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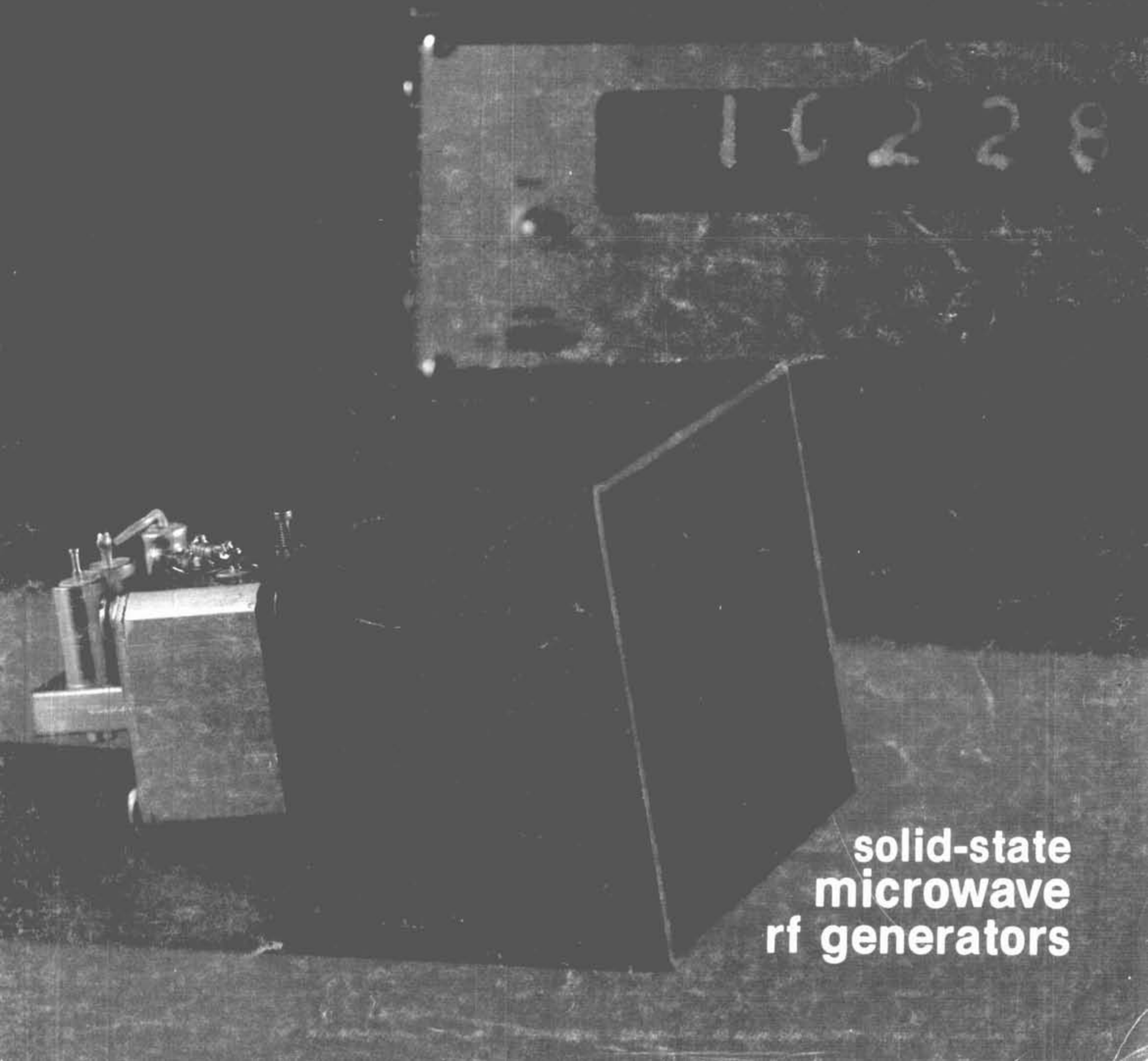
ham radio magazine



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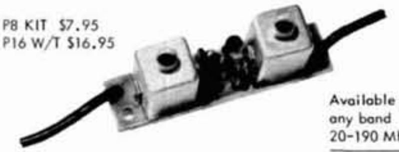

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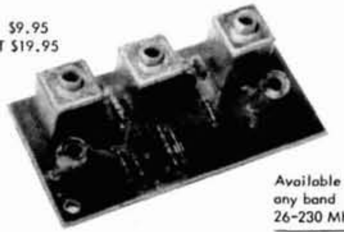

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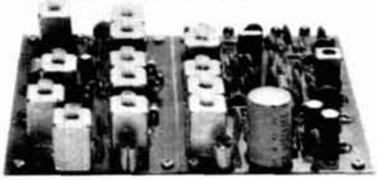

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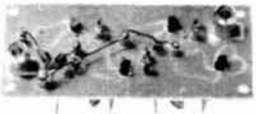

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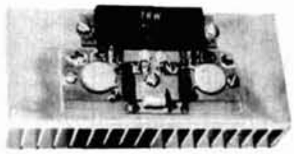

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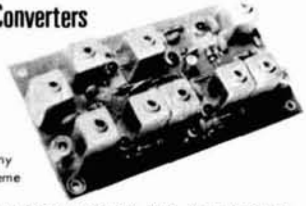

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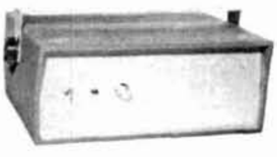

- Uses T80 RF Power Module and P8 Preamp
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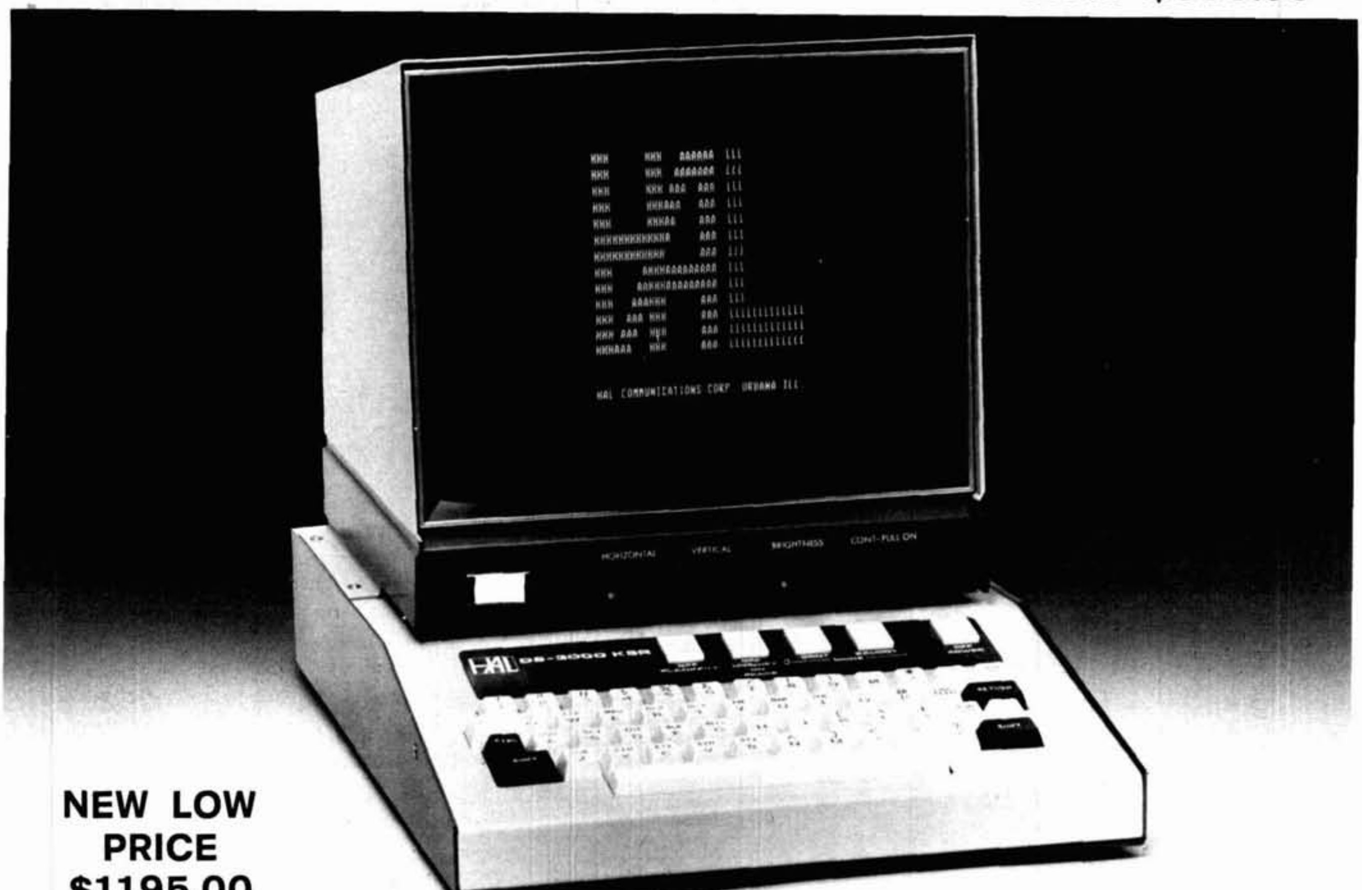

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The HAL ST-6000 operates at standard shifts of 850, 425, and 170 Hz. The tone keyer is crystal-controlled. Loop supply is internal. Active filters allow flexibility in establishing different tone pairs. You can select AM or hard-limiting FM modes of operation to accommodate different operating conditions. An internal monitor scope (shown on model above) allows fast, accurate tuning. The ST-6000 has an outstandingly high dynamic range of operation. Data I/O can be RS-232C, MIL-188C or current loop.

The DS-3000 and DS-4000 series of KSR and RO terminals provide silent, reliable, all-electronic TTY transmission and reception, or read-only (RO) operation of different combinations of codes, including Baudot, ASCII and Morse. The powerful, programmable 8080A microprocessor is included in the circuitry to assure maximum flexibility for your present needs —and for the future. The KSR models offer you full editing capability. The video display is a convenient 16-line format, of 72 characters per line.

These are some of the highlights. The full range of features and specifications for the ST-6000 and the DS series of KSR and RO terminals is covered in comprehensive data sheets available on request. Write for them now—and tune in to the most sophisticated TTY operation you can have today...or in the future.

HAL

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ham radio magazine

APRIL 1977

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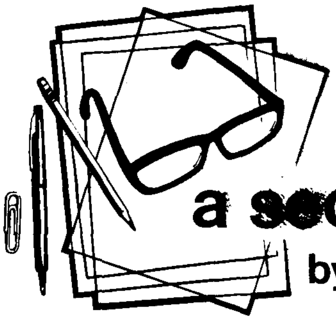

a second look

by Jim Fisk

A new era in community education may well be the outcome if a petition filed recently with the FCC receives favorable action. Based largely on experience gained with vhf repeaters, and more specifically, amateur television repeaters, The Center for Advanced Study in Education of the City University of New York has filed a petition for the establishment of a new community educational radio service to be known as *Communicasting*. Called communicasting because it embodies elements of both communications and broadcasting, the new service would use television channels 70 through 83 for co-channel, multilateral video and audio communications. Using inexpensive terminals in homes, schools, community centers, libraries and hospitals, the system would tie the community together in an interactive educational network. The petition is co-sponsored by the Communicasting Association of America, a non-profit organization headed by W2KPQ which is dedicated to using the radio spectrum for multilateral educational and scientific communications.

If Communicasting is approved, any individual at home would be able to receive the transmissions on one of the unused vhf channels of his television set. The low-power signals from the remote terminals would be transmitted to a translator where they would be re-transmitted on one of the uhf TV channels. The antennas would be high enough to cover the entire community. It is an idea that can effectively and inexpensively implement the concept of "Communicate instead of Commute" by providing electronic classrooms, forums, and lecture halls.

One of the first examples of Communicasting was established on the amateur two-meter band in 1955 when the Albany Medical College started a novel form of post-graduate education: Two or three members of the faculty discussed a medical topic while in direct radio communication with doctors located in outlying hospitals. The conference network consisted of a high-power transmitter at Albany Medical College and lower-powered units at twenty-one hospitals throughout eastern New York and western Massachusetts. The system is functioning to this day.

The principal of Communicasting was further demonstrated on the MARS frequencies in 1958-1960 where it was used for on-the-air scientific and educational forums. Today it is being used for weekly technical nets which are transmitted through a vhf repeater in the New York metropolitan area. Further experiments will be conducted in future months on the uhf television repeater recently unveiled by the Long Island Mobile Amateur Radio Club.

Many educators have recognized the potential of interactive radio and television in traditional classroom activities as well as in continuing and extension programs, and homebound education. At the City University of New York, the Center for Advanced Study in Education and the Institute for Research and Development in Occupational Education have been actively developing courses of study for electronic classrooms, studying the most effective way of delivering the curriculum, and assessing the coverage available with direct transmission and with repeaters. Their research will have a direct bearing on initial efforts to demonstrate a working system in New York State during the 1977-1978 academic year.

The FCC recognized the need for an "Educational Radio Service" in 1963 when they established the Instructional Television Fixed Service (ITFS) in the 2.5-2.69 GHz band. However, since equipment for these frequencies is up to a hundred times more expensive than equipment for the uhf bands, the use of ITFS is effectively limited to large, wealthy institutions who can afford the equipment. The proposed Communicasting network would put it within the reach of everyone.

As Dr. Lee Cohen of the City University of New York said recently, "At present we are in the stone age of multilateral education and scientific communication by radio. We look forward to the day when the FCC will allocate a band of frequencies wherein professional and educational groups could organize radio forums . . . This could eliminate the problems of time and distance in getting some of our foremost minds to communicate by radio, thereby educating a listening/viewing audience."

If you are interested in supporting this worthwhile proposal, or would like to know more about it, write to Ed Piller, W2KPQ, Communicasting Association of America, Inc., 80 Birchwood Park Drive, Syosset, New York 11791.

Jim Fisk, W1HR
editor-in-chief

ICOM INTRODUCES THE REVOLUTION IN VFO TECHNOLOGY

Introducing the IC-245, 144-148 MHz FM Transceiver

The VFO Revolution goes mobile with the unique, ICOM developed LSI synthesizer with 4 digit LED readout. The **IC-245** offers the most for mobile on the market. The easy to use tuning knob moves accurately over 50 detent steps and assures excellent control as easily as steering the vehicle. With its optional adapter, the **IC-245** puts you into all mode operation on 12V DC power with a compact dash-mounted transceiver. In FM, the synthesizer command frequency is displayed in 5 KHz steps from 146 to 148 MHz, and with the side band adapter the step rate drops to 100Hz from 144 to 146 MHz. For maximum repeater flexibility, the transmit and receive frequencies are independently programable on any separation. The **IC-245** even comes equipped with a multiple pin Molex connector for remote control.

The **IC-245** is a product of the revolution in VFO design, from its new style front panel, to its excellent mechanical rigidity and Large Scale Integrated Circuitry. Your **IC-245** will give you the most for mobile.

SPECIFICATIONS

GENERAL	
Frequency Coverage	**144.00 to 148.00 MHz*
Modes	*FM (F3)* **SSB (A3J), CW (A1)*
Supply Voltage	*DC 13.8V ±15%*
Size (mm)	*90H x 155W x 235D*
Weight (kg)	*2.7*
TRANSMITTER	
TX Output	*F3 10W* **A3J 10W (PEP), A1 10W*
Carrier Suppression	*40 dB or better*
Spurious Radiation	*−60 dB or less below carrier*
Maximum Frequency Deviation	*±5 KHz*
Microphone Impedance	*600 ohms*
RECEIVER:	
Sensitivity	**A3J, A1 0.5 microvolt input gives 10 dB S+N/N or better F3 0.6 microvolt or less for 20 dB quieting S+N+D/N at 1 microvolt input, 30 dB*
Squelch Threshold	*−8 dB or less (F3)*
Spurious Response	*−60 dB or better*
SYNTHESIZER:	
Frequency Range	*144 MHz to 148 MHz*
Step Size	*5 KHz for FM* **100 Hz or 5 KHz for SSB*
Stability	*per C in the range of −10 to +60 C, ±0.0000145% per C*

* Valid with SSB Adapter only

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presstop

FCC'S PROPOSED AMPLIFIER BAN would prohibit the marketing of external RF amplifiers capable of operation from 24 through 35 MHz. In its February 18 Public Notice, the Commission specified its concern with the so-called "broad-band linears," which replaced "business-band" linears in the marketplace after those had been banned two and a half years ago, in February, 1975.

In Limiting Its Proposed Ban to 24-35 MHz the FCC also made note that though the Amateur service would be affected Amateurs would still be permitted to build 10-meter amplifiers or modify commercial units to cover 10 meters for their own use, while they "would respect the intent of this regulation and not supply these devices to non-Amateurs." Such construction by individual Amateurs would be limited to a single unit of a given model.

Some Specific Areas the FCC would like Amateurs to address in their comments are:

Any Further Requirements which may be necessary to prevent the use of illegal amplification devices;

Practicality Of Such a prohibition and possible techniques which could be used to produce such an amplifier;

Problems Associated With preventing the few unscrupulous manufacturers from including such features as accessible wiring which could be cut to provide for operation on the prohibited frequencies; and

Controls Which Could provide for operation on these frequencies, or any other concepts which could be used to circumvent this prohibition.

Comments On This Docket, 21116, are due May 25; Reply Comments must be submitted by June 6.

Amateur Transmitters and amplifiers would both require type acceptance under the terms of Docket 21117. In this NPRM the FCC pointed out that most current Amateur equipment is commercial and some of it is capable of operation on CB frequencies, but type acceptance could help control that capability. Furthermore, though Amateurs bear the basic responsibility for the performance of their equipment, type acceptance would be a means by which that responsibility could be shared by the makers.

Specific Exemptions from the type acceptance requirement for equipment built or modified by Amateurs for their own use were also proposed, as were provisions for type acceptance of kit-built designs. Comments on Docket 21117 are also due May 25 with Reply Comments June 6.

In Both Of These far-reaching Notices of Proposed Rule Making the Commissioners have left the door open fot workable alternative-solutions to the problem of non-Amateur use of Amateur equipment on the CB bands. Three months should provide enough time to find some such solutions. Let's hope so!

SECONDARY AMATEUR STATION LICENSES could be abolished entirely or a moratorium imposed on processing Amateur applications other than those from new, upgrading or renewing Amateurs if the tide of multiple applications presently arriving at Gettysburg isn't stemmed. Since license fees were abolished in January an increasing number of new applicants and the newly eligible 1x2 seekers — now threaten to bury the limited portion of the Gettysburg facility devoted to processing Amateur applications.

FCC'S NEW NOVICE EXAM has a circuit diagram error in the Ohm's Law problem which makes it impossible to answer as presented. Gettysburg has been advised to give all Novice applicants credit for that question whether it is answered or not — Novice training instructors please note.

AMSAT AND ARRL HAVE SIGNED an agreement in which the League will provide major assistance to the on-going Amateur space program. The League also provided two technicians — WA1JLD and WA1JZC — to bolster the AMSAT effort on the AO-D spacecraft.

The AO-D Satellite (OSCAR 8) is now scheduled for a November 15th launch. It'll carry 145-28 and 145-435 MHz transponders; and, with its 500-mile high orbit, will be even easier to access — though for shorter periods — than the present Amateur satellites, OSCARS 6 and 7.

"AMATEUR RADIO...IN THE PUBLIC INTEREST" is a very attractive report on Amateur Radio in 1976 published by the ARRL for use in presenting the Amateur service to public officials and the media. Copies for PR use are available on request, but do specify with your request how and where they'll be used.

RF POLLUTION is being studied by the Environmental Protection Agency. The EPA says it's concerned about environmental exposure arising from the ever-increasing number of RF sources, expects its extensive monitoring survey to lead to significant data within the next 18-24 months. They've already determined that "significant portions of the population are exposed to 0.1-1 microwatt/square centimeter range" radiation from the 55 to 1000-MHz part of the spectrum.

Does Your Transmitter Love Your Antenna?

If you're fighting the constant battle of limited band width, high SWR ratios, inefficient low-pass TVI filter operation due to high SWR you're not alone.

DenTron makes the Problem Solvers.

The DenTron tuners give you maximum power transfer from your transmitter to your antenna, and isn't that where it really counts?

Our Super Tuners (A. B. & E.) are the only tuners on the market that match everything between 160 and 10 meters. Whether you have balanced line, coax cable, random or long wire the DenTron Super Tuners will match the antenna impedance to your transmitter.

NEW: The Monitor Tuner (E.) was designed because of overwhelming demand. Hams told us they wanted a 3 killowatt tuner with a built-in wattmeter, a front panel antenna selector for coax, balanced line and random wire. So we engineered the 160-10m Monitor Tuner. It's a life time investment at $299.50

The DenTron 80-10 AT (D.) is a random wire, 80-10 meter tuner which is ideal for portable operation or apartment dwellers.

Every serious ham knows he must read both forward and reverse wattage simultaneously for that perfect match. So upgrade with the DenTron W-2 Dual in line Wattmeter. (C.)

The flexibility we build into our Tuners make any previous tuner you might have owned obsolete.

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B. Super Super Tuner 3 KW PEP	$229.50
C. W-2 Wattmeter	$ 99.50
D. 80-10 AT 500 W PEP	$ 59.50
E. Monitor Tuner 3 KW PEP	$299.50

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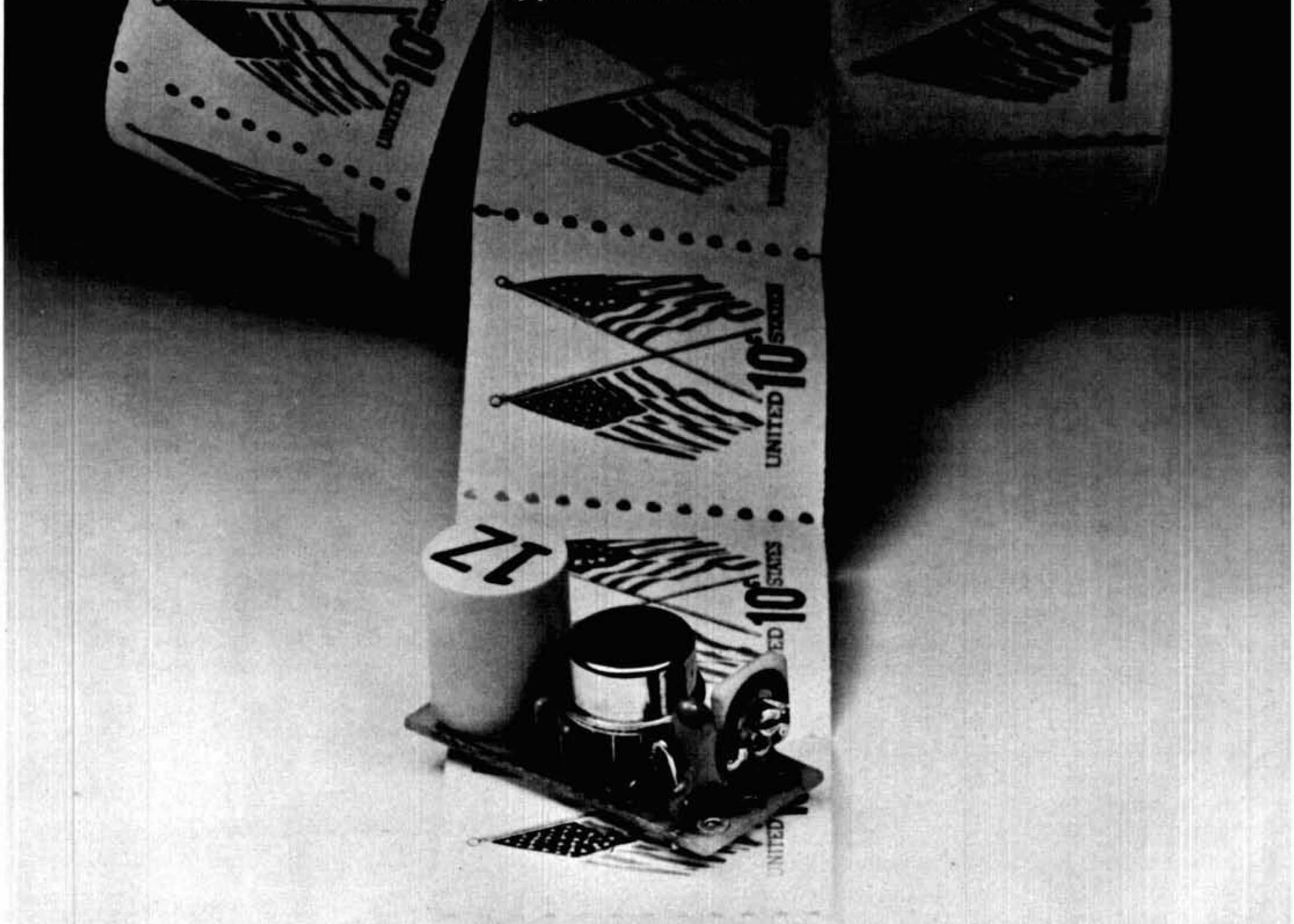

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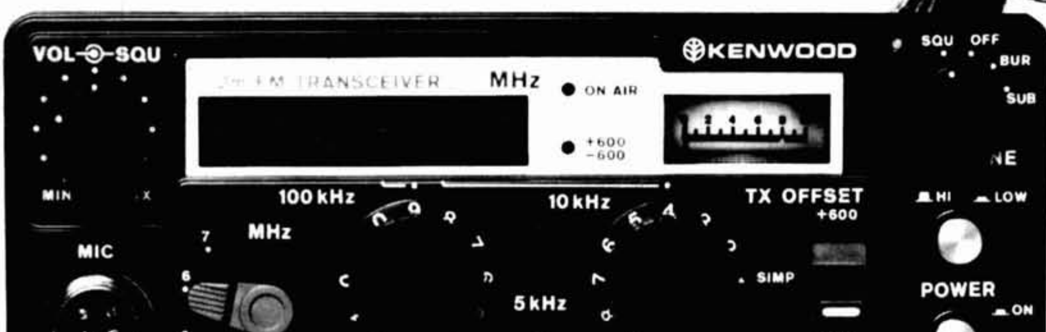

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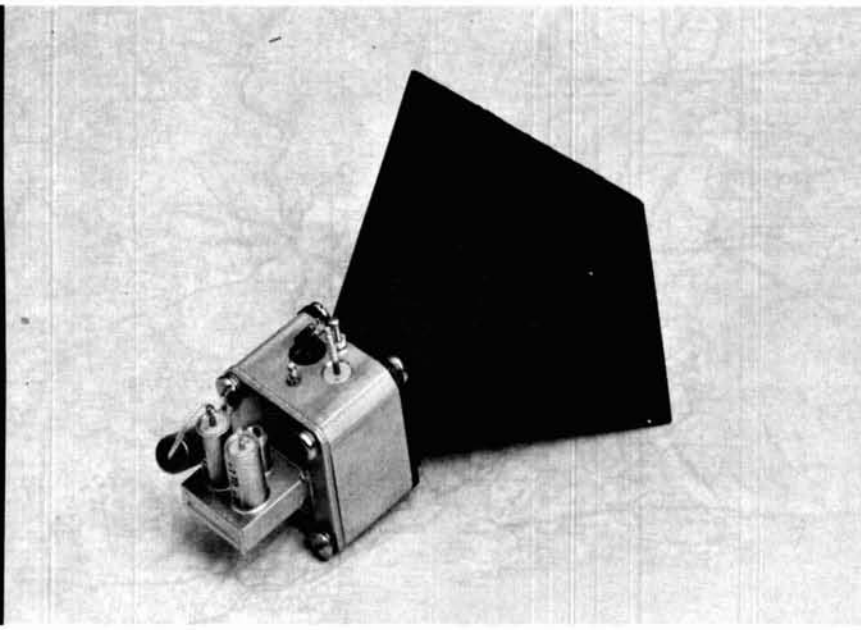


solid-state microwave rf generators

A discussion of microwave diode oscillators, including a description of the Gunnplexer — a complete solid-state transceiver for the amateur 10-GHz band

Of the many interesting trends that have developed in radio, perhaps none has been more spectacular than that toward the higher frequencies. No doubt many readers will remember when the term "high frequencies" was used to distinguish 100 kHz from audio, and when the "ultra highs" meant anything above 10 meters. Although most of the early developments in wireless around the turn of the century were accomplished on radio frequencies below 500 kHz, it's interesting to note that Heinrich Hertz's first experiments with wireless transmission in 1887 were conducted near 60 MHz; he later extended them to 500 MHz. In the first decade following Hertz's discovery the frequency frontier was quickly pushed to 75000 MHz. Marconi, in fact, used vhf in many of his early demonstrations, but quickly switched to the lower frequencies when he recognized that greater distances could be covered with the simple spark equipment then in use.

