340T-24	1.25	LM741-14N	.39
LM304H	1.00	LM350N	1.00	LM747H	.79
LM305H	.60	LM351CN	.65	LM747N	.79
LM307CN/H	.35	LM370N	1.15	LM748H	.39
LM308H	1.00	LM373N	3.25	LM748N	.39
LM308CN	1.00	LM377N	4.00	LM1303N	.90
LM309H	1.10	LM380N	1.25	LM1304N	1.19
LM309K	1.25	LM380CN	.99	LM1305N	1.40
LM310CN	1.15	LM381N	1.79	LM1307N	.85
LM311H	.90	LM382N	1.79	LM1310N	2.95
LM311N	.90	NE501N	8.00	LM1351N	1.65
LM317K	6.50	NE510A	6.00	LM1414N	1.75
LM318CN	1.50	NE529A	4.95	LM1458CN/H	.59
LM319N	1.30	NE531H	3.00	LM1488N	1.95
LM320K-5	1.35	NE536T	6.00	LM1489N	1.95
LM320K-5.2	1.35	NE540L	6.00	LM1496N	.95
LM320K-12	1.35	NE550N	1.30	LM1556V	1.75
LM320K-15	1.35	NE555V	.39	MC1741SCP	3.00
LM320K-18	1.35	NE556N	1.00	LM2901N	2.95
LM320K-24	1.35	NE560B	5.00	LM3053N	1.50
LM320T-5	1.25	NE561B	5.00	LM3065N	.69
LM320T-5.2	1.25	NE562B	5.00	LM3900N(3401)	.49
LM320T-8	1.25	NE565H	1.75	LM3905N	.89
LM320T-12	1.25	NE565N	1.25	LM3909N	1.25
LM320T-15	1.25	NE566CN	1.75	MC5558V	1.00
LM320T-18	1.25	NE567H	1.25	LM7525N	.90
LM320T-24	1.25	NE567V	.99	LM7534N	.75
LM323K-5	5.95	NE570N	10.50	8038B	4.95
LM324N	1.80	LM703CN/H	.45	LM75450N	.50
LM339N	.99	LM709H	.29	75451CN	.39
LM340K-5	1.35	LM709N	.29	75452CN	.39
LM340K-6	1.35	LM710N	.79	75453CN	.39
LM340K-8	1.35	LM711N	.39	75454CN	.39
LM340K-12	1.35	LM723H	.55	75491CN	.79
LM340K-15	1.35	LM723N	.55	75492CN	.89
LM340K-18	1.35			75494CN	.89
LM340K-24	1.35			RC4151	5.95
LM340T-5	1.25			RC4194	5.95
LM340T-6	1.25			RC4195	4.49

74LS00 TTL

74LS00	.23	74LS51	.23	74LS139	.69
74LS01	.23	74LS54	.23	74LS151	.69
74LS02	.23	74LS55	.23	74LS155	.69
74LS03	.23	74LS73	.35	74LS157	.69
74LS04	.29	74LS74	.35	74LS160	.89
74LS05	.29	74LS75	.49	74LS161	.89
74LS08	.29	74LS76	.35	74LS162	.89
74LS09	.29	74LS83	.75	74LS163	.89
74LS10	.23	74LS85	.99	74LS164	.99
74LS13	.49	74LS86	.35	74LS175	.79
74LS14	.99	74LS90	.49	74LS181	2.49
74LS15	.29	74LS92	.59	74LS190	.89
74LS20	.23	74LS93	.59	74LS191	.89
74LS21	.29	74LS95	.79	74LS192	.89
74LS22	.29	74LS96	.89	74LS193	.89
74LS26	.29	74LS107	.35	74LS194	.89
74LS27	.29	74LS109	.35	74LS195	.89
74LS28	.29	74LS112	.35	74LS253	.79
74LS30	.23	74LS123	.99	74LS257	.69
74LS32	.29	74LS132	.79	74LS260	.55
74LS37	.35	74LS136	.39	74LS279	.59
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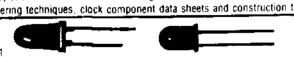

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MAN 72	Common Anode-red	.300	.99	MAN 6780	Common Cathode-red	.560	.99
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22 pin LP	$.37	.36	.35
24 pin LP	.38	.37	.36
28 pin LP	.45	.44	.43
36 pin LP	.60	.59	.58
40 pin LP	.63	.62	.61

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18 pin ST	.35	.32	.30
24 pin ST	.49	.45	.42
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36 pin ST	1.39	1.26	1.15
40 pin ST	1.59	1.45	1.30