About the same time amateurs were opening up the "short waves" above 1500 kHz with their famous Trans-atlantic tests of the early 1920s, researchers E. F. Nichols and J. D. Tear had succeeded in producing radio waves as short as 0.22 mm (135 GHz). In 1923 Madame

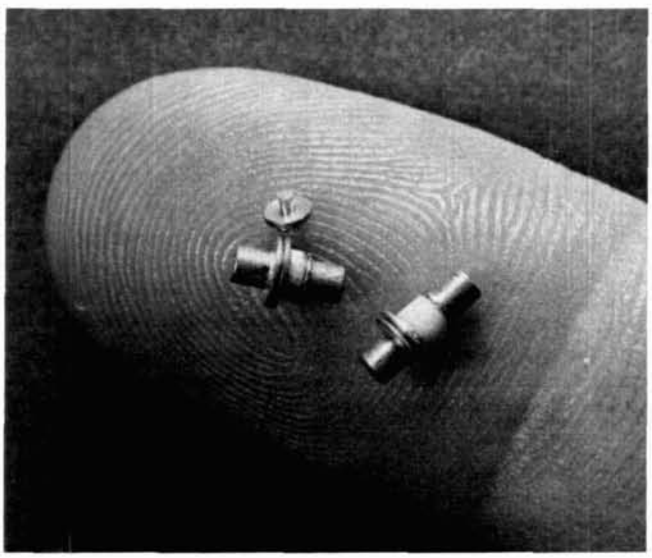
By **James R. Fisk, W1HR**, *ham radio magazine*, Greenville, New Hampshire 03048

Glagowela-Arkadiewa working in Russia extended the frequency limit to more than 35-million MHz.[1] In all of these experiments the microwave energy appeared as a harmonic component of high-energy spark discharges, and the power level was very low – perhaps microscopic would be a better description.

Soon after the invention of the three-element vacuum tube, work was started toward extending its range into the higher frequencies but there were many difficulties to be solved. Since the early tubes were built around techniques borrowed from the electric lamp industry, they were ill suited for the job at best. Extensions in frequency closely followed improvements in vacuum-tube manufacturing methods, but it wasn't too long before researchers were faced with another problem: *electron transit time – the finite time it takes an electron to cross the tube.* At the high frequencies the transit time (about a billionth of a second) is short compared to one complete rf cycle, so the electrons can follow the rf voltage fluctuations on the grid. At the very high frequencies, however, the oscillations are so rapid that the voltage on the grid may go through several complete cycles while the electron travels across the tube, and the grid voltage cannot impose its signal pattern on the electron flow. The regenerative vacuum-tube oscillator could be made to work up to 150 or 200 MHz with tuned Lecher lines, but that was about the limit, even with specially designed tubes (two amateurs, Robert Kruse and Boyd Phelps, extended this to nearly 750 MHz in 1927, but that's getting ahead of the story).

In 1920 two German engineers, H. Barkhausen and K. Kurz, found that if the grid of a vacuum triode was biased positively with respect to the plate, they could produce rf output at a wavelength of 43 cm (697 MHz).[2] With this arrangement the highly positive grid accelerates the electrons from the cathode at high speed

Small size of microwave power diodes is misleading – diodes the same size as those shown here are capable of providing CW power outputs of 500 mW or more at 10 GHz.



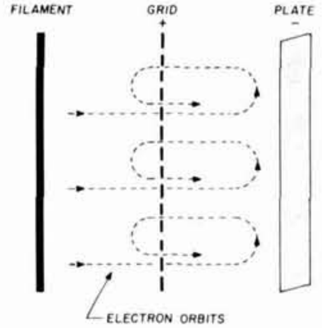


fig. 1. Mechanism of Barkhausen-Kurz oscillators, early method of obtaining uhf rf signals. The positive grid accelerates the electrons from the filament – most strike the grid and give up their energy in heat, but others pass through the openings in the grid, and are repelled by the negatively-charged plate back toward the grid. The electrons continue to orbit between the grid and plate, as shown here. The feeble output is coupled out through a tuned plate line.

(**fig. 1**) – some hit the grid and give up their energy as heat, but others pass through the openings in the grid only to be repelled back toward the grid by the negatively-charged plate. When the electrode voltages are *properly adjusted, the electrons continue to gyrate* between the grid and plate at a very high frequency. Barkhausen and Kurz further reported that the frequency of oscillations was dependent upon the applied voltages, with little regard for the external tuned circuit. This phenomenon created quite a stir among researchers, but American engineers were hard pressed to duplicate their results – because of patent fights there was an embargo on foreign vacuum tubes, and the internal construction of American tubes didn't support this mode of operation (a cylindrical, coaxial plate and grid structure was required and American tubes used a flat, sandwich type construction).

Eventually the Americans were able to "acquire" some suitable tubes from military sources who apparently weren't affected by domestic trade embargos, and the same oscillations were observed. In 1922 E. Gill and J. Morrell found that if the element voltages were very carefully adjusted, the frequency of oscillation could be controlled by tuned Lecher lines. By suitable modification of the electrodes, frequencies of about 6000 MHz were produced by Kohl in Germany as early as 1928. This is the same rf source used at 600 MHz by Yagi and Uda in 1928 when they were developing the parasitic array. Two years later Dr. Esau used Barkhausen-Kurz transmitters and receivers for full duplex operation across the English channel on 1670 MHz.[3]

The efficiency of the Barkhausen-Kurz oscillator was very low because most of the oscillating electrons were intercepted by the grid – which often ran white hot. In 1921 A. W. Hull proposed a solution to the problem: *The magnetron, a device which didn't require a grid;* the electrons were kept in a circular orbit around the cathode by an external magnetic field (**fig. 2**). The original design received considerable modification, notably by Yagi and Okabe in Japan who split the anode

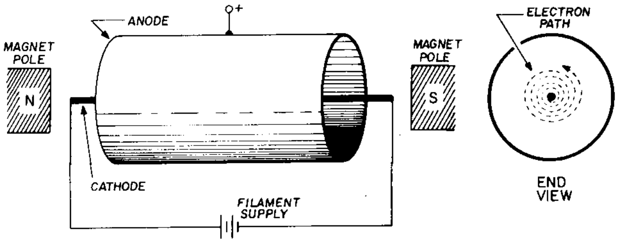


fig. 2. Basic magnetron. The electrons emitted by the cathode do not reach the positively-charged anode if the magnetic field is strong enough. If the magnetic field is properly adjusted, the electrons go into a spiral orbit around the cathode (end view, right), with the turns oscillating at a rate determined by an external tuned circuit. By 1930 devices such as this were being used to generate small amounts of microwave energy at wavelengths of 3cm (10 GHz).

into two or more parts and increased both frequency and power output. Although the magnetron offered considerable promise, it ran into some of the same difficulties as other devices – the dimensions of the component parts became so small at microwave frequencies that it was difficult to dissipate heat in the small space which was available. Nevertheless, by 1936 C. Cleeton and N. Williams at the University of Michigan were operating a magnetron at 50 GHz with very limited power output.

Obviously, before microwaves could be put into practical use, some way had to be found to generate useful amounts of power, but researchers were stymied on two fronts: the electrons moved too slowly, and the electrodes of the tube had to be as small as possible because they formed the capacitance of the resonant circuit. For the generation of liberal amounts of rf energy, however, the electrodes had to be as large as possible to dissipate the heat of electron bombardment. Thus, with existing devices, the requirements of microwave rf and high power could not be fulfilled simultaneously. In 1935 Dr. A. Arsenjewa-Heil and Dr. Oskar Heil proposed a unique solution to this mutual incompatibility. Why not, they asked, use the finite transit time of electrons to control the electron stream and derive the oscillation energy directly from the electron stream? In their proposed design the electrons didn't hit the electrodes at all, so the heating problem was completely avoided. Although the Heils were the first to describe the principle of velocity modulation, others had been thinking along the same lines, including Dr. William W. Hansen at Stanford University. Based on Hansen's calculations, Russell and Sigurd Varian put these ideas to work in 1937 when they built the first klystron, a device which uses transit time and the deceleration of bunched electrons across a vacuum gap to generate rf power (**fig. 3**).

In many respects the development of modern, solid-state microwave devices parallels advances in vacuum-tube technology in the 1920s and 1930s. Transistors suffer from electron drift-time problems at high frequencies, too, and it wasn't too many years ago that amateurs were hoping for a low-cost transistor that they could use successfully on 3.5 MHz. Vhf and uhf transistors are now commonplace, but only because the manufacturers have found ways of making the active region of the transistor wafer thin enough that electron transit time doesn't cause problems.* At microwaves, however, transit time is still the limiting factor, and this is where the analogy between microwave vacuum-tube and semiconductor development comes in: Like the Barkhausen-Kurz oscillator, magnetron, and klystron that preceded them, the operation of Gunn devices, avalanche diodes, and other solid-state microwave rf sources is also based on electron transit time.

In 1953 W. T. Read of Bell Labs proposed a multilayer diode for generating microwave power.[5] Read suggested that the finite delay between an applied rf voltage and the current generated by avalanche breakdown, and the subsequent drift of the generated carriers through the depletion layer of the diode junction would lead to negative resistance at microwave frequencies. The multilayer diode proposed by Read was very difficult to build, but in 1965 R. L. Johnston and his colleagues at Bell Labs experimentally verified Read's principle when they achieved a pulsed power output of 80 mW at 12 GHz from a silicon junction diode driven into avalanche.[6] Advances since 1965 have been so rapid that today avalanche-diode oscillators are established as one

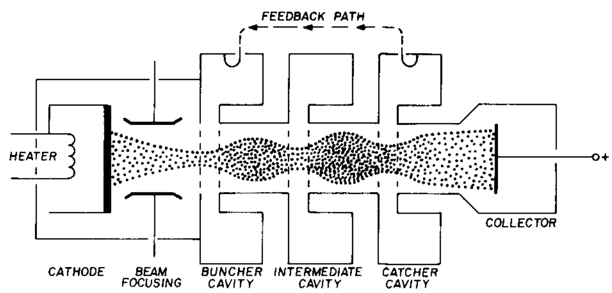


fig. 3. Klystron operation. Electrons are emitted by the hot cathode surface, formed into a beam, and drawn through the buncher cavity toward the positive anode. As the electron beam pours through the buncher grid it induces an rf voltage in the resonant buncher cavity which, for one half cycle, is in a direction that tends to speed up the electrons flowing through the gap; on the following half cycle the electric field tends to slow down the electrons as they cross the gap. This is called "velocity modulation." In the drift tube the electrons which have been speeded up tend to overtake the electrons which have been slowed down during the preceding half cycle, forming clumps or "bunches" of electrons in the drift tube. If the velocity of the electron beam and the length of the drift tube are properly adjusted, the bunched electrons will be completely formed by the time they reach the gap of the catcher cavity. The time between arrival of individual electron bunches coincides with one rf cycle. When a feedback path is provided, the klystron becomes a self-excited oscillator. Intermediate cavities serve to remodulate the electron beam and greatly increase the gain of the device.

*Although not discussed in this article, bipolar transistors, GaAs fets, varactor multipliers, and tunnel diodes are widely used on the microwave frequencies. The upper frequency limit for bipolar transistors is now about 4000 MHz, but GaAs fets (called "gas fets") have been used experimentally at frequencies as high as 15 GHz. Varactor multipliers are used in many microwave applications where good frequency stability is required, and tunnel-diode receivers are used commercially on the frequencies from about 4000 MHz to 15 GHz.

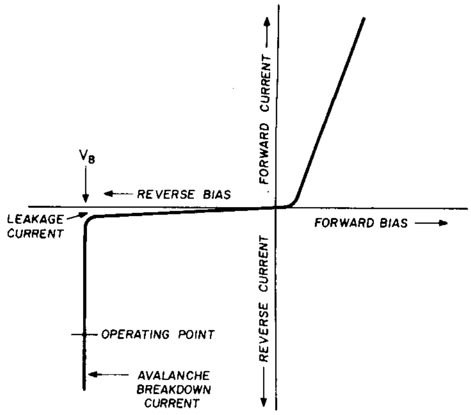


fig. 4. Voltage-current plot for typical junction diode. When reverse bias exceeds the breakdown voltage, V_B, diode is biased into the avalanche breakdown region where large numbers of electrons are generated by secondary emission (also called avalanche multiplication).

of the most important of the solid-state microwave power sources. Since the operation of the Read device is based on a combination of impact avalanche breakdown and electron transit-time effects, diodes of this general type are generally called IMPATT diodes from IMPact Avalanche and Transit Time.

IMPATT operation

The basic construction of an IMPATT diode is similar to that of any pn junction diode. Shown in **fig. 4** is a typical plot of dc current vs voltage for a pn junction diode. When the diode is forward biased, current increases rapidly for voltages above 0.5 volt or so. When the diode is reversed biased, however, a very small current called "leakage current" flows until the breakdown voltage, V_B, is reached. In ordinary rectifier diodes the breakdown voltage, V_B, determines the maximum PIV rating of the diode; if this rating is exceeded the diode will be destroyed.

When a pn junction is reverse biased a depletion region forms in the n-type region of the diode with its width depending on the applied bias. If the bias voltage is less than V_B the depletion zone acts like a nonlinear capacitor – this is the property used in varactors and tuning diodes. When the reverse voltage exceeds V_B by a small amount, the diode is biased into the avalanche region. In this region the small leakage current has a very high probability of creating additional electrons by the process of secondary emission or avalanche multiplication. Biasing some diodes into the avalanche region results in catastrophic failure, because once started, the avalanche current cannot be stopped. However, if the semiconductor material is properly doped, the avalanche process can be controlled (as it is in Zener diodes and avalanche rectifiers).

Fig. 5 is a schematic representation of the electron movement in a reverse-biased pn junction with a large number of electrons generated in the avalanche zone flowing into the *drift* zone. In this condition, a large

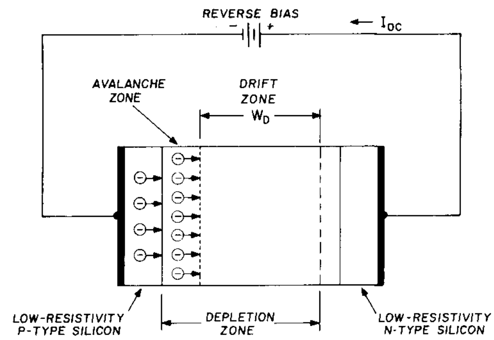


fig. 5. Schematic representation of a reversed-biased PN junction diode. When biased into the avalanche region, a large number of electrons are launched into the drift region. If the width of the drift zone, W_D, is such that the drift time of the electrons is about 37% of the time of one complete rf cycle, the device exhibits negative resistance because the output current will be 180° out of phase with the applied rf voltage (see fig. 6).

current can flow in the reverse direction with little increase in applied voltage. This is the avalanche breakdown current depicted in **fig. 4**. If, in addition to the bias voltage, an rf voltage exists across the depletion region of the diode (as it would be if the diode were mounted in a resonant cavity), under certain conditions the rf voltage induces an rf current that is out of phase with the applied voltage. If the rf current in the external circuit lags the rf voltage by more than 90 degrees, this is equivalent to a negative resistance.

In actual operation the IMPATT diode is biased just above the avalanche point (**fig. 4**) so the diode is biased into avalanche on positive swings of the rf voltage (**fig. 6A**). Since the number of electrons generated in the avalanche zone depends not only on the applied voltage but also on the number of charge carriers that are present, the avalanche current pulse continues to build

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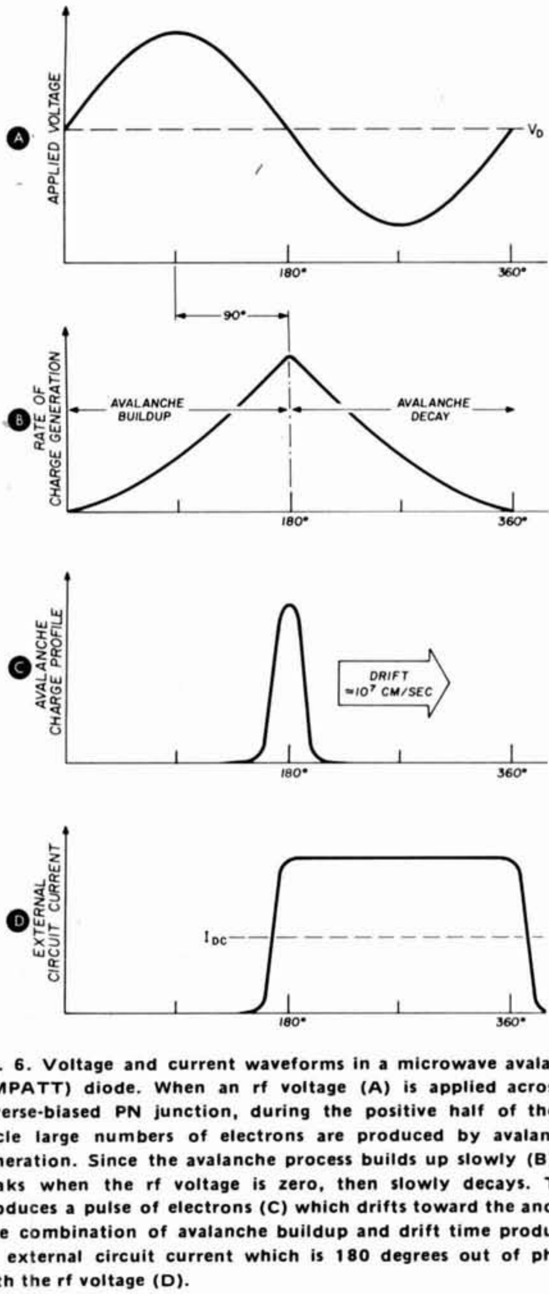


fig. 6. Voltage and current waveforms in a microwave avalance (IMPATT) diode. When an rf voltage (A) is applied across a reverse-biased PN junction, during the positive half of the rf cycle large numbers of electrons are produced by avalanche generation. Since the avalanche process builds up slowly (B), it peaks when the rf voltage is zero, then slowly decays. This produces a pulse of electrons (C) which drifts toward the anode. The combination of avalanche buildup and drift time produces an external circuit current which is 180 degrees out of phase with the rf voltage (D).

up even after the rf voltage has begun to drop (**fig. 6B**). This is because of the highly nonlinear nature of the avalanche generation process. In a properly designed IMPATT diode the excess charge slowly builds up in the avalanche region during the positive half cycle of the rf voltage, and reaches a sharply-peaked maximum in the middle of the rf voltage cycle when the rf voltage is zero. Thus the wave form of the avalanche current, in addition to being very sharply peaked, lags the rf voltage by 90 degrees.

The pulse of avalanche current is launched into the drift zone (**fig. 6C**) and slowly drifts to the right toward the positively charged n-side of the junction. The electrons drift through the semiconductor material at a nearly constant velocity (about 10^7 cm/second) so the time it takes them to pass through the drift zone is simply the width of the drift zone, W_D, divided by the velocity of the electrons, v

$$T = W_{D/v}$$

where T is the drift time. Since drift time is related to the frequency of operation, the width of the drift zone is carefully controlled during the manufacturing process.

While the pulse of electrons is drifting through the diode, they induce an approximate square wave of current in the external circuit as shown in **fig. 6D**. As can be seen in **fig. 7**, the combined delay of the avalanche process and the finite transit time across the drift zone has caused a *positive* current to flow in the external circuit while the rf voltage is going through its *negative* half cycle. Therefore the diode is delivering rf energy to the external circuit because of negative resistance.

The useful frequency range of the IMPATT diode is generally above 3000 MHz. Below this point the long transit times require a structure of such thickness that the breakdown voltage is very high. Most high-power IMPATT oscillators are used in the range from 5 to 13 GHz, although they have also been used successfully above 100 GHz. Most of the early IMPATTs were silicon types, but both germanium and gallium-arsenide (GaAs) have been used successfully. All types are noisy because avalanche breakdown is a noisy phenomenon, but GaAs types are somewhat less noisy than silicon devices. This is a problem in some applications (such as receiver local oscillators), but noise can be reduced substantially by proper circuit design. Another disadvantage is the relative high operating voltage (70 to 135 volts), and the requirement for a constant-current power supply. Operating efficiencies are on the order of 12 to 15 per cent, although careful construction, the use of GaAs, and the so-called double-drift structure can increase efficiency to 25 or 30 per cent.

The double-drift IMPATT diode has four layers instead of the usual three because an additional drift region is implanted in the diode (**fig. 8**). In the single-drift IMPATT described previously, the output current was the result of drifting electrons. The avalanche process, however, generates holes (positive charges) as

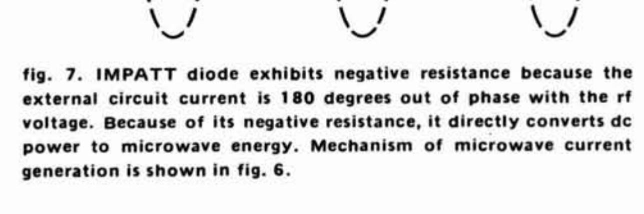


fig. 7. IMPATT diode exhibits negative resistance because the external circuit current is 180 degrees out of phase with the rf voltage. Because of its negative resistance, it directly converts dc power to microwave energy. Mechanism of microwave current generation is shown in fig. 6.

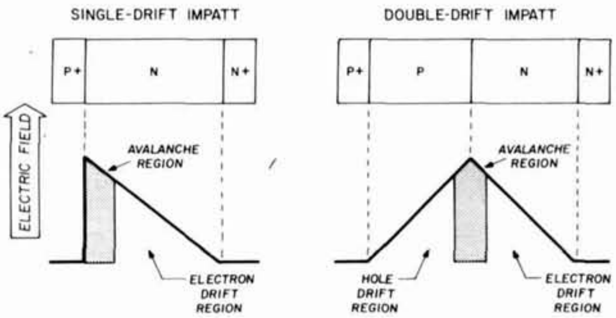
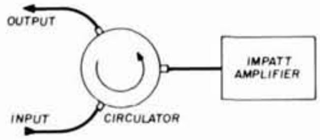


fig. 8. Construction of single- and double-drift IMPATT diodes. Efficiency is greatly increased in the double-drift diode because the holes (positive charges) generated in the avalanche region drift across the P-doped region in phase with the electrons, providing greater power output. In the single-drift IMPATT the holes are simply returned to the cathode.

well as electrons. In the single-drift IMPATT the holes simply are returned to the cathode – in the double-drift structure the holes drift across the added p-doped drift region in phase with the electrons, resulting in greater rf power outputs.

Even with operating efficiencies of 25 per cent, heat dissipation becomes the limiting factor when substantial amounts of rf power are required from an IMPATT diode. With good heatsinking techniques the power output can be increased as much as 20 per cent – diamond heatsinks, which have superb thermal characteristics, are being used extensively in commercial IMPATT applications. It should also be noted that IMPATT diodes are being used almost exclusively as amplifiers – few are used as power oscillators.* When used as an amplifier the IMPATT is coupled into the system through a circulator as shown below. The IMPATT amplifier has only one



REFLECTION AMPLIFIER

port so the circulator, which allows rf energy to propagate in only one direction, is required to isolate the input signal from the output signal. Operation is similar to that used in a parametric amplifier and is called "reflection amplification."

TRAPATT diodes

In 1967 researchers at RCA were trying to develop an avalanche diode that would provide operation around 1000 MHz. According to the drift-time theory, as noted above, IMPATT diodes could not be made to operate at that low frequency, but the engineers had hopes of exciting the diode into some other mode. Within a few months they had found a new mode of operation that had both good efficiency and high power output: 425 watts peak, pulsed output with an efficiency of about 25 percent.[7] As researchers continued to work with the new mode, they found a few diodes with efficiencies as high as 60 percent. They also worked on tuned circuits and eventually developed one that permitted continuous tuning from 900 to 1500 MHz. Since the operation of the high-efficiency diode didn't fit any then-known theory, they called it the "anomalous" mode.

Further work at Bell Labs with computer simulation led to the announcement that the high efficiency resulted from the creation of a trapped voltage plasma state between successive sweeps of the classical IMPATT cycle. According to the Bell Theory, this dense plasma then shielded the interior of the diode from the external voltage so the charges (electrons and holes) drift out of the diode at low velocities, causing very long transit times. This led to the acronym TRAPATT for TRApped Plasma Triggered Transit.

TRAPATT diode configuration built by RCA in 1970 that provided 1200 watts peak power output at 1100 MHz with about 25% efficiency. Five TRAPATT diodes are stacked in a 1N23 rectifier package.

Quantum physicists from Cornell University didn't agree with the theory offered by Bell Labs – they held that though a trapped plasma could undoubtedly be created in an over-driven diode, it was not fundamental to high efficiency. Their contention was that the parametric theory of Avalanche Resonance Pumping (ARP) had broader validity. Many researchers felt that Bell Labs' TRAPATT and Cornell University's ARP were actually two aspects of the same thing. RCA apparently preferred the TRAPATT model but stayed out of the battle of the acronyms. However, in a moment of humor during one heated debate, someone suggested that the original RCA terminology "anomalous" mode could stand for "A Non-Ohmic Maximum Allowable Large Output Uhf Source!"

In the early 1970s it was discovered that many ordinary silicon rectifier diodes could be made to oscillate in the TRAPATT mode,[8,9] but since this mode of operation is very fussy and requires tricky resonant circuit design to generate the required voltage waveform and suppress higher mode harmonics, the device has found limited applications – primarily in military L-band radar systems.

*Researchers in France have reported that the IMPATT diode operates with very high efficiency in diode frequency-multiplier circuits, but few details are available. This application may be widely used in the future.

Gunn devices

In 1963, when John Gunn of IBM was studying the *bulk resistance of a sample of n-type gallium arsenide*, he discovered that when the voltage impressed across the sample was raised above a certain point (**fig. 9**), the current became unstable and began to pulsate cyclically at microwave frequencies.[10] The mechanism which caused this to happen was a mystery at first, but Gunn suspected that a negative resistance was probably responsible, and suggested that a decrease in the mobility (velocity) of the electrons with an increase in applied voltage could account for the negative resistance. This was eventually proved to be the case.

Unlike most other materials, the electrons in gallium arsenide (GaAs) can be in one of two conduction bands, one with much higher electron velocity than the other. As the voltage across the GaAs in increased, more and more electrons are scattered to the low mobility band. This is shown graphically in **fig. 10**. Below the threshold point the current through the material is proportional to the applied voltage, so it behaves as a resistor. As the voltage is increased above threshold, however, sufficient electrons are displaced from the high mobility band that the net electron velocity through the GaAs begins to drop. Since the current through the material is proportional to electron velocity, this means a GaAs resistor (Gunn diode) will have a region of

Construction of Gunn diodes. Most of the package shown here is actually the heatsink — the ceramic-packaged device is only 0.025" (0.6mm) thick. To give you an idea of size, the stud is threaded 3-48 (M2.5).

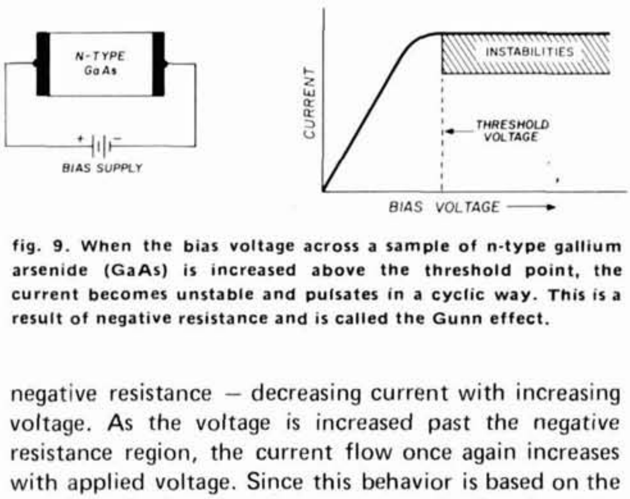


fig. 9. When the bias voltage across a sample of n-type gallium arsenide (GaAs) is increased above the threshold point, the current becomes unstable and pulsates in a cyclic way. This is a result of negative resistance and is called the Gunn effect.

negative resistance — decreasing current with increasing voltage. As the voltage is increased past the negative resistance region, the current flow once again increases with applied voltage. Since this behavior is based on the transfer of electrons from one conduction band to another, microwave oscillators of this type are often called Transferred-Electron Oscillators or TEOs.

Although the negative resistance of GaAs accounts for its current instability characteristics at certain bias levels, the oscillation at microwave frequencies requires further explanation. As was mentioned previously, when the applied voltage is below threshold, the GaAs behaves as a linear positive resistance; under these conditions the internal electric field is constant throughout the material *as shown in* **fig. 11A**. If the applied voltage is increased above threshold, many of the electrons entering at the cathode are entering faster than they leave. This leads to a high field buildup at the cathode with an accumulation of electrons on the cathode side and a depletion of electrons on the anode side (**fig. 11B**). The electric field throughout the rest of the material begins to fall to a value below threshold. The high field domain drifts *rapidly across the GaAs wafer* (**figs. 11C** *and* **11D**) *to the* anode where it is collected (**fig. 11E**). When the domain reaches the anode the bias supply again causes the field at the cathode to exceed the threshold level — a new domain is formed and the process repeats itself.

The current through the GaAs is lower during the transit time of the domain, and increases momentarily

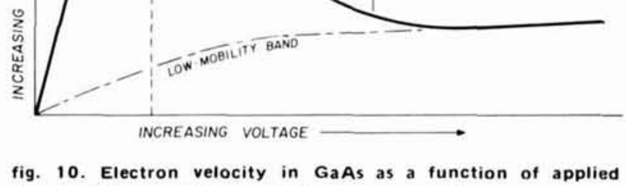


fig. 10. Electron velocity in GaAs as a function of applied electric field. As the electric field is increased, more and more electrons are scattered to the low mobility band, resulting in a net decrease in current flow through the material — this is equivalent to a negative resistance.

when the domain is extinguished at the anode. Thus the output is a series of narrow current spikes with a period equal to the transit time through the wafer (**fig. 11F**). The domain velocity is about 10^7 cm per second so the wafer of GaAs must be about 10 microns thick for operation at 10 GHz. Since the frequency of the output current pulses is a function of drift time, this is called the transit-time mode of operation. It is rarely used, however, because frequency tuning is nil and efficiency is very low.

In practical microwave circuits the Gunn device is mounted in a high-Q resonant circuit — this provides a considerable tuning range because the rf voltage in the circuit influences the formation of the field domain at the cathode by swinging the applied voltage above and below the threshold level as shown in **fig. 12**. If the bias is set so the rf voltage drops below the sustaining voltage, V_s, the field domain will be quenched; when the rf voltage rises above threshold, V_{th}, a new domain will form at the cathode. As the domain starts to drift across the GaAs, the rf voltage will increase to a maximum and then decrease until it, too, is quenched.

The delayed-domain mode occurs when the instantaneous rf voltage never falls below the sustaining value but does go below threshold for a part of the cycle. If the field domain reaches the anode when the applied voltage is between V_s and V_{th}, the formation of a new domain is delayed until the rf swings above threshold. Using these techniques and others, the frequency of the output current spikes will adapt to the resonant frequency of the external tuned circuit and can be tuned over very wide frequency ranges.

The greatest advantage a Gunn device has over IMPATT and TRAPATT diodes is its ability to operate over a wide band with less noise and lower bias voltages at equivalent frequencies — this is an important consideration for amateurs who want to build simple microwave systems that operate from batteries for portable,

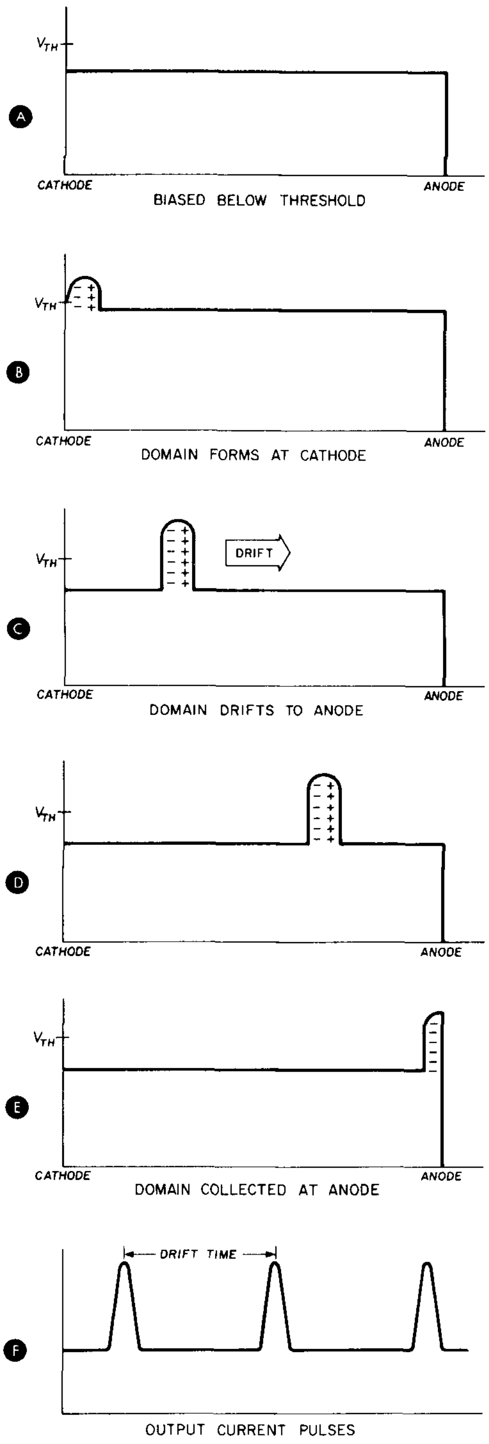


fig. 11. When a wafer of GaAs is biased below threshold, V_{th}, the material behaves as a linear resistance (A). A charge layer (domain) forms at the cathode when the material is biased above threshold (B), and drifts toward the anode at about 10^7 cm per second (C) and (D). Note that when the field domain forms at the cathode, the electric field in the rest of the material drops below the threshold level. When the domain is collected at the anode (E), the field in the material momentarily increases above threshold, a new domain forms at the cathode, and the process repeats itself. The current through the GaAs is lower during the transit time of the domain and increases when the domain reaches the anode, giving a series of sharp current spikes (F).

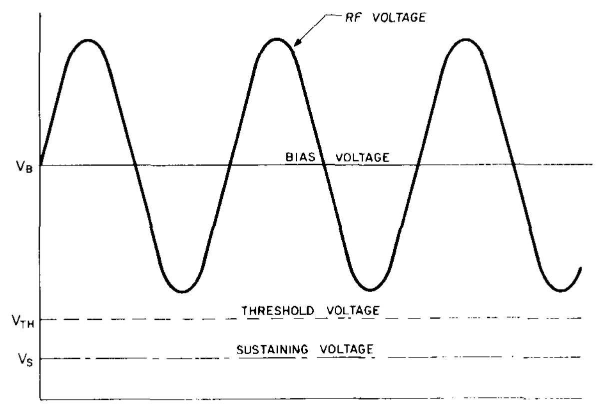


fig. 12. Operating mode of a Gunn diode in a high-Q resonant circuit is determined by the bias level, as shown here. If the diode is biased so the rf voltage drops below the sustaining voltage, V_s, the field domain will be quenched until the rf voltage rises above threshold, V_{th}, and the formation of a new domain at the cathode will be delayed (see text). Therefore, the frequency of the output current spikes will adapt to the resonant frequency of the tuned circuit and can be tuned over wide frequency ranges.

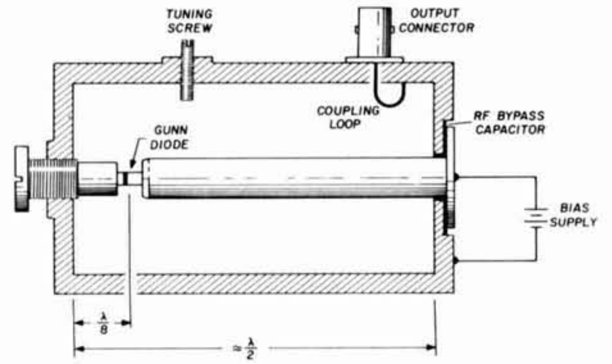


fig. 13. Simple Gunn-diode oscillator uses a half-wavelength coaxial cavity. Impedance matching is provided by the output coupling loop. This type of circuit can be tuned over an octave or more, but difficulties with oscillation at harmonic frequencies are common, and the coaxial cavity is more sensitive to temperature changes and load mismatches than waveguide resonators.

mountain-top expeditions. Moreover, Gunn diodes are easily frequency modulated and lend themselves to automatic frequency control (afc). Compared to tubes the Gunn device operates at lower temperatures and without a vacuum, factors that contribute to longer life. In fact, a number of manufacturers are predicting useful lives of well over 300,000 hours for CW devices (in case you don't want to figure it out, that's equivalent to about 34 years of 24-hour-per-day operation). On the debit side, the Gunn diode is less efficient and has lower power output then other solid-state microwave devices, but this is more than offset by its simplicity of operation, wide tuning range, and lower operating voltage.

Gunn oscillators

Many early attempts to build Gunn oscillators were not all that successful – the results were seldom reproducible. Sometimes the device refused to oscillate in the resonant circuit, or if it did oscillate it wouldn't be on the desired frequency, but this was new technology, fresh from the laboratory; the GaAs manufacturing process was still in its infancy so there were problems with the diodes, and nobody had any comparable experience to fall back on. Gunn diodes have improved over the years, and we now know what types of resonant circuits work best, but the plain truth of the matter is that not much actual circuit design takes place – not in the traditional sense anyway. As one designer at Microwave Associates pointed out recently, "Those who have experience with Gunn diodes usually take a rough cut at what they think will work, and then home in on the final layout by trial, error, and a good deal of sheer feel."

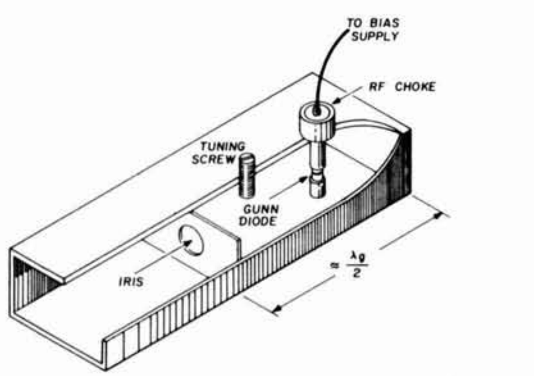
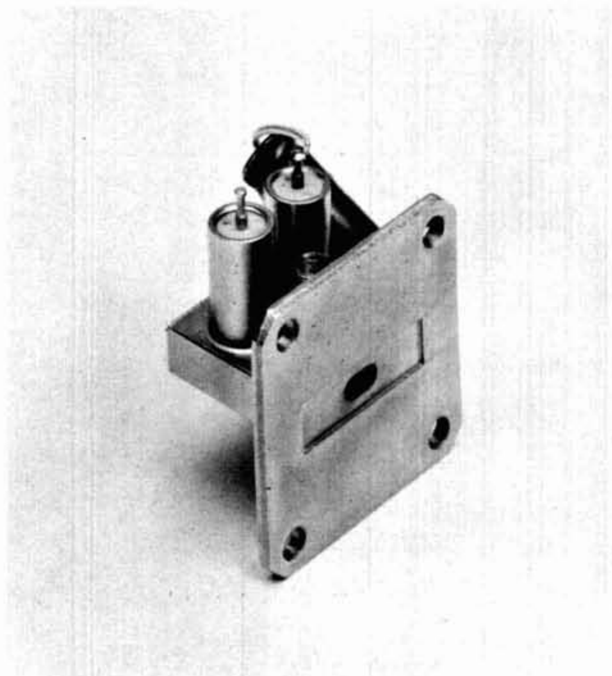


fig. 14. Simple waveguide resonator for Gunn-diode oscillators. In this circuit the microwave energy from the Gunn diode is coupled into the cavity with a post mounted between the narrow dimension of the waveguide. The size of the opening (iris) is optimized for maximum power output and isolation from impedance mismatches. The rf choke requires careful design for minimum rf loss.

One of the simplest forms of Gunn oscillator circuits is shown in **fig. 13**. Here, the diode is mounted at one end of a half-wavelength coaxial cavity which is adjusted



Microwave Associates varactor-tuned Gunn oscillator for the amateur 10-GHz band. Terminals are for the Gunn power supply and varactor bias. Power output is 20 milliwatts.

to the desired operating frequency with a tuning screw. The location of the output coupling loop determines the load impedance presented to the diode. This type of resonator is easy to build and can be tuned over a very wide frequency range, typically an octave, but it has several disadvantages. For one thing, the Q is relatively low and the diode may want to oscillate at a harmonic frequency. In comparison with waveguide cavities, the coaxial resonator is also more sensitive to temperature changes and load mismatches.

For most applications a much better choice is the post-coupled rectangular waveguide cavity shown in **fig. 14** separated from the output waveguide by a coupling iris. The size of the iris is determined experimentally for the best compromise between maximum power output table 1. Operating characteristics of the Microwave Associates MA-87127 series 10-GHz Gunnplexers.

RF center frequency	10250 MHz (4V varactor bias)
RF power output	20 mW
Tuning: mechanical	± 100 MHz
electronic	60 MHz minimum
Frequency stability	−350 kHz/°C maximum
RF power vs temperature and tuning voltage	6 dB maximum
Frequency pushing	15 MHz per volt, maximum
Noise figure	12 dB maximum
Input requirements:	
Gunn voltage	+10 Vdc typical
Gunn current	500 mA maximum
Tuning voltage	+1 to +20 volts
Rf connectors	mates with UG-39/U waveguide
Operating temperature	−30 to +70 °C

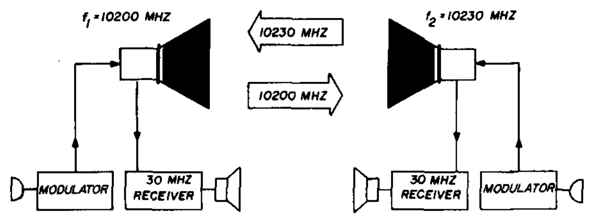


fig. 16. Gunnplexer operation. Since the same oscillator is used as both a transmitter and local oscillator for the mixer, the i-f at each end of the link must be at the same frequency, and the frequencies of the Gunn oscillators must be separated by the i-f. In the example shown here the Gunnplexer at one end of the link is tuned to 10200 MHz — 30-MHz i-f receivers are used so the Gunnplexer at the other end must be tuned 30 MHz higher or lower (10230 or 10170 MHz).

and isolation from changes in diode impedance and load. The tuning rod may be either metal or a low-loss dielectric. The diode must be properly decoupled from the bias supply to minimize any rf resonances in the bias circuit and to prevent any rf loss. None of these things is trivial, so if you're interested in building your own Gunn oscillators for the 10-GHz amateur band, I suggest that you try one of the proven designs published in the RSGB's *VHF - UHF Manual*.[11]

If you don't have any previous experience with microwave circuits, you may find it easier and less frustrating to purchase one of the new 10-GHz *Gunnplexers* which are being offered to amateurs by Microwave Associates.* These Gunn-oscillator transceivers provide 20 mW of output power, include a built-in low-noise Schottky mixer diode, and are provided with varactor tuning. A ferrite circulator isolates the receiver and transmitter functions. The electrical specifications of the Gunnplexer are listed in **table 1**; a cut-away view of the transceiver is shown in **fig. 15**.

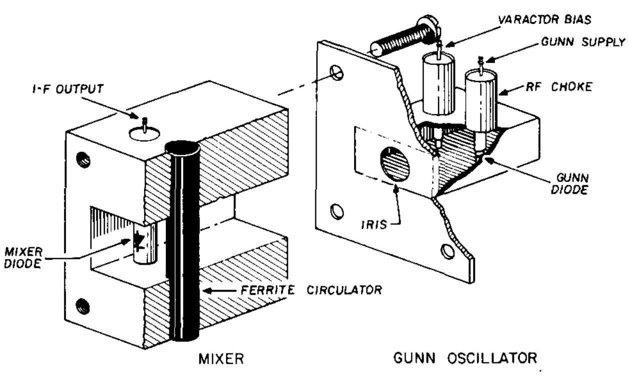


fig. 15. Cutaway view of the 10-GHz Microwave Associates Gunnplexer. The post-coupled Gunn diode is tuned to the desired frequency with the dielectric tuning screw, and the rf energy is coupled out through an iris. The ferrite circulator couples a small amount of energy into the Schottky mixer diode and isolates the transmit and receive functions. Mixer injection can be adjusted with the small screw mounted in front of the circulator. A horn antenna provides 17 dB gain.

In the Gunnplexer the Gunn oscillator provides both the transmit power and LO injection for the mixer diode. Therefore, the i-f used at each end of a communications link must be tuned to the same frequency, and the frequencies of the Gunn oscillators at each end of the link must be separated by the intermediate frequency. This is the same system used with klystron polaplexers and is shown in **fig. 16**. If the i-f is at 30 MHz, for example, and one Gunn oscillator is tuned to 10200 MHz, the other oscillator is tuned 30 MHz higher (or lower) to 10230 MHz (or 10170 MHz). Gunnplexers can also be used for two-way communications with stations which use polaplexers or separate 10-GHz transmitters and receivers. If a polaplexer is used at one end of the link, however, you should expect about 3 dB loss because of the difference in polarization.

All you need to put the Gunnplexer on the air is a well regulated 9-volt power supply (200 mA maximum), a bias supply for the varactor, an fm receiver, and a microphone and speech amplifier. This is the system used by W1HR and W1SL in their first two-way communications with Gunnplexers. Later you may want to add automatic frequency control (afc) or a phase-locking system, but this isn't necessary to get started.

The 30-MHz i-f has been the standard for amateur microwave communications for a number of years, but there's no reason why you can't use a standard 88-108 MHz fm broadcast receiver as a tunable i-f. If you choose to go this route, be sure to pick a frequency that's not occupied by a nearby fm transmitter — otherwise you may have problems with i-f feedthrough. In metropolitan areas it may be impossible to find a clear

*The Microwave Associates MA87127 *Gunnplexer* is a complete 10-GHz transceiver consisting of a Gunn oscillator, tuning diode, detector, and circulator and is priced at $85. Also available is the MA87108 which consists of the Gunn oscillator and tuning diode ($60), and the MA87140 which includes a complete transceiver and a 17-dB gain horn antenna ($108). Two complete transceivers with horn antennas, part number MA87141, are priced at $185. Write to Microwave Associates, Inc., Burlington, Massachusetts 01803 for the name of your nearest sales representative.

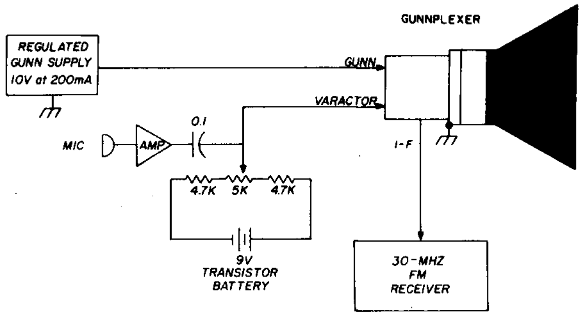


fig. 17. Basic system used by W1HR and W1SL for short-range communications with the 10-GHz Gunnplexers. Output frequency is adjusted ±30 MHz with the 5k potentiometer across the 9V varactor bias supply.

channel, but most fm receivers can be "tweaked" slightly so they will tune to a clear spot above or below the fm broadcast band. If you use an fm broadcast receiver you won't be able to work stations using a 30-MHz i-f, but this is an inexpensive way to get started, and you can always to to 30-MHz i-f later.