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16 pin SG	.38	.35	.32
18 pin SG	.52	.47	.43
24 pin SG	$.70	.63	.57
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40 pin SG	1.75	1.59	1.45

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14 pin WW	.39	.38	.37
16 pin WW	.43	.42	.41
18 pin WW	.75	.68	.62
22 pin WW	$.95	.85	.75
24 pin WW	1.05	.95	.85
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36 pin WW	1.59	1.45	1.30
40 pin WW	1.75	1.55	1.40

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ASST. 2	5 ea.	68 OHM, 82 OHM, 100 OHM, 120 OHM, 150 OHM, 180 OHM, 220 OHM, 270 OHM, 330 OHM, 390 OHM					1/4 WATT 5%	50 PCS.
ASST. 3	5 ea.	470 OHM, 560 OHM, 680 OHM, 820 OHM, 1K, 1.2K, 1.5K, 1.8K, 2.2K, 2.7K					1/4 WATT 5%	50 PCS.
ASST. 4	5 ea.	3.3K, 3.9K, 4.7K, 5.6K, 6.8K, 8.2K, 10K, 12K, 15K, 18K					1/4 WATT 5%	50 PCS.
ASST. 5	5 ea.	22K, 27K, 33K, 39K, 47K, 56K, 68K, 82K, 100K, 120K					1/4 WATT 5%	50 PCS.
ASST. 6	5 ea.	150K, 180K, 220K, 270K, 330K, 390K, 470K, 560K, 680K, 820K					1/4 WATT 5%	50 PCS.
ASST. 7	5 ea.	1M, 1.2M, 1.5M, 1.8M, 2.2M, 2.7M, 3.3M, 3.9M, 4.7M, 5.6M					1/4 WATT 5%	50 PCS.

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EXAR

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XR215	4.40	XR1800	3.20	XR3403	1.25
XR320	1.55	XR2206	4.40	XR4136	1.25
XR555	.39	XR2207	3.85	XR4151	2.85
XR556	.99	XR2208	5.20	XR4194	1.45
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XR567CT	1.25	XR2211	5.25	XR4212	2.05
XR1310P	1.30	XR2212	4.35	XR4558	.75
XR1468CN	3.85	XR2240	3.45	XR4739	1.15
XR1488	1.39	XR2264	4.25	XR4741	1.47

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TYPE	VOLTS	W	PRICE	TYPE	VOLTS	W	PRICE
1N746	3.3	400m	4/1.00	1N4005	600 PIV	1 AMP	10/1.00
1N751A	5.1	400m	4/1.00	1N4006	800 PIV	1 AMP	10/1.00
1N752	5.6	400m	4/1.00	1N4007	1000 PIV	1 AMP	10/1.00
1N753	6.2	400m	4/1.00	1N3600	50	200m	6/1.00
1N754	6.8	400m	4/1.00	1N4148	75	10m	15/1.00
1N959	8.2	400m	4/1.00	1N4154	35	10m	12/1.00
1N965B	15	400m	4/1.00	1N4305	75	25m	20/1.00
1N5232	5.6	500m	.28	1N4734	5.6	1w	.28
1N5234	6.2	500m	.28	1N4735	6.2	1w	.28
1N5235	6.8	500m	.28	1N4736	6.8	1w	.28
1N5236	7.5	500m	.28	1N4738	8.2	1w	.28
1N456	25	40m	6/1.00	1N4742	12	1w	.28
1N458	150	7m	6/1.00	1N4744	15	1w	.28
1N485A	180	10m	5/1.00	1N1183	50 PIV	35 AMP	1.60
1N4001	50 PIV	1 AMP	12/1.00	1N1184	100 PIV	35 AMP	1.70
1N4002	100 PIV	1 AMP	12/1.00	1N1185	150 PIV	35 AMP	1.70
1N4003	200 PIV	1 AMP	12/1.00	1N1186	200 PIV	35 AMP	1.80
1N4004	400 PIV	1 AMP	12/1.00	1N1188	400 PIV	35 AMP	3.00

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C38M	35A @ 600V	SCR	1.95
2N2328	1.6A @ 300V	SCR	.50
MDA 980-1	12A @ 50V	FW BRIDGE REC	1.95
MDA 980-3	12A @ 200V	FW BRIDGE REC	1.95