If you decide to use a 30-MHz i-f, military surplus i-f strips are inexpensive, or you can use a tunable fm receiver such as the old Hallicrafters S-36 or S-27, or the military surplus BC683. W1SL and I used 35-year-old S-36 receivers in our initial experiments, but are planning to build some solid-state replacements in the near future. With the wide availability of ICs designed for rf amplification and fm demodulation, this should be a relatively easy task.

communications range

One of the first questions you're probably asking is what kind of communications range can I expect with a 20 mW Gunnplexer system? As shown in **fig. 18**, this is a function of the bandwidth of the i-f system. This graph assumes a noise figure of 12 dB at 10 GHz which should be no problem with a low-noise (about 1 dB) i-f strip, and 17-dB horn antennas at each end of the link. For the bandwidth of an fm broadcast receiver, 240 kHz, the *maximum* line-of-sight range is about 25 miles (40 km). However, this graph is based on "threshold" (the beginning of reception of intelligible speech) and allows no margin for fading due to rainfall, multipath propagation, or other environmental effects. Therefore, for a practical system with good signal-to-noise ratios, somewhat less range should be expected. Range can be increased considerably by using parabolic reflectors, as shown by the dashed lines in **fig. 18**, but this entails additional cost and you may have trouble getting the antennas properly lined up.

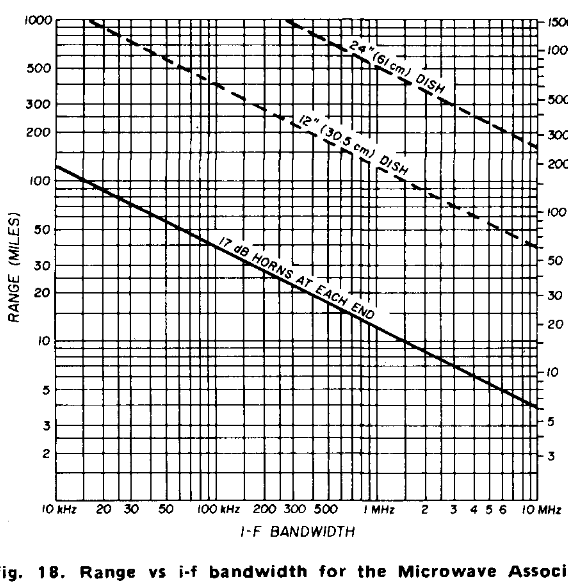


fig. 18. Range vs i-f bandwidth for the Microwave Associates 10-GHz Gunnplexers. Dashed lines show increased range available with 12-inch (30cm) and 24-inch (61cm) parabolic reflectors.

Range can also be improved by using a narrower i-f bandwidth, but this requires the use of automatic frequency control or a phase-lock arrangement. Since the drift characteristics of the Microwave Associates Gunnplexer is about −350 kHz per °C maximum (downward drift with increasing temperature), it's doubtful that the 240-kHz bandwidth of an fm broadcast receiver would be practical without continually adjusting varactor bias or using afc. The basic afc system is shown in **fig. 19**. To prevent the two Gunnplexers from chasing each other all over the band, only one end should use afc of the Gunn diode. Afc of the i-f LO is permissible at both ends of the link.

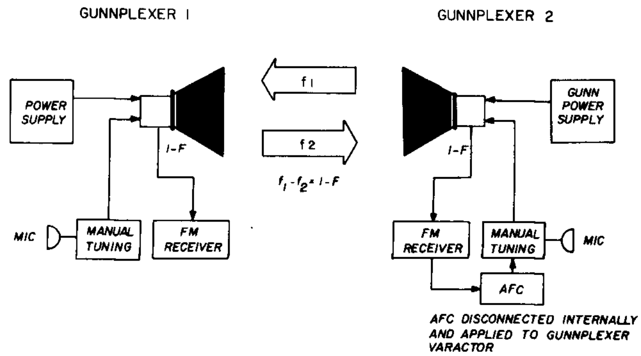


fig. 19. Basic arrangement for automatic frequency control (afc) of the Gunnplexer. Manual frequency control is used at one end of the link; afc at the other end of the link is derived from the fm receiver and applied to the varactor.

Stability can also be improved by placing the Gunnplexer in an insulated box and heating it with a lightbulb or other heat source (such as a power resistor). For maximum temperature stability, a proportional temperature control system is suggested (reference 12 describes a proportional control circuit for crystal ovens that could be easily adapted to the Gunnplexer).

For an idea of what can be accomplished with low power on 10 GHz, consider the new world's record on this band which was set in August, 1976, by GM3OXX in Scotland, and G4BRS, in Cornwall, England — a distance of 324 miles (521 km). Both stations used 10 mW Gunn oscillators with parabolic reflectors – a 24 inch (61cm) dish in Scotland and a 30-inch (76cm) reflector in England. This was their ninth try over this particular path, so it wasn't as easy as it sounds, but it should give you an idea of what can be done with simple equipment, patience, and a *lot of persistence*.

phase locking

One of the best ways to improve communications range with the Gunnplexer is to decrease i-f bandwidth but this can only be done by phase locking the transmitter to a stable crystal oscillator. I don't have any practical, tried and true circuits to offer yet, because I don't know of any amateurs who have built a phase-locked Gunnplexer system, so the following are merely suggestions.

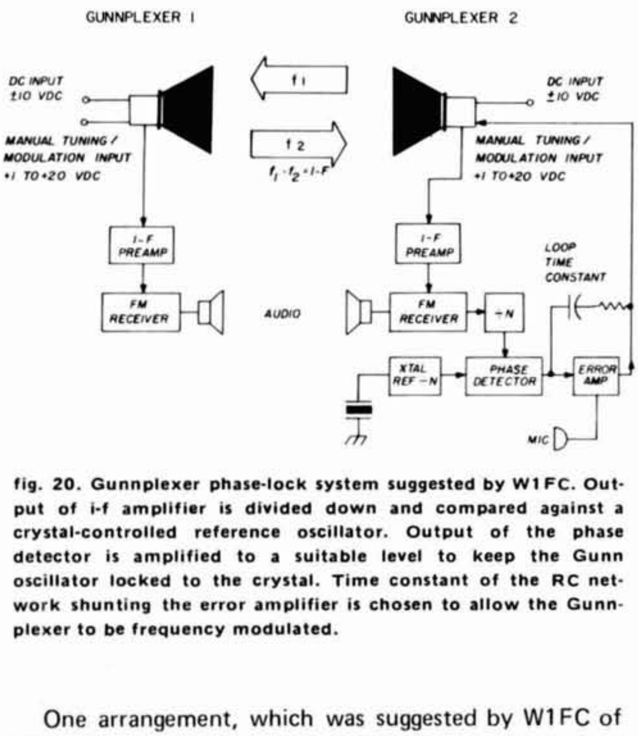


fig. 20. Gunnplexer phase-lock system suggested by W1FC. Output of i-f amplifier is divided down and compared against a crystal-controlled reference oscillator. Output of the phase detector is amplified to a suitable level to keep the Gunn oscillator locked to the crystal. Time constant of the RC network shunting the error amplifier is chosen to allow the Gunnplexer to be frequency modulated.

One arrangement, which was suggested by W1FC of Microwave Associates, is shown in **fig. 20.** In this system the output of the i-f amplifier is divided digitally and phase compared against a stable crystal reference. The dc output information from the MC4044 phase detector is amplified and fed to the Gunnplexer tuning varactor. The time constant of the RC network which shunts the dc error amplifier permits the Gunnplexer to be frequency modulated. Without the RC circuit, the phase-lock system would tend to cancel the modulation.

Another, more complex phase-lock circuit, based on the San Bernadino Microwave Society's *Rocloc* system for klystron polaplexers,[13] is shown in **fig. 21.** In this circuit a small portion of rf from the Gunnplexer is mixed with the output from a harmonic multiplier – the output from the mixer is then phase compared with a low-frequency crystal-controlled source. The rest of the system is similar to that shown in **fig. 20.**

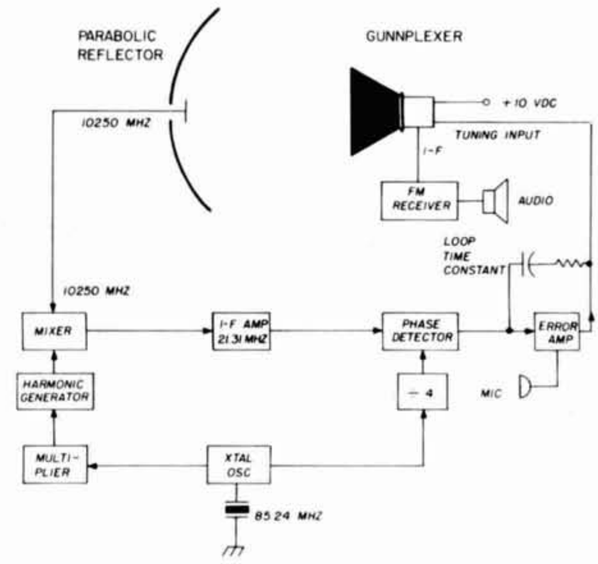
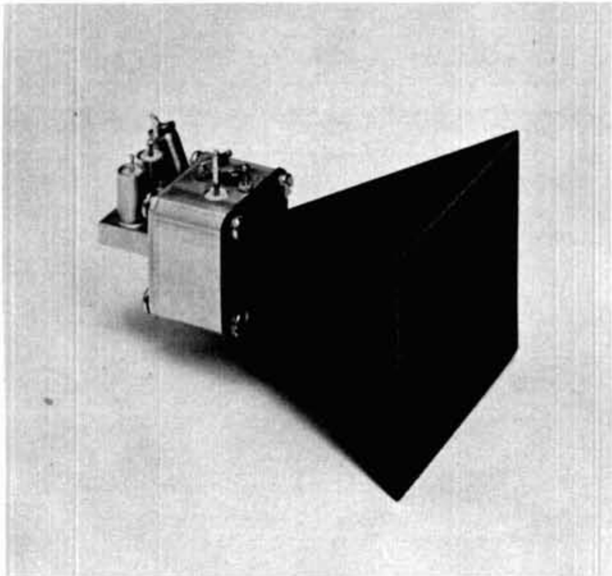


fig. 21. Proposed phase-lock system for a Gunnplexer operating on 10250 MHz. In this system the 120th harmonic of the 85.24-MHz crystal oscillator is mixed with a sample of rf energy from the Gunnplexer. The difference output is amplified, then phase compared with a signal derived from the crystal oscillator. The error voltage is amplified and fed to the tuning varactor in the Gunnplexer. With proper design, frequency stability of the Gunnplexer will be the same as the crystal reference oscillator.

Phase locking can also be used to good advantage in the receiver as a tracking filter. Whereas the threshold of intelligible speech in a conventional fm system occurs when the fm signal is about 10 dB above the noise, a phase-locked tracking filter can linearly demodulate fm signals buried 20 to 30 dB in the noise.[14] One IC on the market which was designed specifically for this task is the Exar XR-215.

Microwave Associates 10-GHz Gunnplexer and 17-dB horn antenna. Receiver section is housed in waveguide section machined from large block of metal. This improves thermal stability of the unit.



Microwave Associates 10-GHz Gunnplexer.

i-f circuits

Good afc capture range requires a wider bandwidth than optimum signal processing, so a dual i-f system is recommended. One suggestion here is an input at 30 MHz followed by conversion to 10.7 MHz for signal processing. A parallel i-f at 30 MHz would be used for afc control. The 30-MHz preamp should have a noise figure on the order of 1 dB and should be impedance matched to the Schottky mixer in the Gunnplexer (about 200 ohms). The capacitance of the mixer diode (approximately 27 pF) can be resonated with an rf choke connected from the i-f output terminal to ground (use 1 μH at 30 MHz, 0.33 μH at 100 MHz).

radiation hazard

Although 20 mW isn't very much power, remember that it's concentrated at the small, open end of the waveguide so power density is about 6.2 mW per square cm. This is considerably higher than OSHA's 1 mW/cm^2 safety limit. Fortunately, rf power density falls off to safe levels a few inches (15 cm) away, but remember that your eyes are especially susceptible to damage from rf radiation — *never* look into the open end of a Gunnplexer when it's operating.

summary

The amateur microwave bands have been badly neglected by most amateurs, but this is one area where amateurs can still make contributions to modern rf technology. Presented here are the basic requirements for a modern, solid-state, amateur microwave system, and some suggestions for putting your own Gunnplexer on the air. In future months we will try to publish practical, successful circuits — if you have suggestions for improvements, or have solved a problem that may give others trouble, we would like to have the opportunity to publish it in *ham radio*.

I would like to thank Dana Atchley, W1CF, of Microwave Associates for making Gunnplexers available to the amateur community, and Fred Collins, W1FC (ex W1FRR), and Dr. Ron Posner (ex K6DJB) for their circuit suggestions.

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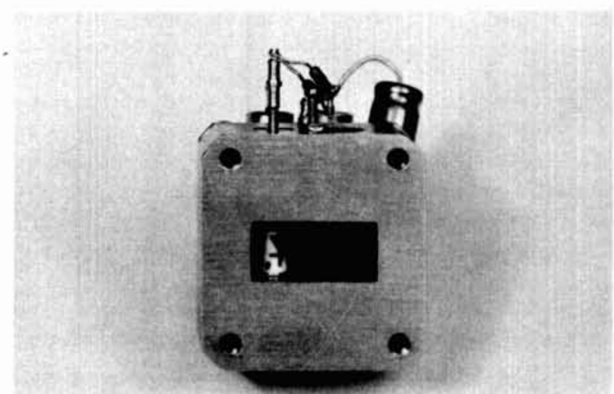

Head-on view of the Gunnplexer showing the mixer diode, left, and ferrite circulator (black cylinder to right). The small screw which protrudes through the top of the waveguide is used to adjust mixer injection.

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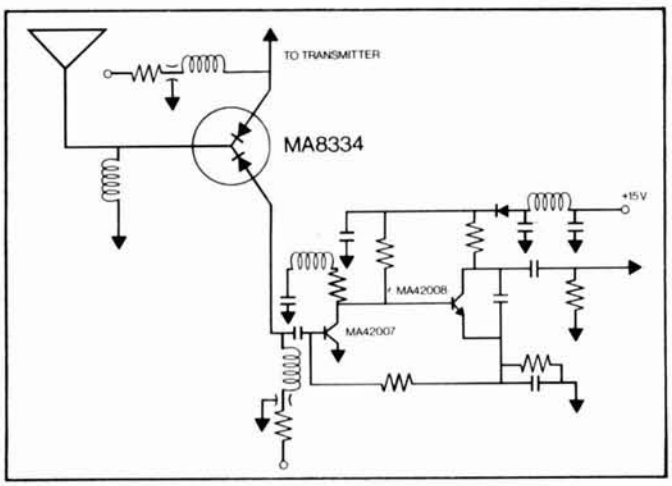

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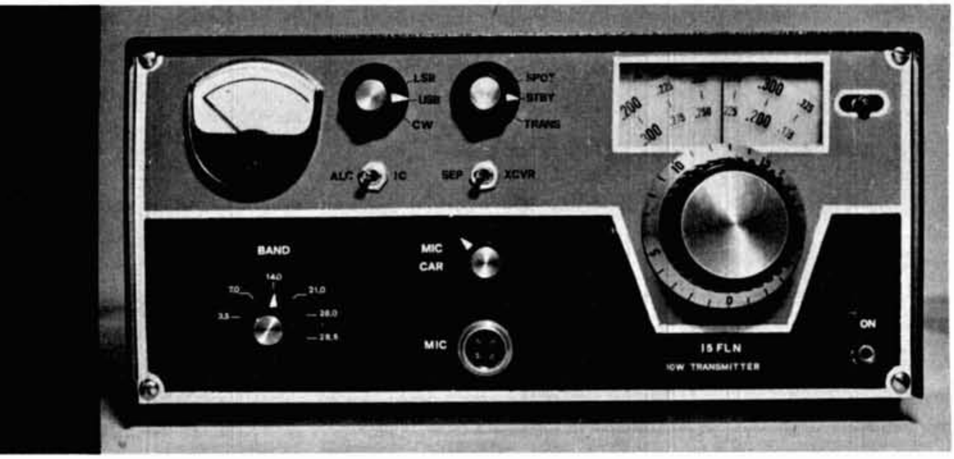


solid-state five-band transmitter

A companion transmitter to the I5TDJ receiver features all bands, solid-state circuitry, and 10-watt output

On the air, amateurs speak proudly of using the "S-Line" or the "Drake Line." To complement the receiver I described last year in these pages,[1] I built a companion five-band solid-state transmitter. Thus I felt I could proudly speak of using the "TDJ Line!"

I actually started the design in 1973 when high power rf transistors were not available in Italy. Thus, the output is only a modest 10 watts. None the less, I've made contacts with all continents and had a nice QSO on CW and ssb with JD1ACH during the Ogasawara (Bonin Island) DXpedition.

The transistors I used are a mixture of Phillips and U.S. types, collected over the years. Despite that, the 3rd-order IMD measures −28 dB below one tone of a 2-tone test. Output is actually greater than ten watts on all bands, with no tuning. A SPOT switch and carrier insertion for linear amplifier tuneup are included for convenience. Again, my friend I5FLN has duplicated my circuitry.

circuit description

The transmitter was designed to transceive with my receiver; the rf mixing scheme being identical. In the block diagram, **fig. 1**, note that the two modules, vfo and vfo converter, are common to both units. The builder has three options: you may use these two modules in the receiver, you can duplicate just the vfo module, or you can duplicate both modules and have a completely self-contained transmitter with frequency control from either unit – just like Drake. Don't forget a light to indicate which dial is operative!

In the ssb generator module, shown in **fig. 1**, the microphone signal is amplified in two stages and then applied to a diode-ring balanced modulator. The carrier is initially generated at 454 or 455 kHz, for CW/ssb. Undesired products are attenuated and suppressed by adjustment of the modulator and also FL1, a 2.4-kHz mechanical filter. The output from the filter is applied through a buffer to a balanced mixer comprised of two fets. Another pair of crystals at 8545 and 9455 kHz provide the signal, which is simultaneously applied to the fet mixer, for USB or LSB selection. The mixer output, after a three-section LC filter, is now a 9-MHz ssb signal.

In the next module, the 9-MHz signal is buffered and applied to a second diode ring mixer for addition to the signal from the vfo converter. Five wide-band LC filters at the output of the diode mixer select the desired band. The combination of diode mixer and LC filters again make for excellent suppression of undesired mixer products. The signal is now fed to a wide band amplifier and then to the power amplifier (PA) module.

Two transistors are used for the driver in the power amplifier module. The final transistor operates as a linear amplifier and has been protected against thermal runaway. No mismatch protection was deemed necessary. On several occasions the antenna was not connected, and no damage has resulted to date! In order to make the

By Piero Moroni, I5TDJ, Cosseria 10, Florence, Italy 50129 (translated by Joe Darmento, W4SXK)

most of the modest power available, a tuned circuit is used to match the final transistor to a 50-ohm output. Each circuit need only be set once during the initial adjustments. A sample of the output is rectified and used for alc. The 0-1 mA meter reads the supply voltage or may be switched to read alc voltage and thus monitor the output.

should be selected for equal forward resistance at 1 mA of current. Crystals Y1 and Y2 provide the 454/455-kHz carrier. On CW, the 454-kHz crystal causes the carrier to be transmitted 1 kHz above the vfo frequency. No problem – your contact receives a 1-kHz signal in the usb position of his receiver. R1 and C1 are used to null the carrier during the initial adjustment.

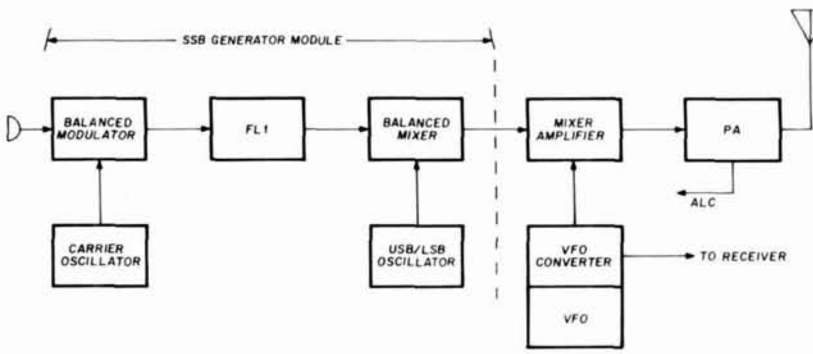
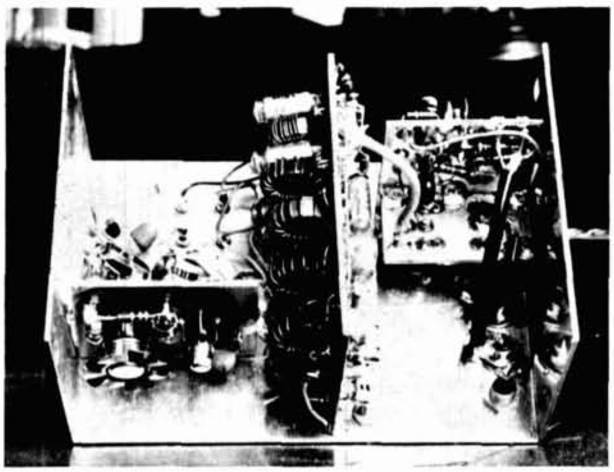


fig. 1. Block diagram of the solid-state transmitter. The vfo and mixing scheme are identical to the one used in the receiver described by the author in the October, 1975, issue of ham radio.

Except for the auxiliary circuit, each module was enclosed in a standard aluminum box, 3 x 4¼ x 2 inches (77x107x49mm). The PA module is slightly larger, 3 x 5.7 x 2 inches (77x145x49mm). Each module may be arranged as desired to fit a commercial case, as done by I5FLN. All circuitry is mounted on single-sided epoxy-glass PC boards. I used an isolated pad drill rather than printed-circuit technique; it's easier and quicker for a one-time project. Signal interconnections are made with phono jacks and plugs. All other interconnections enter the modules through 1500-pF feedthrough capacitors. Points designated as +12 always have power applied; +12T implies voltage on with PTT. When winding the toroids, make sure you twist the wires 7 to 8 turns per inch (2 to 3 turns per cm).

ssb generator module

Referring to **fig. 2**, Q1 and Q2 amplify the microphone signal to a few hundred millivolts for application to the balanced modulator. The four 1N270 diodes



Mixer and amplifier module. The individual band filters are mounted on the vertical circuit board. The balanced mixer is on the left and the output circuits on the right.

The gain of Q3 is controlled by the alc developed from the final transistor. Q6 and Q7 form the balanced mixer for signal conversion to 9 MHz. R2 is the balance pot, and when properly adjusted the mixer, in conjunction with L1, L2 and L3, will provide good attenuation of spurious products without resorting to an expensive commercial filter. One precaution, do not mount any oscillator components near the ends of the mechanical filter or you will not obtain good carrier suppression. Component placement is shown in the photograph.

mixer amplifier module

In the schematic, **fig. 3**, transistor Q1 is a buffer amplifier; the 100-ohm potentiometer adjusts the 9-MHz signal level into the diode mixer. Unfortunately, this type of mixer attenuates the desired signal approximately 6 dB and requires about 5 mW of oscillator injection. However, it does attenuate the oscillator signal at least 30 dB with satisfactory performance from 0.5 to 50 MHz. This latter characteristic is eminently desirable to provide constant output on all bands. Thanks to these characteristics the simple filters on the output are perfectly adequate. Except for component values these five filters are all identical to the 3.5 to 4 MHz filter shown in detail in the schematic. The filters are diode selected by the bandswitch. With feedback to improve linearity, the signal is amplified to about 50 mW by Q2 and Q3.