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MPSA06	5/1.00	MJE3055	1.00	2N3906	4/1.00
TIS97	6/1.00	2N3392	5/1.00	2N4013	3/1.00
TIS98	6/1.00	2N3398	5/1.00	2N4123	6/1.00
TIS133	5/1.00	PN3567	3/1.00	PN4249	4/1.00
TIS135	5/1.00	PN3568	4/1.00	PN4250	4/1.00
40409	1.75	PN3569	4/1.00	2N4400	4/1.00
40410	1.75	MPS3638A	5/1.00	2N4401	4/1.00
40673	1.75	MPS3702	5/1.00	2N4402	4/1.00
2N918	4/1.00	2N3704	5/1.00	2N4403	4/1.00
2N2219A	3/1.00	MPS3704	5/1.00	2N4409	5/1.00
2N2221A	4/1.00	2N3705	5/1.00	2N5086	4/1.00
2N2222A	5/1.00	MPS3705	5/1.00	2N5087	4/1.00
2N2369	5/1.00	2N3706	5/1.00	2N5088	4/1.00
2N2369A	4/1.00	MPS3706	5/1.00	2N5089	4/1.00
MPS2369	5/1.00	2N3707	5/1.00	2N5129	5/1.00
2N2484	4/1.00	2N3711	5/1.00	PN5134	5/1.00
2N2906	4/1.00	2N3724A	.65	PN5138	5/1.00
2N2907	5/1.00	2N3725A	1.00	2N5139	5/1.00
2N2925	5/1.00	2N3772	2.25	2N5210	5/1.00
MJE2955	1.25	2N3823	1.00	2N5449	3/1.00
2N3053	2/1.00	2N3903	4/1.00	2N5951	3/1.00

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22 pf	.05	.04	.03	.0047µF	.05	.04	.035
47 pf	.05	.04	.03	.01µF	.05	.04	.035
100 pf	.05	.04	.03	.022µF	.06	.05	.04
220 pf	.05	.04	.03	.047µF	.06	.05	.04
470 pf	.05	.04	.035	.1µF	.12	.09	.075

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.0022	.12	.10	.07	.047mf	.21	.17	.13
.0047mf	.12	.10	.07	.1mf	.27	.23	.17
.01mf	.12	.10	.07	.22mf	.33	.27	.22

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.22/35V	.28	.23	.17	3.3/25V	.31	.27	.22
.33/35V	.28	.23	.17	4.7/25V	.32	.28	.23
.47/35V	.28	.23	.17	6.8/25V	.36	.31	.25
.68/35V	.28	.23	.17	10/25V	.40	.35	.29
1.0/35V	.28	.23	.17	15/25V	.63	.50	.40

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1.0/50V	.16	.14	.11	.47/50V	.16	.14	.11
3.3/50V	.14	.12	.09	1.0/16V	.15	.13	.10
4.7/25V	.15	.13	.10	1.0/25V	.16	.14	.11
10/25V	.15	.13	.10	1.0/50V	.16	.14	.11
10/50V	.16	.14	.12	4.7/16V	.15	.13	.10
22/25V	.17	.15	.12	4.7/25V	.15	.13	.10
22/50V	.24	.20	.18	4.7/50V	.16	.14	.12
47/25V	.19	.17	.15	10/16V	.14	.12	.09
47/50V	.25	.21	.19	10/25V	.15	.13	.10
100/25V	.24	.20	.18	10/50V	.16	.14	.12
100/50V	.35	.30	.28	47/50V	.24	.21	.19
220/25V	.32	.28	.25	100/16V	.19	.15	.14
220/50V	.45	.41	.38	100/25V	.24	.20	.18
470/25V	.33	.29	.27	100/50V	.35	.30	.28
1000/16V	.55	.50	.45	220/16V	.23	.17	.16
2200/16V	.70	.62	.55	470/25V	.31	.28	.26

More Details? CHECK—OFF Page 142

PRODUCTS OF SOUND RESEARCH

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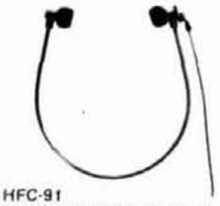
COMMUNICATIONS, INC.

Lightweight Headphones



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$19.95

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Flight Line audio quality and performance for that FM transceiver. Rugged selectable noise cancelling, 43 dB sensitivity, variable gain shaping amplifier. Use on high or low Z applications. Available in grey Cycolac or Clear Tenite.
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Lightweight Headset
VHF operation will never be the same. Noise cancelling magnetic microphone, 8 ohm earphone, in line PTT and variable gain shaping amplifier deliver performance Hams have been waiting for – 43 dB maximum microphone sensitivity at 600 ohms – complete with eyeglass adapter.
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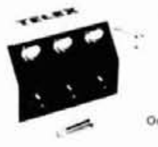

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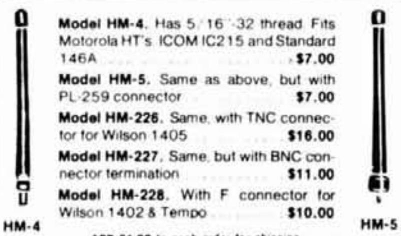

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Legal ammunition available

The tragedy is that suits are being lost that could have been won. But TVI/RFI and tower cases fall into a little-known area of the law. Unless your lawyer is a specialist, he could spend hundreds of hours researching court decisions. And still not be sure he's put together the strongest defense possible. It's expensive (expect to spend an average $4,000 to $8,000 if you're sued). And risky. Which is why we formed the non-profit Personal Communications Foundation.* To provide your lawyer with legal ammunition.