Components for this module are mounted on three small boards approximately 3/4 x 1 inch (15x25mm). One board holds the buffer and mixer, the second the 5 filters, and the third board contains Q2 and Q3. Separating the boards is helpful in home brewing. In case you make a fatal error, you ruin only a small part of your work.

power amplifier module

Two transistors in push-pull, **fig. 4**, raise the signal level to 0.5 watt. CR1 is in thermal contact with one transistor to maintain collector currents within safe

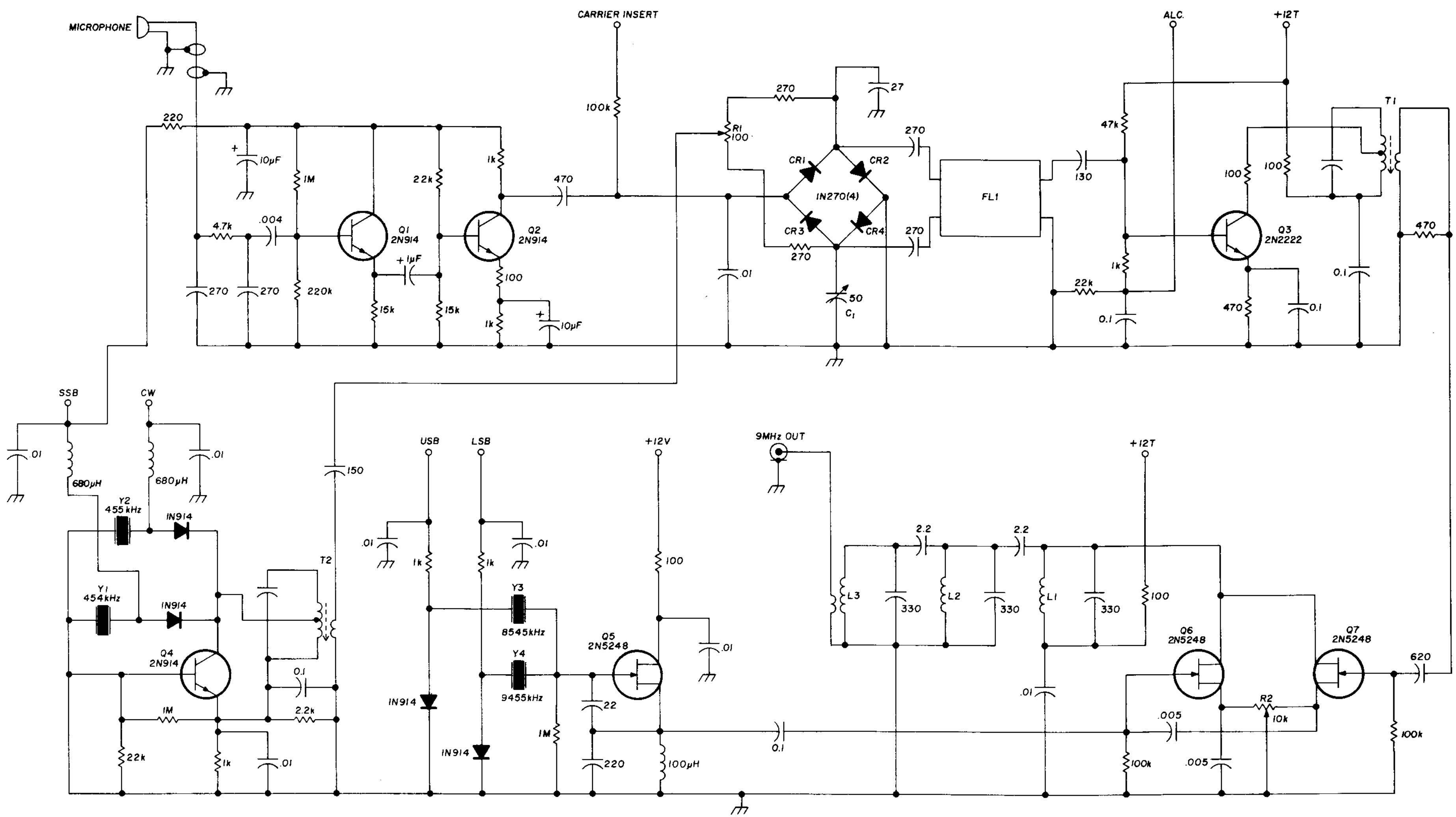


fig. 2. Schematic diagram of the ssb generator. C1 is a 50-pF (ARCO) compression trimmer. FL1 is a Collins F455-Z7 filter. L1 through L3 are each 15 turns of no. 22 AWG (0.6mm) wire wound on a Micrometals T44-6 core. L3 also has a one-turn link. Each coil should be 0.95 μH. T1 and T2 are 455-kHz transistor i-f transformers (Miller 2031). The crystals are parallel resonant with 32-pF load capacitance.

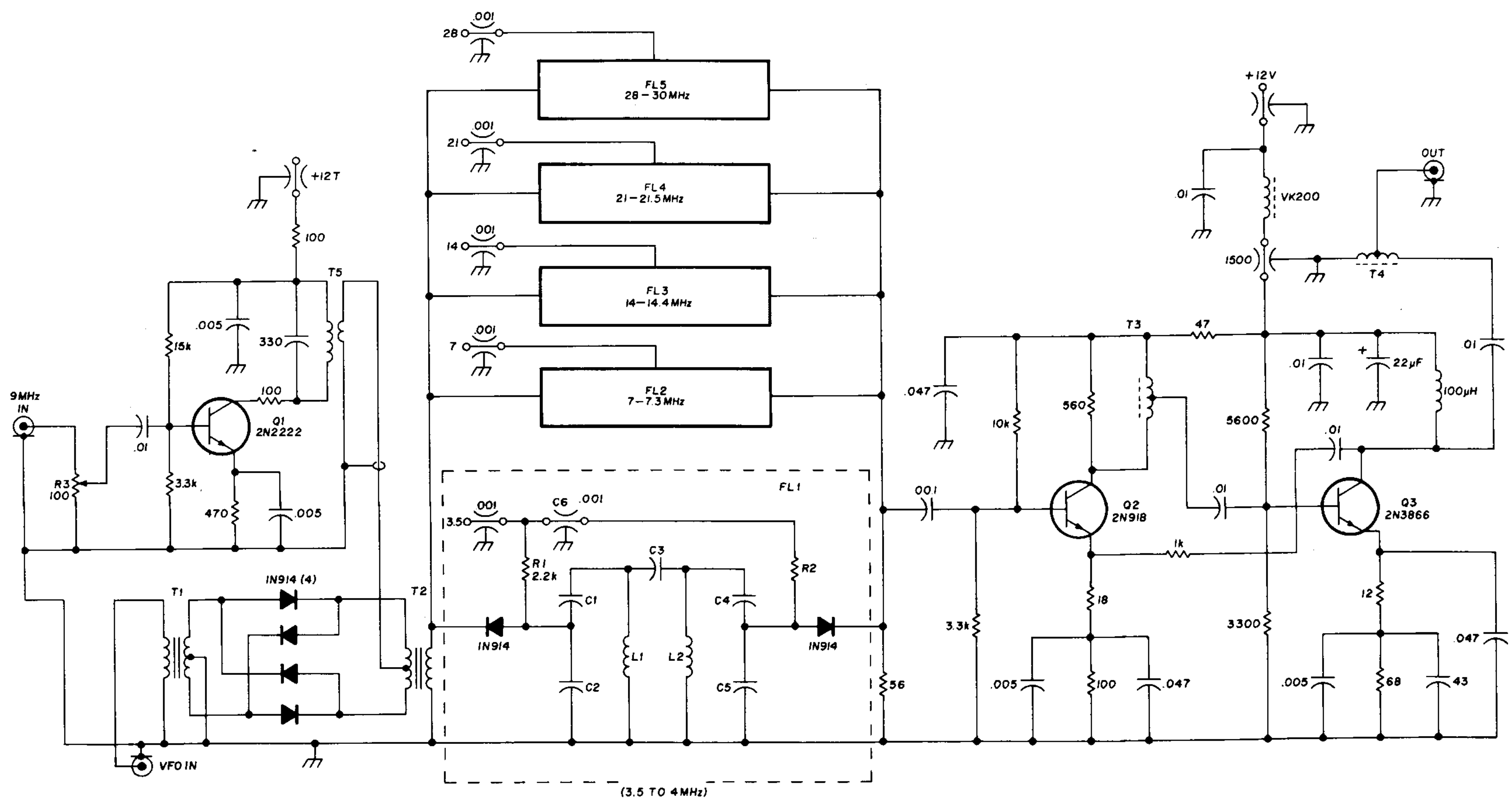


fig. 3. Mixer and amplifier module schematic diagram. T1 and T2 are each 10 turns of no. 32 AWG (0.2mm) trifilar wound on a hi-μ core (Indiana General CF-2). T3 and T4 are identical to T1 except they are bifilar wound. T5 is wound on an Amidon T44-6 core. The primary is 15 turns of no. 24 AWG (0.5mm) with a secondary of 2 turns wound over the ground side of the primary. The rf choke, VK200 19/4B, is available from Elna Ferrite Labs, Woodstock, New York 12498.

	C1, C4	C2, C5	C3	L1, L2
FL1	500 pF	3000 pF	39 pF	3.4 µH, 30 turns no. 24 AWG (0.5mm)
FL2	330 pF	3000 pF	8.2 pF	1.6 µH, 20 turns no. 24 AWG (0.5mm)
FL3	150 pF	1600 pF	3.3 pF	0.9 µH, 15 turns no. 22 AWG (0.6mm)
FL4	100 pF	1000 pF	3.3 pF	0.6 µH, 13 turns no. 22 AWG (0.6mm)
FL5	68 pF	430 pF	3.3 pF	0.5 µH, 11 turns no. 20 AWG (0.8mm)

Note: The coils for FL1 to FL3 are wound on Amidon T50-6 toroids and FL4 and FL5 coils are wound on Amidon T50-10 cores.

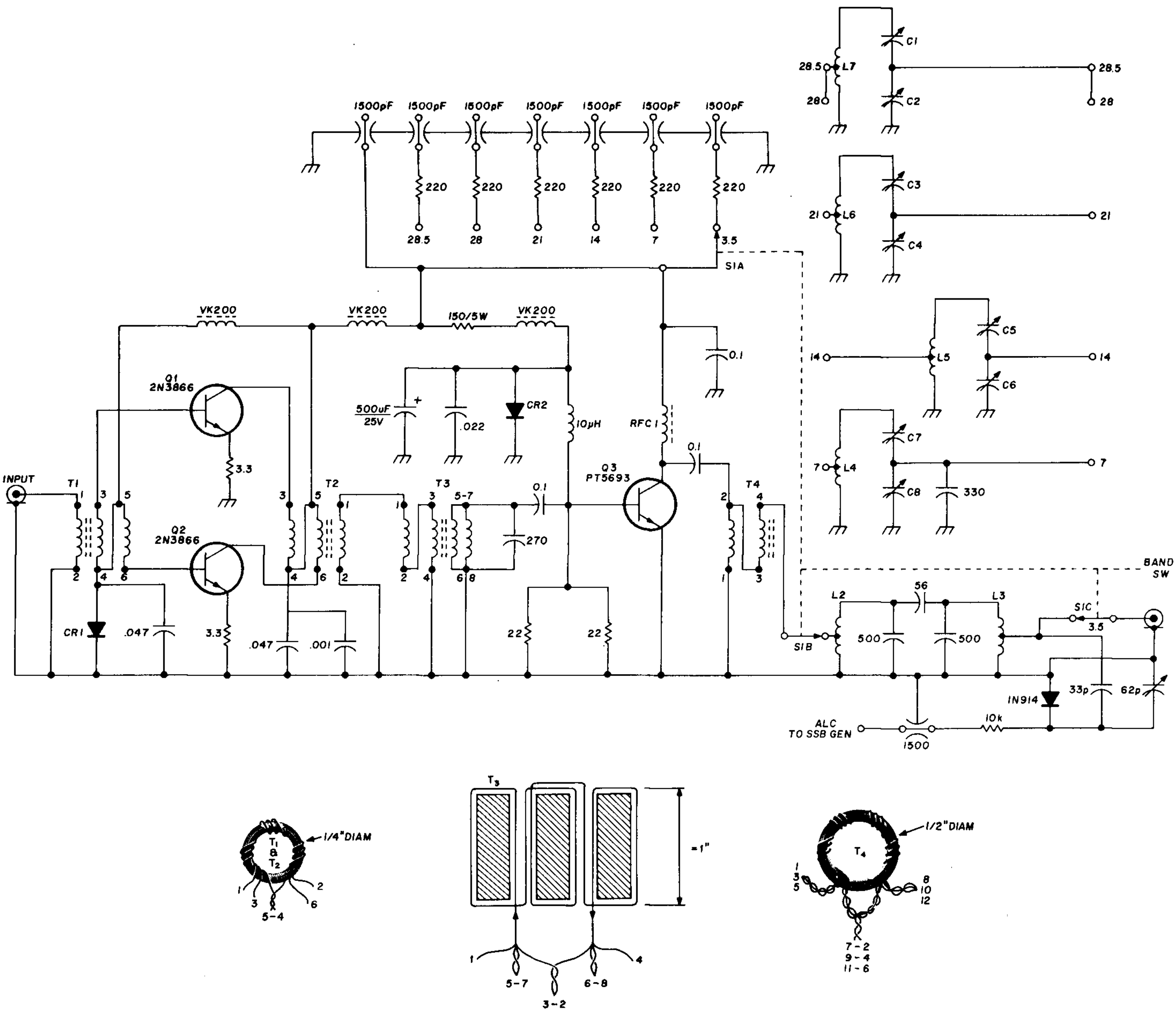


fig. 4. Schematic diagram of the power amplifier. RFC1 is 2 turns of no. 22 AWG (0.6mm) wound on a large Amidon bead. CR1 is a silicon diode rated at 0.4 A and 50 PIV. CR2 is also a silicon diode, 3 A and 50 PIV, stud mounted. T1 and T2 are each 6 turns no. 28 AWG (0.3mm), trifilar wound on a hi-μ 1/4-inch (6.5mm) core (Indiana General CF-2). T3 is a TV balun core wound with four twisted no. 28 AWG (0.3mm) wire. T4 is a high-μ Ferroxcube core wound with 5 turns of 6 twisted no. 28 AWG (0.3mm) wires.

	capacitors	inductors
80	—	L2 - 3.8 μH, 30 turns no. 24 AWG (0.5mm) tapped 10 turns from the ground end
40	C7 - 360 pF C8 - 600 pF	L4 - 2 μH, 22 turns no. 24 AWG (0.5mm) tapped 5 turns from the ground end
20	C5 - 180 pF C6 - 300 pF	L5 - 1 μH, 15 turns no. 20 AWG (0.8mm) tapped 4 turns from the ground end
15	C3 - 120 pF C4 - 200 pF	L6 - 0.6 μH, 13 turns no. 20 AWG (0.8mm) tapped 3 turns from the ground end
10	C1 - 90 pF C2 - 150 pF	L7 - 0.5 μH, 11 turns no. 20 AWG (0.8mm) tapped 3 turns from the ground end

Note that L2 through L5 are wound on Amidon T50-6 cores while L6 and L7 are wound on T50-10 cores.

The ARCO trimmers are type 42 or 46 with the values indicated being for resonance. The trimmers, Amidon, and Miller parts are available from Circuit Specialists, Box 3047, Scottsdale, Arizona.

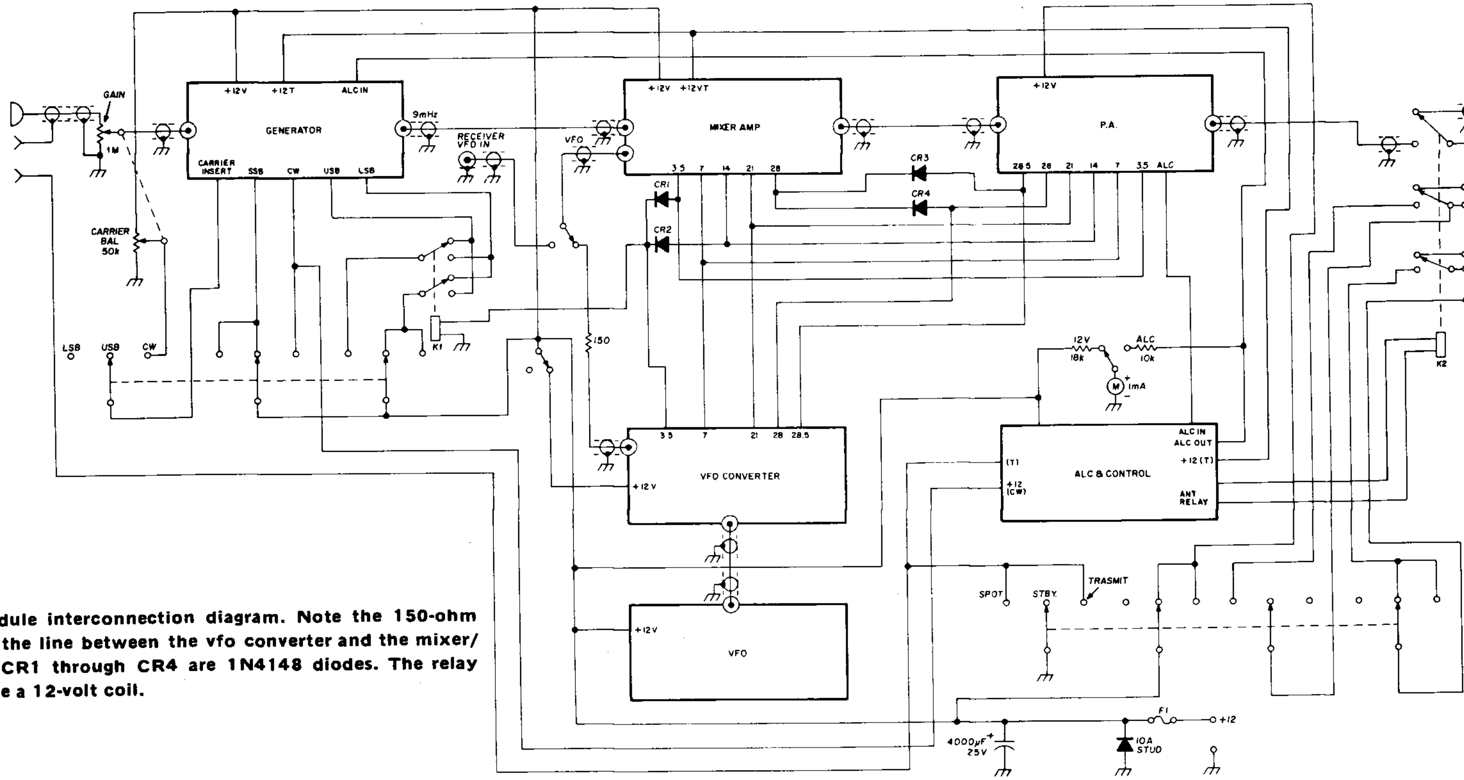


fig. 6. Module interconnection diagram. Note the 150-ohm resistor in the line between the vfo converter and the mixer/amplifier. CR1 through CR4 are 1N4148 diodes. The relay should have a 12-volt coil.

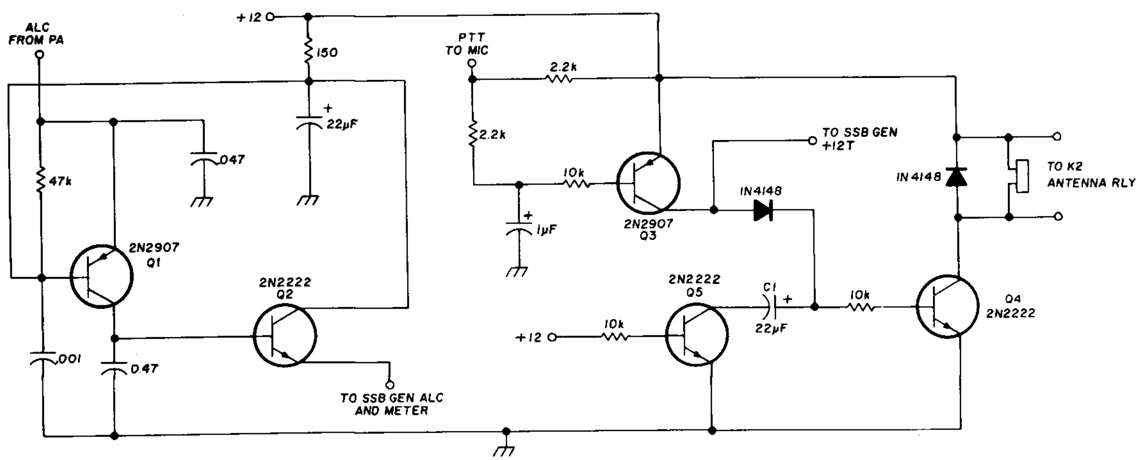
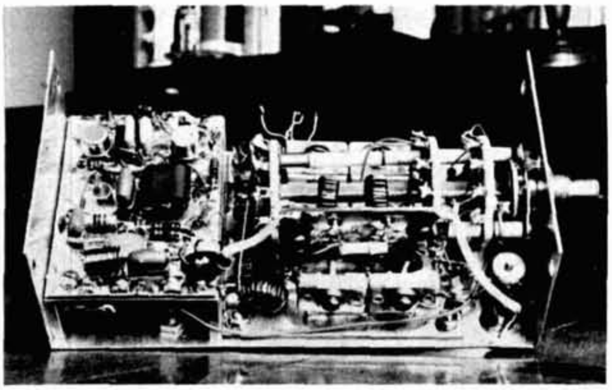


fig. 5. Schematic diagram of the alc and control module.

limits. T1 and T2 are wide-band toroid transformers, while T3 matches the collectors to the base of the final power transistor.

The final power transistor is a TRW PT5693, which was my pride and joy when I first acquired it. It is now a discontinued model but better substitutes are available: PT5649, PT8710, 2N6081 to name a few. The PT5693



Construction of the power amplifier module. The output transistor and associated circuitry is mounted on the heatsink at the left. The large switch selects the correct output network.

is rated at 17½ watts for 175-MHz fm. As a linear amplifier it will provide 10 watts. Forward bias, to improve linearity, is provided by CR2. When rf drive is applied, the necessary increased base current is "robbed" from the diode. CR2 also provides temperature compensation, and should be mounted on the heatsink close to the transistor.

auxiliary circuits

The alc and receive/transmit control circuits are shown in **fig. 5**. Q1, with the base and emitter at +12 volts, is normally off. The positive peaks of the rf are filtered and applied to the emitter of Q1. The negative peaks have been clipped by CR3. As the emitter becomes more positive, the transistor starts conducting. The gain of Q3 in the ssb generator is then controlled by this change in conduction level. Transmit control is by Q3, Q4, and Q5. Q3 and Q4 are normally off with K2 de-energized. Pushing the PTT button turns Q3 on, applying +12T power to the ssb generator, turning K2 on. To inhibit K2 opening up between characters on CW, capacitor C1 introduces a delay. In my case 22 μF worked fine; it may be varied for different dropout times.

checkout and alignment

I suggest you check each stage individually. As a minimum you'll need a vom or vtvm with an rf probe, a receiver that will tune around 9 MHz, and a grid-dip oscillator. An oscilloscope and frequency counter will help, and of course, a source of 12 volts at 3 amperes.

Starting with the ssb generator, apply power to Q4 and verify there is about 2 volts of rf on the secondary of T2; adjust the core for maximum. Switch to CW, add power to Q3 at +12T and connect an oscilloscope to the secondary of T1. It should indicate about 100 millivolts; again adjust the core for maximum voltage.

Connect the grid-dip oscillator through a 1 to 2 pF "gimmick" to the source of Q5, apply +12 volts and verify that the transistor oscillates with either crystal. With a receiver connected to the 9-MHz output jack, apply power to Q6 and Q7 and check for output. It may be necessary to adjust the turns slightly on L1/L2/L3 for maximum output.

Turn Q4 off, switch to the 9455-kHz crystal and tune a receiver to this frequency. Adjust R2, the 10k pot, for minimum signal. After turning Q4 on again, switch to ssb, and alternate the adjustment of R1 and C1 in the balanced modulator, for minimum signal. In some cases C1 must be connected to the opposite side of the balanced modulator for better suppression. In my transmitter the carrier was down 80 dB.

With the ssb generator connected to the mixer amplifier module, temporarily solder a 47-ohm resistor on the output of the mixer module. The output of the vfo converter, as measured at the 150-ohm interconnecting resistor, should be approximately 0.5 volts. Switch to CW, set the vfo to 3.5 MHz, and apply power to +12 and +12T. You should measure at least 1.6 volts with an rf probe on the output. It may be necessary to adjust the turn spacing of L1 and L2 for maximum output. Repeat this procedure for all five bands, making any slight adjustment necessary to the respective filters.

The power amplifier is normally adjusted for maximum output. A temporarily-connected rf voltmeter is used to measure the output across a 50-ohm dummy load. The respective trimmers, C1 through C8, are adjusted for each band. At this point the entire transmitter can be interconnected as shown in **fig. 6**.

Set the carrier balance pot to its midpoint and adjust R3 in the mixer amplifier module for 200 millivolts on the secondary of T5. Change to 28-MHz CW and put a voltmeter in the alc line from the PA module. As the carrier level is increased you should get 10 watts output on all bands. Note also that the alc responds. Now, connect the alc line to the ssb generator. Locate the 62-pF trimmer on the PA module and adjust it for proper level of alc action – it should limit the output to 10 watts. Switch to ssb and, using a two-tone test or your mike, verify that this condition also exists. Depending on your voice quality and language (Italian needs lots of alc) you may need to readjust the alc trimmer. The transmitter is now ready for operation.

reference

1. P. Moroni, I5TDJ, "Solid-State Communications Receiver," *ham radio*, October, 1975, page 32.

ham radio

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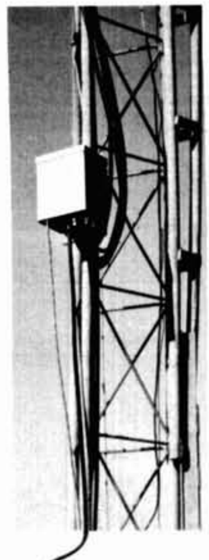

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the remote base:

an alternative to repeaters

Recommended reading for those wishing to relieve congestion on the vhf bands — a definitive description of the difference between remote-base and repeater stations

We feel that the case for remote base stations, as opposed to repeaters, is a very strong one for those interested in the advancement of vhf/uhf amateur communications. In this article we discuss the remote base-station concept with emphasis on its advantages over repeaters in today's crowded vhf/uhf spectrum.

Appreciable differences exist in the technical details between remote bases and repeaters. The former require a far more flexible command and control system than for repeaters, but they are potentially capable of performing many more functions. Furthermore, the remote base is designed and built with the systems approach in mind and with an eye toward modernization and expansion, whereas repeaters tend to be limited to one or two functions and are generally designed as "common-carrier" machines.

background

Radio amateurs have explored the characteristics of frequency-modulated communications systems since the 1930s, when Edwin H. Armstrong demonstrated the feasibility of this mode of transmission.[1] Initially failing to win acceptance on the hf bands because of the superiority of ssb in spectrum conservation and weak-signal

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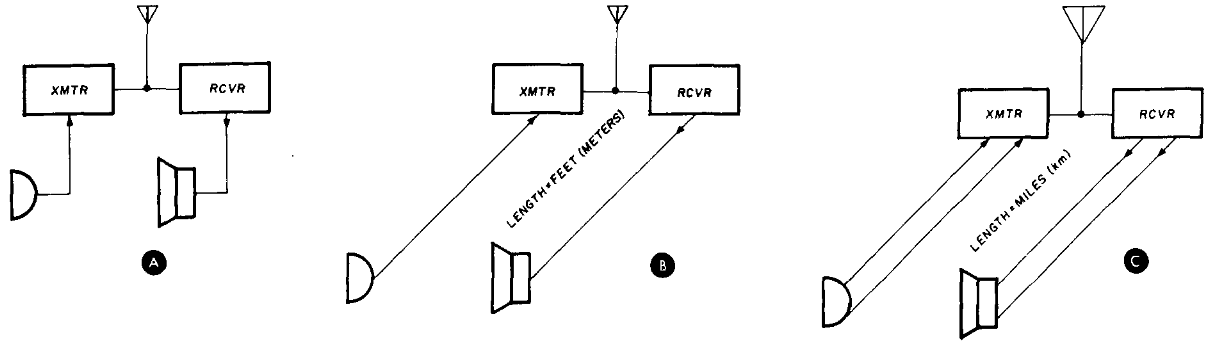


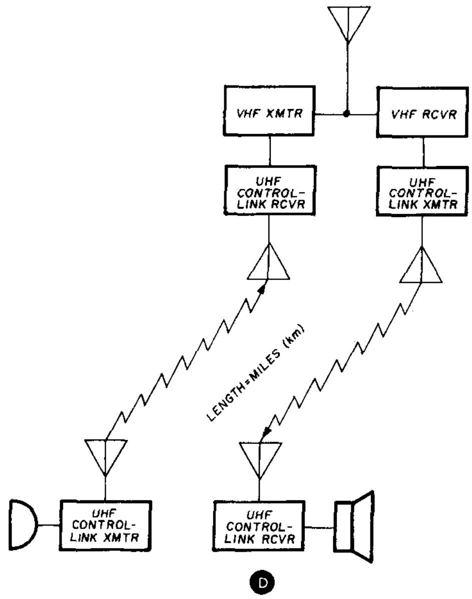


fig. 1. Evolution of a vhf remote-base station. A typical locally controlled amateur station is depicted in A. "Extended" local control is shown in B in which the microphone, speaker and control lines are routed from the operating position to equipment located elsewhere on the premises. A wire-line-controlled remote base is shown in C (transmitter and receiver are located at an elevated site to increase operating range). A radio-controlled remote base station, D, is the same as in C except that the wire link is replaced with a pair of uhf radio channels.

reception, fm entered into general amateur use on the vhf bands in the late 1950s.

Amateurs associated with the commercial* two-way radio business (land mobile service) purchased obsolete police and taxicab radios, retuned them to operate on adjacent amateur vhf bands, and began to experiment with the new mode. Having radios that generally offered one, or at most two, crystal-controlled transmitting and receiving frequencies (or channels), local fm groups quickly adopted standardized channels on which all radios would be operated. In the uncrowded vhf bands of those golden days, these few fm channels were placed well away from existing a-m and CW activity, and the new fm operators were generally ignored. With pretuned radios transmitting and receiving on the same frequency and with effective squelch circuits silencing receiver noise between transmissions, a natural party-line type of operation ensued. Thus the very first amateur fm operations were of a simplex nature – direct, point-to-point transmissions on a single frequency.

In southern California, the first simplex channels were established on 146.760 and 146.940 MHz. Since an a-m repeater (K6MYK) had been in operation at this time, fm operators saw no need for duplication. Instead they concentrated on extending the range of their simplex stations. In the mid 1960s several groups of fm operators established remotely controlled 2-meter fm transmitters on several southern California mountains. These transmitters were operated by radio-control links on the 450-MHz amateur band. Soon thereafter 2-meter fm receivers, tuned to the transmitting frequency, were added to the remotely controlled installations. These early groups of fm experimenters had established base stations (*i.e.*, stations designed to be operated at fixed locations), which were on mountains to increase range. They were remotely controlled and were operated by uhf radio links. These were among the first remotely controlled base stations, or remote bases, as they are more commonly known.

*The term "commercial radio" applies to a radio originally designed and manufactured for operation in the commercial two-way Land Mobile Radio Service and adapted to amateur use.

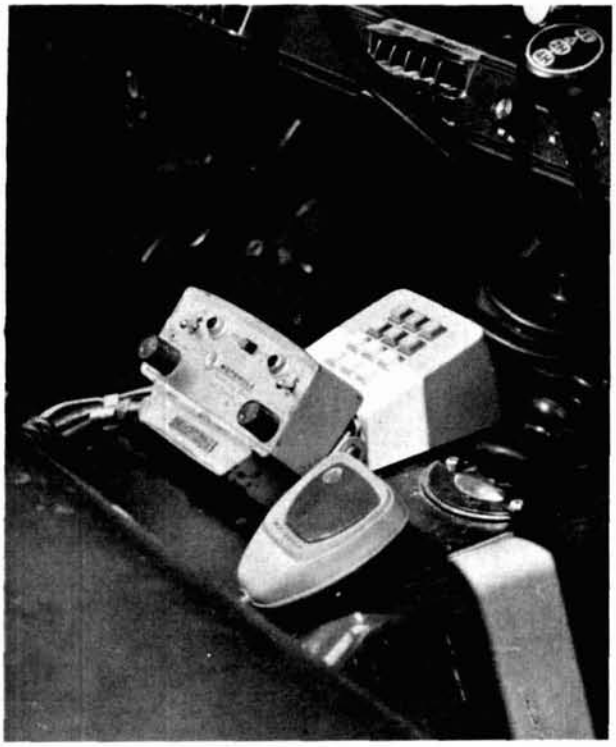
Early remote bases in southern California included those of W6YY, WB6SLR, WB6CZW, WB6LXD, and WB6QEN. From this beginning the number of southern California remote bases has increased to over 100 at present, with a smaller number in northern California, Nevada, and Arizona. Remote bases have been established in other parts of the country though nowhere in the numbers found in California.

the remote-base concept

While both remote bases and vhf fm repeaters operate from elevated locations, it should be clearly understood that a remote base is *not* a repeater station. Major differences exist between them in construction, operation, and licensing; these differences are discussed later in greater detail. Most important, however, is the difference in intent of the two stations. Repeaters exist primarily to extend the intracommunity range of user mobile and hand-held portable stations, most operators of which are *not* owners or control operators of the repeater. Remote bases, on the other hand, are extensions of the personal stations of their owners and are operated generally only by control operators.

Fig. 1 presents the evolution of the remote base concept. A typical amateur station is depicted in **fig. 1A**; for the sake of discussion let's assume that it's an fm base station. The owner/control operator talks on the local microphone, listens on the local speaker, and manually turns the transmitter on and off. All controls are at arms' length. This has been the typical style of amateur operation on all bands since the inception of ham radio.

Let's now assume, for reasons of space limitations, that it is inconvenient for the amateur to keep his fm base equipment at his operating position. Since fm stations are operated on crystal-controlled, fixed-tuned channels, it's not necessary to have direct physical access to the transmitter and receiver for tuning purposes. Therefore the amateur may elect to place his base equipment in his basement, attic, or garage, and extend the microphone, speaker, and push-to-talk lines back into his operating position (**fig. 1B**). Many commercial fm base stations include provisions for doing just this. The amateur is now operating his base station remotely by wire line, although for licensing purposes the station is still under direct control as long as it is entirely contained within the amateur's fixed station license location.



Several radios may be installed at the remote base and operated through one uhf control station, which avoids duplication of radios between home and car or between several cooperating amateurs.

In **fig. 1C** we extend the operating range of the fm base station by relocating it to a higher elevation. It might be situated at a friend's house on a hill, on a mountain top commercial two-way radio site, or on top of a tall building – all depending on the local geography. The station is still controlled and operated by wire line, but in this case the length of the control line is measured in miles (km) rather than in feet (m). (The technical details of the control system depend on the characteristics of the wire-line pair, its length, and whether or not it is leased from the telephone company). This installation is now a remotely-controlled base station, or remote base, and it must be licensed as a remotely-controlled station. Few, if any, southern California remote bases are wire-line controlled, but the idea has merit for other areas of the country where distances and topography permit.

Now, let's assume that no wire lines can be run to the proposed remote base-station location because of expense, distance, or inaccessibility. It then becomes necessary to control and operate the remote base by radio (**fig. 1D**). FCC rules, (*Part 97.109a*), require that radio remote-control links operate on frequencies above 220 MHz. While some remote bases operate with 220-MHz radio links and a few others use the amateur microwave bands, the vast majority of remote base operators have elected to control and operate their stations through radio links on the 420-450 MHz amateur band. The reason for this is the availability of high-quality, surplus, commercial fm equipment designed to operate in the 450-470 MHz land mobile service band, or the 406-420 MHz government service band. The former set of radios can be easily retuned to operate in the 440-450-MHz segment of the amateur 3/4-meter band, while the latter set converts easily to the 420-430-MHz segment.

Note from **fig. 1D** that the control link must be bidirectional. Speech and control information is sent from the local uhf control-link transmitter to the remote base uhf control-link receiver. The information is demodulated and used to operate the vhf fm transmitter. Signals received by the remote base vhf receiver are used to modulate the remote uhf control-link transmitter and are then recovered by the local control-link uhf receiver. The entire control link could be operated on a single uhf channel but this is technically cumbersome. It has become customary to use separate channels for the uhf uplink and downlink. Spacing between the two control-link channels is typically on the order of 5 MHz, a separation sufficient to allow *all* uhf receivers to function properly while their associated transmitters are operating. Thus full two-way duplex operation of the control link is permitted; the control operator can simultaneously transmit signals to, and receive signals from, the remote-base station.

The locally-controlled fm base station in **fig. 1A** has now grown to become the radio-controlled remote base station in **fig. 1D**. Fundamentally, however, the only significant change between the two stations has been the replacement of three pairs of wires by one pair of 450-MHz radio links: the pair connecting the microphone to the transmitter, the pair between the receiver and its speaker, and the push-to-talk line pair.

remote-base advantages

To this point we've discussed the concept of the radio-controlled remote-base station operating on fm simplex channels. While many southern California remote bases have been established to do just this, the description above is actually a restricted view of the capabilities of remote bases. In point of fact, the existence of the basic radio link and control equipment, together with the physical location of the remotely controlled station, represent a resource that can be developed: radios of any type of emission on any amateur band, from 1800 kHz to 10 GHz or higher can be operated remotely. The remote base, for example, allows operation of high-power transmitters, such as on the 50-MHz band, in areas where TVI is a problem. It allows operation on any amateur band where antennas can't be erected at the control operator's location. It affords improved operation on the hf bands where space for efficient antenna systems may be more easily available at the remote-base site.

A remote base offers the opportunity for a group of amateurs to relocate all their radios at one central point while achieving antenna space advantages on hf and height advantages on vhf/uhf. This relocation includes not only home-station radios but mobiles as well. All may be replaced with one uhf radio per location, thereby saving on duplication of radios among several home station and mobile installations.

Those remote-base stations that operate on the fm simplex channels promote spectrum conservation in several ways. With their extended local operating range, they provide interference-free regional-area communications. This can relieve congestion on the hf phone bands by shifting local-area communications to vhf. Because remote bases operate as simplex stations, each occupies only one vhf channel at a time (*i.e.*, 146.940 MHz) rather than two required by a repeater (*i.e.*, 146.340 and 146.940 MHz). Additionally, by the nature of the remote-base design, a control operator always monitors the channel of operation with a mountaintop receiver before transmitting. Thus activity on the operating channel over the entire remote base transmitting range can be easily detected and inadvertent interference avoided. The same is true for repeaters only when a separate receiver and auxiliary link system is used to monitor the output frequency from the repeater site.

Finally, a remote base usually represents the desire of a group of active vhf/uhf amateurs to build a communications system. In deciding to build a remote base, the constructing group does not require the use of the limited set of 2-meter repeater channels. This translates to spectrum conservation. In the southern California area it would be impossible to fit more than the one-hundred remote-base groups into individual 2-meter repeater pairs, even when using 15-kHz channel spacing and all the simplex channels. While it's true that each remote base requires a pair of dedicated channels, these channels are in the spacious 440-450 MHz region. On a narrow-band deviation (±5 kHz) basis, this region of the spectrum contains a potential 200 pairs of channels, with another 200 pairs in reserve between 420 and 430 MHz.

constructing a remote base

Occasionally an individual will undertake the entire job of designing, building, and installing a remote base. He will then either operate it as his own station, or may invite his friends to use the remote base as co-control operators. More often, in southern California, at least, a group of individuals will be formed to build and operate the remote base, thereby sharing the financial and technical responsibilities. The following comments, although addressed primarily to the group-ownership case, apply as well to single-owner bases.

administrative and technical responsibility. A remote base is a communications system that contains separate but intercommunicating radios. The cost and effort to build and operate a remote base is greater than that required to operate a home station, so careful attention should be given to financial and technical responsibilities. One member of the group should be responsible for handling and reporting finances. Provisions should be made for one owner selling his equity in the remote base in the event he must move out of the area. Provisions also should be made for including new members or owners. Lack of adequate preparation in this area has been an historical source of conflict in many remote base groups.

One individual should be responsible for obtaining the site for the remote base, which should be the first task undertaken and completed. When the site involves rental of space at a commercial two-way radio installation, it has been found best to have a single individual from the group maintain relations with the site owner. One individual will have to arrange for licensing the remote base, whether it is in his name or in that of a club station. Additional non-technical duties that may need to be delegated include a) obtaining supplementary permits (for example, from the Forest Service, Bureau of Land Management, local governmental authority) to operate the station, b) maintaining memberships in regional amateur radio associations, and c) providing for fulfillment of public-service commitments.

Technical responsibilities in establishing a remote base should be divided into design, construction, and installation and maintenance areas. A single individual should have overall responsibility for the design of the entire system, although he may wish to delegate specific design projects to others. Particular attention should be given to interfacing between the various subsystems, such as audio and control signal levels between rf hardware and the control system.

Once original equipment designs are complete, construction of individual components can be delegated to group members. Emphasis should be placed on building for reliability, both in selection of components and in construction practices. One or two individuals should assume the responsibility of tuning the rf hardware, integrating the amateur-constructed subsystems into the final assembly, and performing on-the-ground checkout.

Maintenance. When installation of the remote base is completed, the maintenance team assumes responsibility for continued operation. These people should be equipped with the specialized test equipment (wattmeters, signal generators, frequency and deviation meters) for servicing fm communications systems. Inevitably there will be an initial period of system debugging as various design and subsystem deficiencies become apparent. Frequent trips to repair and service the remote will later taper off to occasional visits for scheduled maintenance. At this point the design team will probably begin work on improved subsystems to be retrofitted into the existing remote base, or perhaps better quality rf hardware will be acquired and put into service. Few remote bases are ever truly "completed."

Rf hardware. The remote base typically will consist of commercially manufactured rf hardware and amateur-built control systems. Antennas may be either commercially manufactured or home built. In the selection of transmitter and receiver strips, southern California remote-base groups invariably use late-model commercial equipment. All or partially solid-state equipment is preferred for greater reliability, although high-quality all-tube equipment has performed well at some installations for many years. Receivers should have good sensitivity (fet preamps may be added) as well as excellent rf selectivity and cross-modulation rejection; many busy commercial radio sites contain very heavy rf fields. Vhf receivers and transmitters should be capable

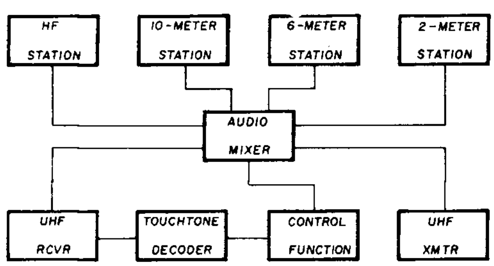


fig. 2. Block diagram of a typical remote-base station. Control signals from the uhf receiver are decoded in the Touch-Tone decoder, processed in the control-function circuits, and used to operate hf or vhf base station. Speech information is routed to the selected base station through the audio mixer.

of operation on several different channels, so that the remote base may be switched to operate on whatever channel the control operator wishes to use. The vhf transmitters should be capable of moderate power output (30 - 100 watts), and should be free from spurious output. A remote base operating from an elevated location with a few hundred milliwatts of spurious output will certainly make its presence known.

Commercially manufactured resonant cavities are often used ahead of the entire vhf portion of the remote base to provide additional rf selectivity. The uhf remote base control-link radio should be the best that can be purchased, since it will be the limiting factor in using remote base from distant locations. Matching commercially manufactured 110-Vac power supplies for fm installations are preferred to home-built supplies since they provide the exact voltages required, have provisions for properly interconnecting the transmitter and receiver to other equipment, and are usually rated for continuous-duty operation under severe environmental conditions.

Antennas. Antennas and transmission lines for the remote base should be selected with regard to survival under severe weather conditions. Antenna gain, easily obtainable at vhf and uhf is an additional factor to be considered. Remember, however, that many "gain" antennae have major radiation lobes directed at the horizon; for a mountaintop installation it may be preferable to select antennas that radiate their major lobes below the horizon. Transmission lines should exhibit the lowest loss possible; weak received signals and expensively generated vhf and uhf power can be lost in inferior coax. If available, commercially manufactured Foamflex should be used.

Control systems. Control systems are the heart of a remote base; they are always amateur constructed. In southern California they vary in complexity from simple audio-tone decoders that drive rotary stepping switches to sophisticated multilevel digital logic circuitry. These advanced systems allow any piece of rf hardware in the remote base to be interconnected with one or more of the remaining pieces in various combinations. Control systems reflect individual needs and capabilities; space prohibits giving specific examples.

A control system performs several functions in addition to enabling transmitters to be turned on and off. In general, the control system must provide for:

1. Authentication and decoding of the received control signals.

2. Selection and activation of the required transmitters and receivers.

3. Selection of specific frequencies to be used within each transmitter and receiver.

4. Processing and conditioning of audio.

5. Automatic indentification of active transmitters.

6. Automatic timing of transmission length to provide ultimate shutdown protection should the control link fail.

Typically, remote bases are controlled by specific audio tones sent along with speech on the uhf uplink channel. The use of *Touch-Tone** audio encoders for this purpose has become relatively standard. The control link usually also contains a subaudible continuous tone squelch signal (*Private Line, Channel Guard*, etc.) as a verification device. Audio-tone decoders, logic circuits, and audio processors are matters of personal preference and design, although some circuits have been published. Timers and IDers are well documented in amateur literature.

**Touch-Tone* is a trademark of American Telephone and Telegraph.

packaging. It is considered good construction practice to build all control circuits, timers, audio processors, and identifiers on standard-size edge-connector cards for insertion into a card rack. Interconnections to the individual pieces of rf hardware from the control system are made from the contacts at the rear of the card rack. Provisions should be included in any control system for expansion; the use of individual cards for specific circuits facilitates this goal.

New designs for amateur-built components should be breadboarded and thoroughly tested on the bench before being constructed in final form. In testing, provisions should be made for checkout of the new designs under conditions of continuous duty in temperature and humidity extremes. **Fig. 2** shows a typical remote base station.

One other design feature should be included in any remote base: "series audio." This is illustrated in **fig. 3**. In a series-audio system, the remote-base vhf receiver runs continuously, even when the control operator transmits; his speech is sent from his 450-MHz control transmitter to the 450-MHz remote base receiver and is then transmitted by the remote base vhf transmitter. The vhf receiver remains on, and although not connected to an antenna during this time, still receives a signal from the vhf transmitter operating nearby. This signal is retransmitted back to the control operator over the 450-MHz downlink. The control operator can listen to his voice as it is being transmitted on vhf by the remote base and can verify that the vhf transmitter in the remote base is being properly modulated. The speech from the control operator follows a path from the control-station microphone back to the control-station loudspeaker, with the remote base vhf transmitter and receiver in "series" with the duplex control link.

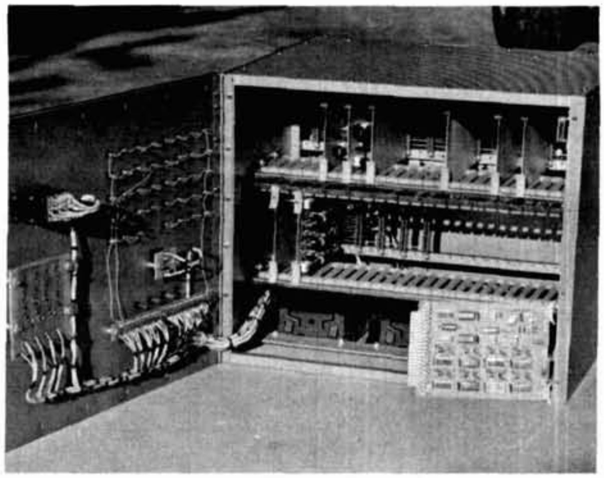
operating a remote base

What can be done with a remote base is limited only by the imagination and ingenuity of its owners. First and foremost, however, southern California remote bases operate on the area's simplex channels: many can be heard on 146.940 and 146.760 MHz. This is the historical rationale for the establishment of a remote base; and in fulfilling this function, remote bases have helped to remind fm operators – in a time of rapidly expanding numbers of repeaters – that much good work can be accomplished on a point-to-point simplex basis. Occasionally a remote base will be used to transmit bulletins of interest to the regional fm community on 146.940 MHz (a channel that every fm operator can monitor). With their height advantage, many southern California remote bases can be heard from Santa Barbara to the Mexican border; they provide an invaluable resource for tying together an entire region by radio.

Because of the large number of remote bases in southern California, an agreement has been reached that use of the 146.460-MHz simplex channel will be limited to an "intercom" channel among remote bases. This allows two or more remote bases to avoid monopolizing 146.760 or 146.940 MHz, which would prevent mobiles and home-base stations from using these channels. This arrangement has worked well in practice. A number of remote bases also have provisions for operating on 52.525 MHz and 29.600 MHz, the National simplex frequencies for these bands.

While it's possible to equip a remote base to transmit on a repeater input channel and listen to the corresponding repeater output channel, such practice is not often done. An exception would be where the repeater to be contacted is so far from a majority of the remote bases' control operators that they couldn't transmit on vhf directly into the repeater from their locations.

Several remote bases have been equipped with hf ssb transceivers. Notable was the former WA6ZRB remote base, which contained provisions for transmitting on 40 meters including remote tuning of the transceiver vfo. The remote was often used by control operators to check into the WCARS net.



Control system for the remote-base station of WA6ZOI. Card-rack construction is featured. A typical edge-connector card, containing one circuit element, is shown in the lower-right corner.

Many remote bases contain autopatches. The use of a remote base for this purpose is particularly fortuitous because it removes the autopatch operation from repeaters in the busy 2-meter band thus reducing congestion and increasing repeater availability for mobile users. Generally, because of nonavailability of telephone lines at the remote base site, a special pair of auxiliary link channels operating in the 420 - 430 MHz region are used to transmit control-link audio from the remote base site to a telephone ground station at a convenient location.

Many remote bases contain auxiliary uhf radios, which link to other remote bases in other areas. Often two or more remote base groups will enter into reciprocal operating agreements, so that by means of the radio links the members of one group, transmitting through their own remote, can control and operate the other remote bases. For example, the Gronk Radio Network can be activated so that stations in southern California can talk to and operate through remote bases in central and nothern California, and in Nevada and Arizona (and vice versa). This is an area where advanced fm operators are awaiting FCC rules and regulations to catch up to the state of the art.

Several remote bases contain special functions, such as telemetry of prevailing environmental and equipment conditions at the remote base site, or television surveillance of the site.

Remote bases have participated in emergency activities. With their great range and ability to contact virtually any fm-equipped amateur through vhf simplex channels, they provide a natural focus for emergency and disaster operations. Of particular note is the participation of remote bases in the rescue effort after the San Fernando Valley earthquake of 1971.[2] The use of remote bases to relay traffic accidents and other emergencies to public service agencies is a common occurence.

licensing

Before adoption of Repeater Docket RM 18803, remote bases were routinely licensed by the FCC after the required showings had been submitted. The FCC, then as now, wanted to be convinced that the remotely controlled station would not be tampered with or operated by unauthorized people, and that provisions had been made for automatic shutdown of the transmitters should a failure of the control link occur. Remote-base licenses could be single "additional station" licenses, or primary-station licenses with authorization for remote control. Control operators required no special licenses but were listed as control points on the remote base license.