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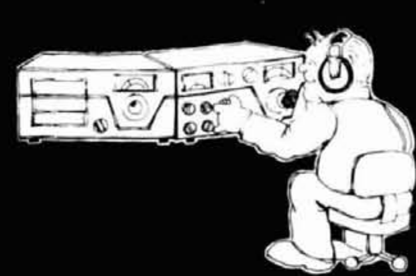

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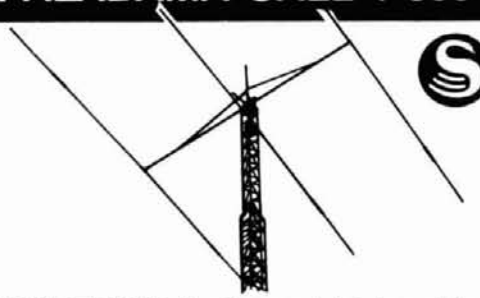

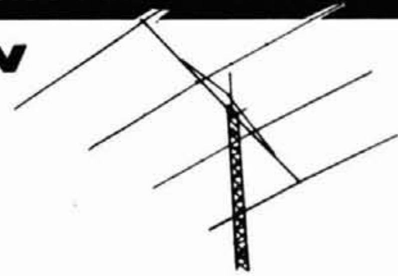

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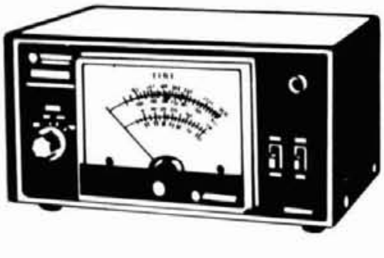

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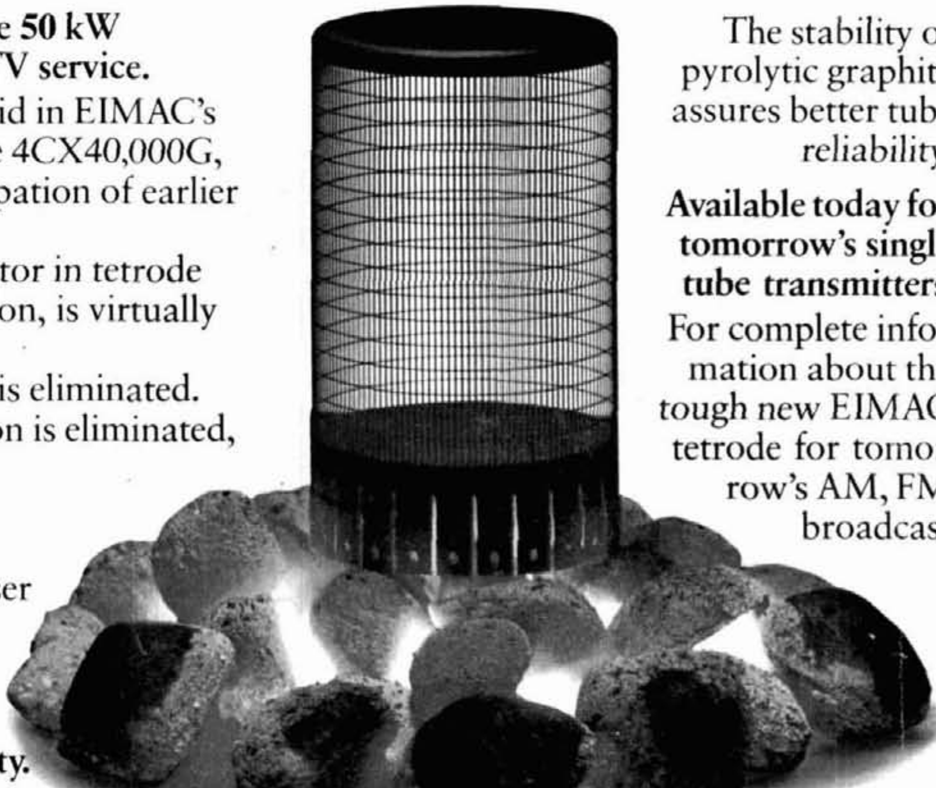

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