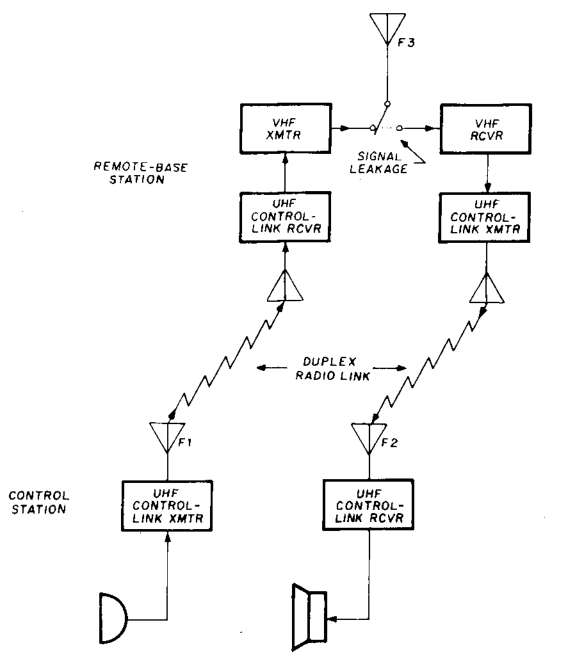


fig. 3. Remote-base station except for series-audio feature. All transmitters and receivers operate simultaneously. Speech information travels from the control-operator's microphone through the remote base station, then back to the control-station speaker.

Southern California remote-base operators became concerned with the status of licensing after the adoption of RM 18803. Apparently under the misapprehension that only a handful of remote base licenses would be requested, the FCC devoted its time to the increasing number of repeater applications. But along the way, they released a set of "interpretations" of the new Part 97 rules, which completely changed the nature of remote-base operation.

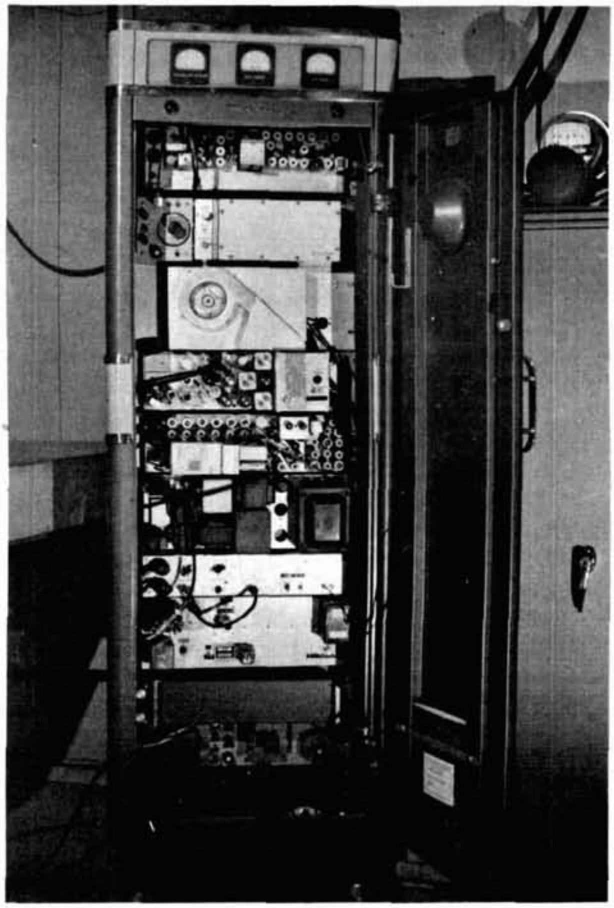
The interpretations included a requirement for a) the licensing as auxiliary-link stations of all uhf transmitters that carry speech to and from the remote base, b) the licensing as control stations of all uhf transmitters sending control information to the remote base, and c) the use of separate uhf frequencies for remote base speech and control uplink channels. A subsidiary effect of these interpretations was to declare as "illegal" the operation of the remote base from portable and mobile locations since, by definition, auxiliary links must operate between two fixed points. The FCC has since dropped the requirement that separate channels must be used for speech and control uplinks.

Nevertheless, remote-base operators are faced with a cumbersome and expensive licensing procedure and with operating restrictions more severe than those before RM 18803. The current licensing procedure, under which the FCC is processing and issuing remote base licenses, is as follows:

The mountaintop remote base must be licensed as a "secondary station" or "club station." This basic license covers the hf and vhf portion of the station; an "auxiliary link" license is required to cover the uhf down-link transmitter. Both privileges may be combined on a single station license for a single application fee. Each control operator must modify his primary station license to include both "control station" and "auxiliary link" privileges; this also can be accomplished with one application fee. The FCC has deleted the requirements for submitting many parts of the required showings, making them instead a required part of the station log.

During the ensuing years the FCC has come to better understand the remote base concept, and has shown increasing willingness to allow remote base (and also repeater) licensees more latitude in the operation of their stations. Docket 21033, which is based in part on a Rule Making petition by the authors of this article, if adopted, will grant essentially complete freedom to operate remote-base stations in the traditional ways described above. For example, Commission restrictions against operation of a remote base from portable and mobile control locations will be eliminated, licensing will be greatly simplified, and the distinction between the remotely controlled base station (with its associated control operators) and the true repeater will be clearly drawn. Southern California remote base operators are generally pleased with the contents of this Docket, and are looking forward to increased flexibility and freedom to innovate.

Remote bases are completely different in intent and operation from repeater stations. Repeaters are operated to extend the communications range for operators of specific mobile and hand-held portable stations interested in communications among themselves. Remote bases are operated as *extensions* of the owners' personal



View of the WA6ZOI remote-base station. Equipment includes a 50-MHz base station, 450-MHz receiver, power supplies and control system. The station is located on Johnstone Peak in ***southern California.***

stations for purposes of communicating with *all* amateur stations. Almost all users of repeaters are *not* control operators, and the act of activating a repeater by transmitting on its input channel is *not* an act of controlling the repeater. Repeater control station operators are responsible for activating the station to repeat the transmissions of other amateurs and for suspending operations in the event that FCC rules are not complied with. By contrast, in southern California, every user of a remote base station has been a control operator. The remote base must be commanded by the control operator through the uhf radio link to operate for each transmission; it is not designed to *automatically* retransmit signals.

The comment has been made that, because the operation of a remote-base station involves speech transferred between hf-vhf and uhf frequencies, the remote base operates as a crossband repeater. From the discussion above it should be clear that the remote-base station does not fit the basic definitions of a repeater. The act of monitoring a vhf channel through the remote-base vhf receiver and uhf downlink channel is not an act of repeater usage. The system *could* be used as a crossband repeater if a) two nonremote base simplex stations were to transmit on a channel being monitored by a remote base, and b) each were to listen to the other through the remote base 450-MHz downlink instead of directly on the vhf channel. In practice this seldom happens; if it should happen it is the responsibility of the remote-base control operator to suspend operations on that vhf channel.

It is our feeling, which is shared by a majority of southern California remote base operators, that liberalization of the present FCC rules (and interpretations of these rules) is required. Ideally each remote base could be licensed as a remotely controlled station, with one license covering the entire mountaintop station including the uhf radio links. Each user would be required to be an authorized control-station operator, having control-station privileges added to his primary station license. There would be no limit to the number of control stations that could be *conveniently* licensed including other remote base stations operating as control stations (many remote bases have 15 control operators at present). The control-station license would confer the privileges of both controlling and operating the remote base and would be usable in portable and mobile operation in addition to its customary fixed-station use. Were these proposed changes to be adopted, remote-base operators would achieve more flexibility to innovate in the amateur vhf and uhf bands.

conclusion

For those vhf/uhf-oriented groups wishing to expand their interests from individual circuits and individual stations to building an entire communications system, the remote-base concept has several advantages. It offers the chance to experiment with systems engineering – to design and build a system constructed from individual pieces of equipment. The final system can reflect the designers' needs, wishes, and abilities, rather than the standardized requirements of the marketplace. The remote base offers reliable, interference-free local communications capability on the vhf bands, thus helping to relieve congestion on the crowded hf bands. It abolishes the need for duplication of radios between home and car, or duplication among several cooperating owners, and permits the establishment of high-power transmitters and large antennas at the remote-base site. It fosters spectrum conservation on popular bands by removing the requirement for dedicated repeater channels, substituting instead the need for a dedicated pair of channels in the far-less congested 420-450 MHz band. It promotes the use of simplex communications, thus reducing the load of busy repeaters.

Southern California amateurs developed the remote-base concept more than ten years ago. It has proved to be a useful adjunct to the amateur vhf community. We look forward to its adoption in other parts of the country.

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1. Thomas McMullen, W1SL, *FM and Repeaters For the Radio Amateur*, ARRL, Newington, Connecticut, 1972, page 7.
2. "The Southern California Earthquake," *QST*, June, 1971, page 60.


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graphical aid for winding rf coils

You'll find this family of parametric curves indispensable when designing small inductors for rf work

Winding a small inductor can be frustrating and usually involves several trials. Here's a simple graphical method that will produce accurate inductance values on the very first try. Carbon composition resistors, ¼ through 2 watts, and standard-size coil forms up to ½ inch (12.5mm) diameter are used for winding the coils.

literature methods

The usual method for winding coils is to use Wheeler's approximate formula

$$L = \frac{r^2 n^2}{9r + 10l} \qquad (1)$$

where L = inductance (μH)
r = coil radius (inches)
l = coil length (inches)

If all dimensions are in millimeters, **eq. 1** becomes

$$L(\mu H) = \frac{0.0394 r^2 n^2}{9r + 10l}$$

These formulas are accurate with one percent for $l > 0.8r$ (*i.e.*, if the coil is not too short).[1] The procedure for winding the coil is described in *The Radio Amateur's Handbook* and several other publications. It usually involves the solution of Wheeler's formula for n, searching a wire table for a suitable wire size, then spacing the wire along the coil form to get the required number of turns in the calculated coil length. Also coil inductance slide rules are used such as the ARRL *Type A Lightning*

By Earl Palmer, W7POG, 17510 Military Road South, Seattle, Washington 98188

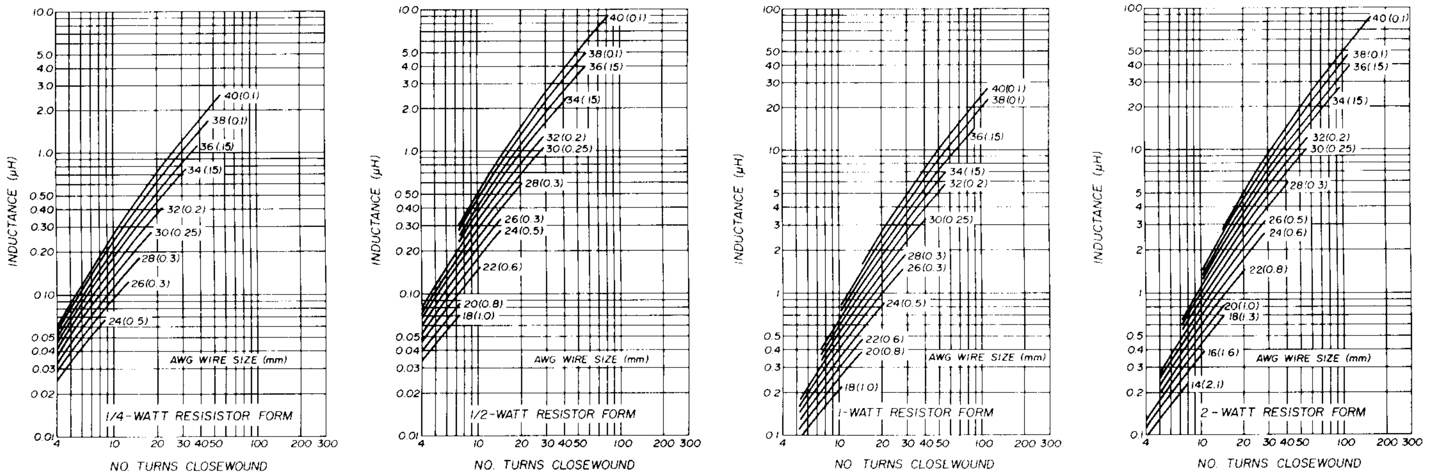

fig. 1. Curves for finding inductance as a function of turns closewound on composition resistor forms.

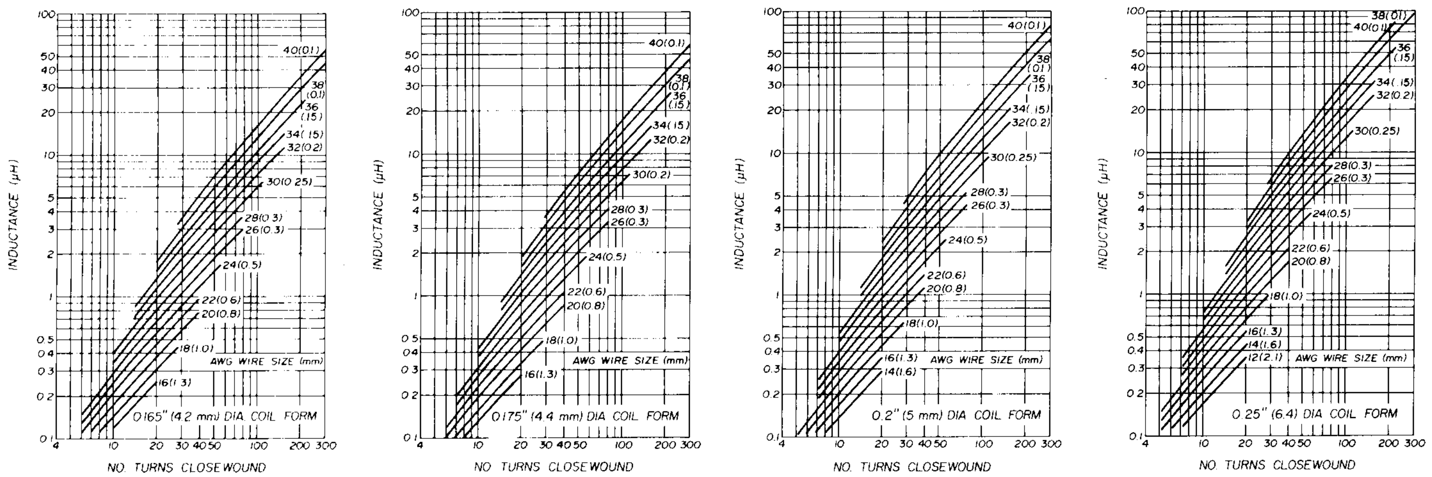

fig. 2. Curves for finding inductance as a function of turns closewound on standard-size coil forms 0.165 in. through 0.25 in. diameter (4.2 through 6.4mm).

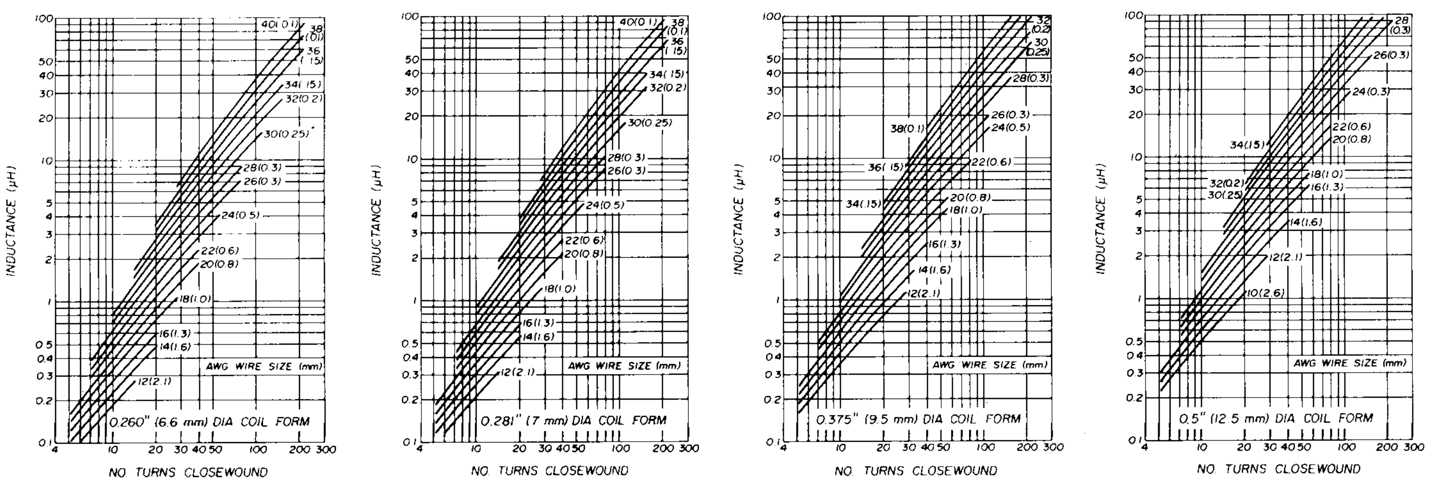

fig. 3. Curves for finding inductance as a function of turns closewound on standard-size coil forms 0.260 in. through 0.5 in. diameter (6.6 through 12.5mm). A complete set of full-size curves is available from the author for $2.00 postpaid.

Calculator. These work fine for large coil diameters, ½-inch (12.5mm) or greater, but are not calibrated for smaller coil forms.

Accurate small coils can be made if the windings are closewound. Then the wire size determines the coil length (number of turns x wire diameter = coil length). For closewound coils, Wheeler's formula can be written:

$$L = \frac{d^2 n^2}{18d + 40n/T} \qquad (2)$$

where L = inductance (μH)
d = coil diameter (inches)
n = number of turns closewound on coil
T= number turns/inch of the particular wire size used for the coil

For dimensions in millimeters, **eq. 2** is

$$L\ (\mu H) = \frac{0.0394\ d^2 n^2}{18d + 40n/T}$$

This formula is no simpler than the previous formula, but it is in a form that can be plotted in terms of inductance vs turns for a given wire size and coil diameter. This was done using a Monroe 1666 desktop calculator and plotter. With these graphs and a fair assortment of enameled copper wire, accurate inductors up to approximately 100 μH can be wound.

graphical solution

Fig. 1 is used for winding inductors on carbon composition resistors. Composition resistors make excellent forms as their size is standard throughout the industry. The graphs will not let you try to wind more turns on the resistor than it will hold, and the minimum number of turns will be high enough to ensure reasonable accuracy. In winding these inductors keep the following in mind:

1. If the resistor is not going to be removed from the coil use a high resistance value, 100k or more.

2. If the Q of the inductor is important or the frequency of resonance is above 30 MHz, it would be best to remove the resistor form. For very small wire sizes this is impractical as the resultant coil would be too fragile.

3. If the accuracy of the inductor is important, then use as many turns as possible. The greater the number of turns, the more accurate the formula and the graph.

4. When an inductor is to be used as an rf choke, its self-resonant frequency is important, as it exhibits a high impedance at that frequency. Commercially manufactured inductors usually have specified self-resonant frequencies and may be used in a circuit for that reason. So beware — a hand-wound inductor may not work in a particular circuit even though it has the same inductance as a manufactured inductor. To find the self-resonant frequency of an inductor, short its leads and measure the resonant frequency with a grid-dip meter.

Figs. 2 and **3** are used for winding coils on standard-size forms that are normally used for printed circuit work. Larger coil-form sizes of 1/2, 3/8, and 1/4 inch (12.5, 9.5, and 6.5mm) are also included. The same precautions concerning Q and accuracy apply. For these inductors keep the following in mind:

1. If the coil is to be tuned with a ferrite or powdered-iron slug, the frequency used to determine the value of required inductance should be 15 to 20 per cent above the desired value. This is only a rule of thumb and does not take into account the difference in permeability of different slug materials. For brass slugs use a frequency 10 to 15 per cent lower.

2. For slug-tuned coils, the length of the windings should be less than the length of the slug. Typically it should be 75 per cent or less.

3. For air-wound coils, use a frequency 10 per cent below that desired. The coil can then be spread slightly to obtain the desired frequency.

using the graphs

To gain confidence in these graphs, wind several inductors and check their actual value. A simple way to do this is to make a tuned circuit with a known capacitance then check the resonant frequency with a grid-dip meter. The expression for inductance is:

$$L = \frac{25330}{f^2 C} \qquad (3)$$

where L = inductance (μH)
f = frequency (MHz)
C = capacitance (pF)

I have wound at least one inductor from each graph and find the accuracy better than 10 per cent, which is satisfactory for most home-construction projects.

Example: Suppose a 3.3-μH inductor is needed for a low-power lowpass circuit and it is desired to keep the size at a minimum. Looking at the graph using a ¼-watt resistor as a form, the maximum inductance is approximately 2.5 μH, so the next size (or ½ watt) must be used. On the ½-watt resistor graph, the 3.3-μH inductance line intersects the wire-size lines at 36 turns AWG no. 40, 41 turns AWG no. 38, and 48 turns AWG no. 36 (0.08, 0.10, and 0.13mm). Any of these combinations can be used; however, it would probably be somewhat easier to work with the larger wire size.

For calculating the inductance of toroids see references 2 and 3.

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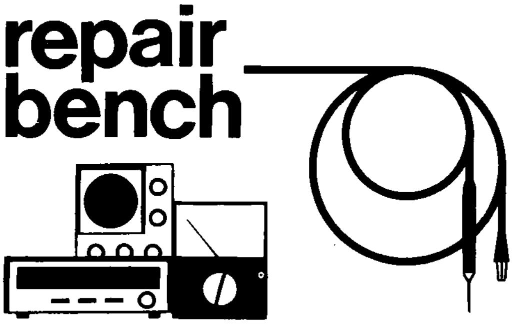
1. Frederick E. Terman, *Radio Engineer's Handbook*, McGraw-Hill Book Company, New York, 1943, page 55.
2. Michael J. Gordon, Jr., WB9FHC, "Calculating the Inductance of Toroids," *ham radio*, February, 1972, page 50.
3. Charles Miller, W3WLX, "Toroidal Coil Inductance," *ham radio*, September, 1975, page 26.

ham radio



Bob Stein, W6NBI

how to use the lab-type rf power meter

When the average amateur hears the words "rf power meter," they bring to mind instruments such as those made by Drake, Swan, Collins, Heath, and others. These are all designed for relatively high powers in the high-frequency and low vhf regions, having full-scale ranges from 20 to 2000 watts. In a more professional class are the through-line instruments manufactured by Bird and Sierra, which use plug-in elements for various frequency ranges between 2 and 1000 MHz, and for full-scale power levels between 1 and 1000 watts.

Power meters such as these are suited for both field and home station use, and require no power source other than the rf energy being measured. However, when it comes to measuring low power, such as the output of a local-oscillator chain, a signal generator, or low-power stages in a solid-state transmitter, they are relatively useless. In such applications, it is necessary to use a laboratory-type power meter which is capable of measurements at the milliwatt and microwatt levels. And, as with most electronic test equipment, a sensitive instrument can be used to make measurements above its range by means of certain auxiliary equipment. Thus, this type of power meter can be a versatile measurement tool for all power levels normally encountered by the experimenting amateur.

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instrument availability

As with the first article[1] in this series, this one is intended to demonstrate the use of test equipment which is more or less generally available on the surplus market. The instrument which is most often seen is the Hewlett-Packard model 430C Microwave Power Meter. Similar instruments also available from surplus sources are the Sperry Microwave Average Power Meter model 31A1 and the Narda model 440C Microwave Power Meter. Before the word "microwave" turns you non-uhf types off, read on. Microwave is a misnomer, although it must be admitted that these instruments were designed for use at frequencies up to 40 GHz. But they may also be used at 10 MHz and lower; it all depends on the bolometer being used with the power meter.

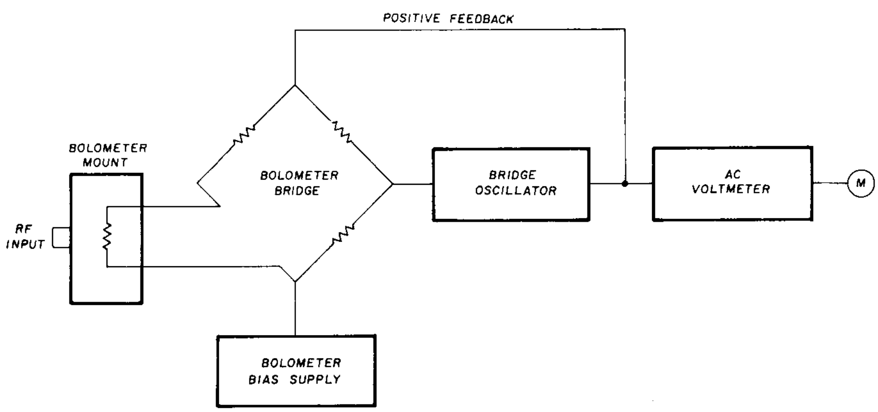


fig. 1. Simplified block diagram of the basic power meter. The bridge oscillator is tuned to approximately 10 kHz.

Which brings us to the next point – without a bolometer, a power meter is rather useless. The bolometer is a power-sensing device which is mounted separately from the power meter and connected to it by means of a cable. There are two types of bolometers: *barretters,* which are normal resistance elements with a positive temperature coefficient, and *thermistors*, which are manufactured from metallic oxide materials which exhibit a negative temperature coefficient. In its simplest form the barretter may be a short length of very fine wire, such as an instrument fuse, or a metallized film resistor.[2]

The three power meters mentioned above can be used with either a thermistor or a barretter mount.* Typical of these (for 50-ohm coaxial systems) are the Hewlett-Packard model 476A Bolometer Mount, which employs barretter fuses, and the model 477B Thermistor Mount. Several others are listed in **table 1**. Obtaining the bolometer mount is generally more difficult than finding a power meter, but they are around. On the other hand, a simple barretter mount suitable for use to over 500 MHz has been built by W6VSV and will, hopefully, be the subject of a forthcoming article.

Later model power meters, such as the Hewlett-Packard model 431 series, are also beginning to show up surplus. This type requires a temperature-compensated thermistor mount, typically a Hewlett-Packard model 478A or 8478B. The instruments may be differentiated by the bolometer connector on the front panel. The older type power meter and bolometer are interconnected via a simple coaxial cable; the newer temperature-compensated thermistor mount connects to the power meter by means of a cable with multi-pin connectors.

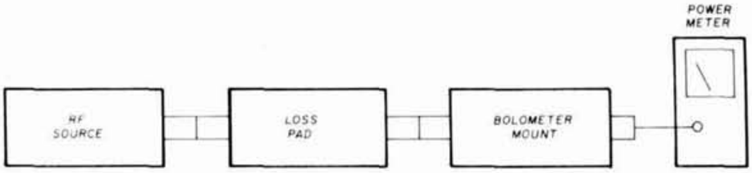


fig. 2. Using a loss pad to extend the range of the power meter. The pad must be capable of dissipating the power output of the rf source.

how it works

The bolometer-power meter combination functions because of the fact that the bolometer element is essentially not frequency sensitive within its specified range (although there are some variations which are plotted at selected points on the newer mounts). This means that equivalent amounts of power, from dc to the maximum specified frequency, will produce the same resistance in the element.

In the earlier power meters, the bolometer forms one leg of a bridge which is connected in the feedback loop of an audio oscillator; see **fig. 1**. Dc bias is also applied to the bolometer element. This configuration results in a self-balancing circuit. The combination of dc bias and audio-frequency power applied to the bolometer causes it to assume a resistance value which balances the bridge. In practice, this balance is achieved by adjusting the bolometer bias controls on the power meter so that the meter indicates zero with no external rf power applied to the bolometer.

When the bolometer is connected to a source of rf power, it heats and its resistance changes, unbalancing the bridge. Since the bridge-oscillator circuit is self-balancing, and the dc bias is fixed when the meter is zeroed, the bridge rebalances itself by reducing the audio oscillator power by an amount equal to the external rf power. This *decrease* in oscillator power is measured by a voltmeter circuit which indicates an equivalent power *increase* representing the external rf power applied to the bolometer.

The Hewlett-Packard model 430C and similar power meters are designed to work with both positive (barretter) and negative (thermistor) temperature-coefficient bolometers which have operating resistances of 100 or 200 ohms. Front-panel switches on the power meter change the bolometer bridge configuration to accommodate these variations.

Because simple bolometers, especially the thermistor type, are extremely sensitive to changes in the ambient

The Hewlett-Packard model 430C Microwave Power Meter is commonly available on the surplus market at reasonable prices. It may be used with either a thermistor or a barretter mount.

*It appears to be standard in the industry to designate the sensor as either a thermistor mount or a bolometer mount, the latter being a barretter mount. In this article, the two will be differentiated as thermistor or barretter mounts, with the term bolometer used as the overall category which encompasses both types.

temperature, improved thermistor mounts have been developed which are temperature compensated by incorporating additional thermistors in the mount. This requires additional circuitry in the power meter which uses the compensating thermistors to maintain a balanced detection bridge under changing temperature conditions.

thermistors vs barretters

A comparison of thermistors and barretters applies only to the Hewlett-Packard model 430C, the Sperry Microwave model 31A1, the Narda model 440C, and other similar types which are designed for uncompensated bolometers. (The later, improved power meters all use temperature-compensated thermistor mounts).

In general, thermistors are more sensitive, have a greater power range, and are less susceptible to overload and burnout than barretters. On the other hand, a barretter responds more quickly because it has a shorter time-constant, and thus is able to follow a modulation envelope better. However, either may be used to measure the *average* power of a modulated signal.

The previously mentioned power meters will operate with either 100- or 200-ohm bolometers. (This refers to the bridge resistance, not the bolometer's input impedance which is usually 50 ohms). Since all of these power meters are able to provide a wide range of bias currents, any bolometer which allows the power meter to be zeroed is suitable for use with that instrument.

Because of the susceptibility of low-level barretters to burn-out, the coarse zero-set control on the power meter must always be turned fully counter clockwise before the bias-current switch is turned on *or* off. This avoids putting a switching transient through the barretter, which might cause it to burn out.

measuring power below 10 milliwatts

Measuring rf power within the power range of the bolometer and power meter (usually 10 milliwatts maximum) is as simple a procedure as can be imagined. First, connect the bolometer mount to the power meter with an appropriate cable, then set the power meter resistance

The Hewlett-Packard model 477B Thermistor Mount is a 200-ohm, negatively-temperature-coefficient bolometer. It can be used with the Hewlett-Packard model 430C, the Sperry Microwave 31A1, or the Narda model 440C power meters.

and polarity switches to the positions which correspond to the bolometer being used. If a temperature-compensated thermistor mount is used, set the calibration-factor switch on the power meter to the factor which is specified on the mount.

Energize the power meter and zero the meter on the power range to be used. If you use an uncompensated thermistor or barretter, this only entails setting the bias-

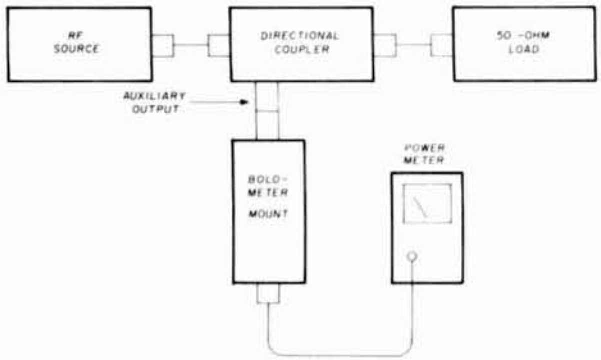


fig. 3. Using a directional coupler to extend the range of the power meter. The range may be further increased by inserting a loss pad between the auxiliary output port and the bolometer mount.

current switch to the lowest current range which will allow the meter to be zeroed by means of the zero-set controls. When using a barretter, be sure to observe two precautions:

1. Make sure the coarse zero-set control is turned fully counterclockwise before you turn the bias-current switch, and

2. Do not exceed the maximum safe current specified for the barretter.

When using a temperature-compensated mount, there may be some differences among the various power meters made by different manufacturers. Therefore, the instruction book for the power meter being used should be consulted.

Allow the power meter to warm up. In the case of the older types used with uncompensated bolometers, an hour or more may be required to reach a stable operating temperature, especially on the two lowest power ranges. It is advantageous to have the bolometer mount connected to the device under test during this warm-up period so that the mount and the device under test are at the same temperature.

After the power meter and bolometer have warmed up, re-zero the meter. Turn on the rf power source to be measured and read its output directly from the power meter. Since bolometers have time-constants as high as one or two seconds, you must take this delay into account before reading the meter.

When using the older types of power meters and bolometers, severe drift occurs on the two most sensitive ranges (0.1 and 0.3 milliwatt), even after several hours of warm-up. This drift can be minimized, but by no means eliminated, by protecting the bolometer mount from drafts and other environmental changes. One way of doing this is to enclose the mount in a block of styrofoam which has been cut and formed to fit closely around the mount. Even wrapping several layers of cloth around the housing will be an improvement over a "bare" mount.

Despite such attempts to stabilize the bolometer, I have found it necessary to re-zero the meter after every reading on the 0.1-milliwatt range. Consequently, I generally take between two and six readings, and average them to arrive at a meaningful measurement.

At this point there may be a question as to why powers of 0.1 milliwatt or less would be of interest to an amateur. A good example of this is checking or calibrating the output level of a signal generator. For instance, there are a great many TS-497/URR signal generators available surplus. Although this military version of the venerable Measurements Corporation model 80 has seen its day, it is quite adequate for the experimenter who is interested in a 2- to 400-MHz instrument. The main problem with such military surplus is that the output level may have to be readjusted. Since the calibration must be made at 50 millivolts output, a sensitive indicating device is required over the frequency range of the signal generator. A power meter and bolometer can do this nicely above 10 MHz, since 50 millivolts across 50 ohms is 50 microwatts, which is half-scale on the 0.1-milliwatt range of the power meter.

measuring power above 10 milliwatts

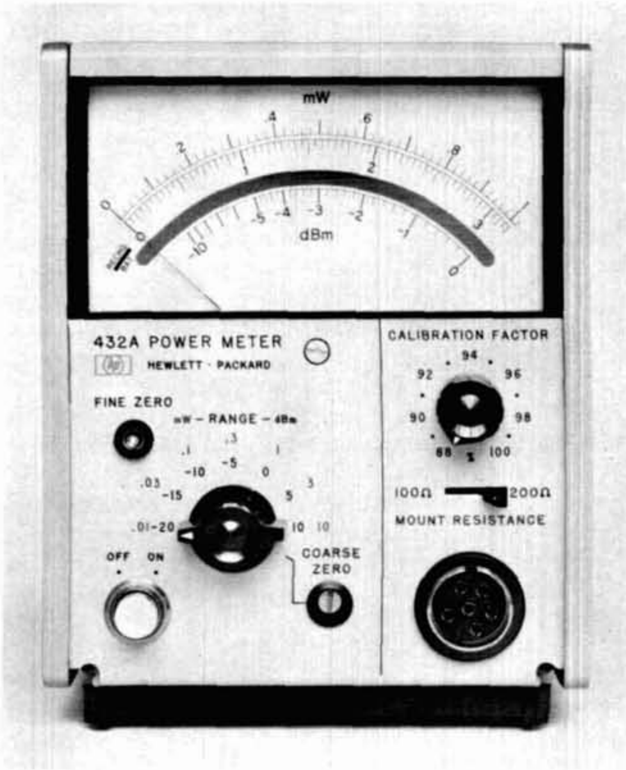
Because most of the power meters and bolometers which are likely to be available have a maximum power limit of 10 milliwatts, signals above that power level must be reduced by some means to use the instruments.

The most convenient method is to introduce attenuation between the rf source and the bolometer in 10-dB steps, which will allow you to multiply the power-meter readings by multiples of 10. Thus, to increase the range to 100 milliwatts, a 10-dB loss pad can be inserted between the source and the bolometer mount, as shown in **fig. 2.** In the same manner, the power-meter range can be extended to 1 watt by the use of a 20-dB pad, although you must be certain that the attenuator can dissipate 1 watt. The 10-watt level can be reached by using a 30-dB pad, but this must be a power attenuator rated at 10 watts dissipation or greater.

table 1. Frequency ranges and standing-wave ratios of typical bolometer mounts.

model	type	frequency range and swr
Hewlett-Packard 476A	barretter	20 - 500 MHz: less than 1.15 10 MHz - 1 GHz: less than 1.25
Hewlett-Packard 477B	thermistor	50 MHz - 7 GHz: less than 1.3 10 MHz - 10 GHz: less than 1.5
Hewlett-Packard 478A	thermistor (temperature compensated)	10 - 25 MHz: 1.75 maximum 25 MHz - 7 GHz: 1.3 maximum 7 - 10 GHz: 1.5 maximum
FXR N218A	thermistor (temperature compensated)	10 MHz - 10 GHz: 1.5 maximum
Narda 560	*	20 MHz - 1.5 GHz: 1.5 maximum
Narda 561	*	0.5 - 10 GHz: 1.5 maximum

*May use any one of several bolometer elements, either thermistor or barretter types. Model number is for mount only.



The Hewlett-Packard model 432A Power Meter is an improved instrument designed to be used with 100- or 200-ohm temperature-compensated thermistor mounts.

For powers over 1 or 2 watts, a directional coupler will usually be more convenient to use than a loss pad. This arrangement appears in **fig. 3.** A 50-ohm load, capable of dissipating the output power of the rf source, is connected to the main output port of the directional coupler, which must also be able to handle the full output power. The bolometer mount is connected to the auxiliary output port. Knowing the coupling factor (in dB) of the directional coupler, you need only to multiply the power indication by the power ratio equivalent to the coupling factor to obtain the actual power.

It is also possible to use a combination of both methods to reduce the power to 10 milliwatts or less. For example, let's assume that the power to be measured is expected to be 20 watts, but a 30-dB directional coupler is all that is available. Since 30 dB represents a power ratio of 1000, this would only extend the power-meter range to 10 watts. Thus attenuating the 20-watt input by only 30 dB would result in 20 milliwatts at the auxiliary output port of the coupler, but this can be reduced by inserting a low-power loss pad between the coupler and the bolometer mount; a 10-dB pad would increase the overall attenuation to 40 dB, permitting measurements up to 100 watts.

measurement accuracy

As with all rf measurements, there are many factors which affect the accuracy of the power measurements described. The one of major significance for amateur purposes is the loss caused by a mismatch between the rf source and the bolometer mount. **Table 1** shows the frequency range and swr of several mounts. Note that, except for the Hewlett-Packard model 478A between 10 and 25 MHz, all have maximum swrs of 1.5:1 or less. This is a reasonably good load, especially if the source can be tuned to conjugately match the bolometer mount, as evidenced by maximum output power. Of course, using the measurement configurations shown in **figs. 2** and **3** reduces the measurement uncertainty to a virtually insignificant figure because of the matching improvement provided by the directional coupler and load and/or the loss pad.

If, however, the bolometer is fed directly from the power source, and the output of the source is fixed, there will be a loss in available power because the load (bolometer mount) impedance will probably not provide a conjugate match to the source impedance. The limits of this loss can be determined from **fig. 4**, where the solid diagonal lines represent the minimum loss and the broken lines the maximum loss.

As an example, assume that you want to measure the output of an amplifier which has been adjusted for maximum output power into a known 50-ohm load. We can assume then that the output swr of the amplifier (the source) is 1.0:1. Measurements are being made with a bolometer mount (the load) having a maximum specified swr of 1.5:1. The intersection of the 1.5:1 mount-swr line and the 1.0:1 source-swr line lies at

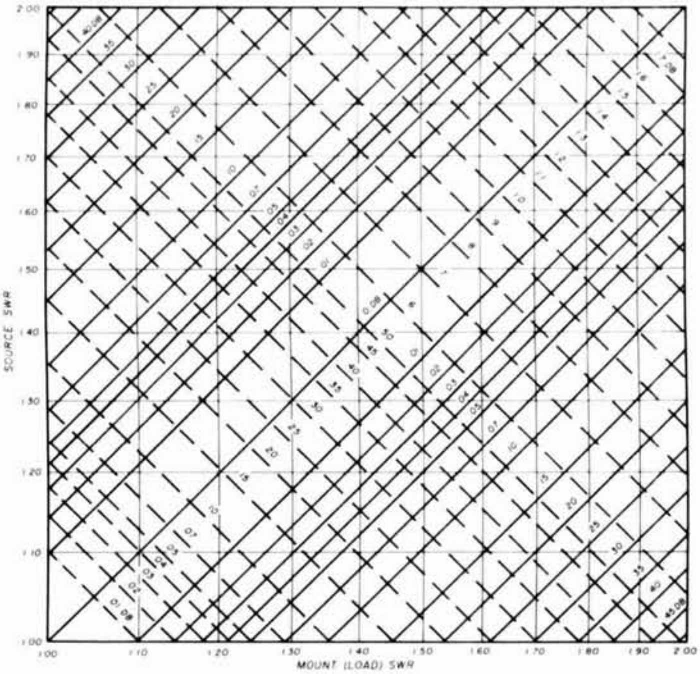


fig. 4. Mismatch loss as a function of source and load swr. The solid diagonal lines indicate the minimum possible loss, the broken lines the maximum loss.

approximately 0.17 on the solid diagonal lines and at approximately 0.17 on the broken lines. Since the minimum and maximum losses are equal, there will be a 0.17-dB mismatch loss.

If the same bolometer mount is used to measure the output of a signal generator having a specified output swr of 1.2:1, it can be seen that the intersection of the 1.5:1 mount-swr line and the 1.2:1 source-swr line is at about 0.055 on the solid diagonal lines and 0.37 on the broken lines. Therefore the mismatch loss will be between 0.055 and 0.37 dB.

Other factors which affect measurement accuracy are instrument error, miscellaneous rf loss, dc-to-microwave substitution error, and thermoelectric-effect error. A complete discussion of these errors appears in reference 3, along with a more detailed explanation of mismatch loss.

The Hewlett-Packard model 478A Thermistor Mount is typical of temperature-compensated units which can be used with the Hewlett-Packard model 431 and 432 series of power meters.

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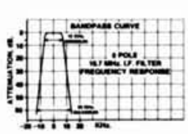
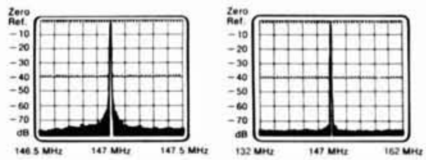


An 8-pole IF crystal filter greatly reduces adjacent channel interference.



Actual spectrum analyzer photos of the HW-2036 transmitter output operating at 147 MHz. Spurs within 20 MHz of carrier are down a full 70 dB!

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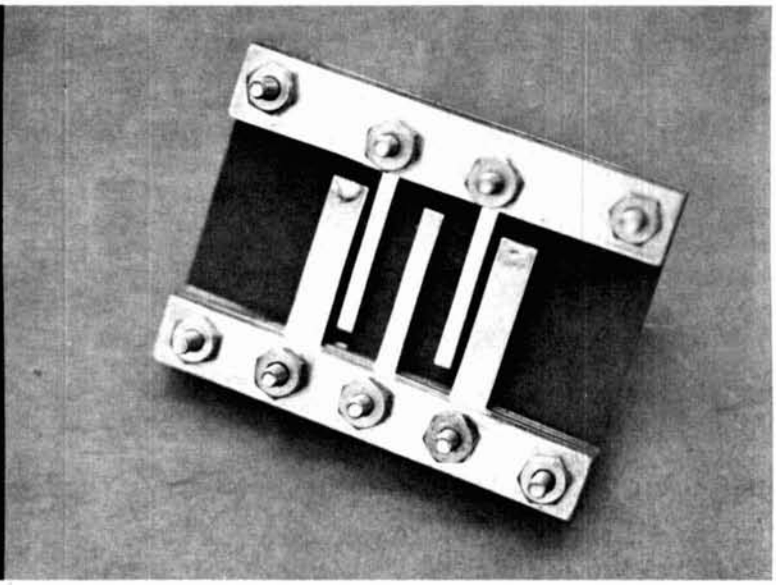


stripline bandpass filter for 2304 MHz

Interdigital filters can be easily made by using simple hand tools and this stripline design

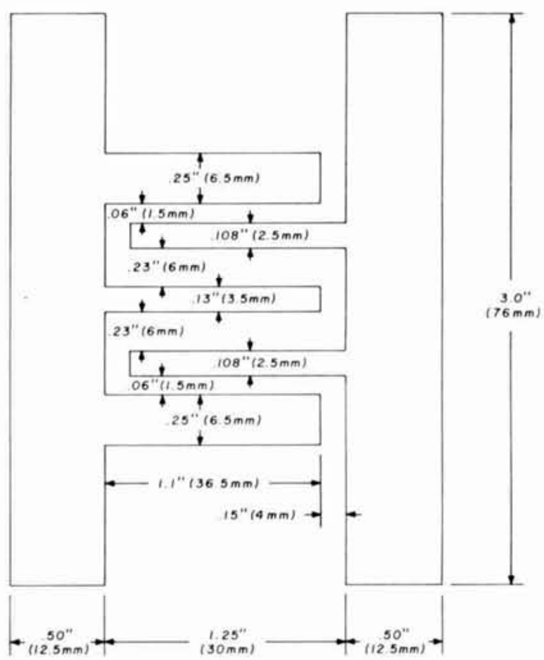


fig. 1. Details of the center conductor measurements and spacings.

By John M. Franke, WA4WDL, Apt. 225, 1006 Westmoreland Avenue, Norfolk, Virginia 23508, and **Norman V. Cohen, WB4LJM**, 7719 Sheryl Drive, Norfolk, Virginia 23505

Prior to the use of interdigital filters,[1,2] amateurs used only simple half- and quarter-wavelength coaxial-line cavities as filters. Such filters, while easy to construct, lack the needed sharp skirts – interdigital filters have the steep skirts and are only moderately more difficult to construct. However, the need for a lathe to square up the rod ends prompted us to investigate another form of the interdigital filter. An article by John R. Pyle[3]

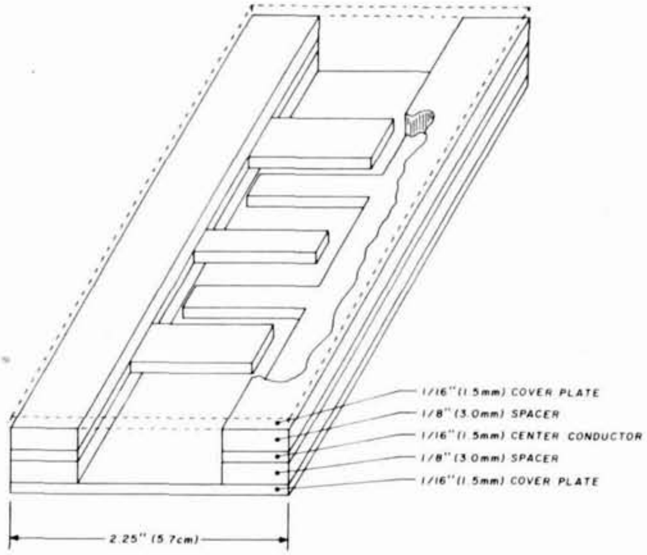
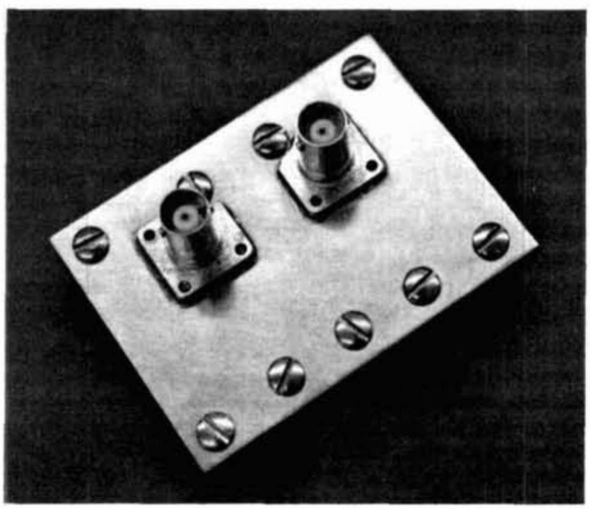


fig. 2. The sandwich construction of the interdigital filter showing the center conductors and the spacers.

provided the necessary design curves for fabricating stripline interdigital filters. His paper provided the needed graphs for filters of 1 to 10 percent bandwidth with up to 8 fingers. The curves are normalized for a 1/16-inch (1.5mm) thick center conductor between two ground planes separated by 5/16-inch (8mm).

filter construction

One such filter recently constructed is shown in the photographs. This filter was designed for 2.5 GHz with a bandwidth of 10 percent. The filter preceded a times-4 multiplier to 10 GHz. The dimensions of the center conductor are shown in **fig. 1**. For other frequencies, the finger widths and spacings are held constant and the fingers are made one quarter-wavelength long. The overall sandwich construction is illustrated in **fig. 2**. The



Top view of the filter showing the input and output connectors.

center conductor was made from brass sheet to avoid having to solder the fingers to the root strips. The 1/8-inch (3.2mm) spacers were made from a copper ground strap although brass could be used. The cover plates were fabricated from 1/16-inch (1.6mm) sheet brass. The input and output BNC connectors are sweat soldered to one of the cover plates. The connector center pins were trimmed to just touch the center conductor when assembled, and were soldered to the center conductor before attaching the other cover plate. The sandwich is bolted together with eight 6-32 (M3.5) machine screws. Note that a screw is located near the root of each finger to reduce the contact resistance where there is high circulating current. Alternately, the entire assembly could be sweat soldered together once the fingers have been trimmed. After assembly, the entire unit was given two coats of clear lacquer for

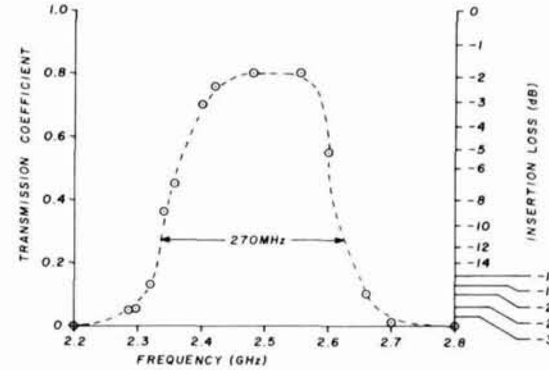


fig. 3. The measured response of the completed filter.

corrosion protection. The only tools used in the construction were a hacksaw, an electric drill, and an assortment of hand files.

The response of the completed filter is shown in **fig. 3**. The midband loss could be reduced by silver plating the unit but we felt this was not worth the added cost. If the center frequency is too high, file away at the spaces between the fingers.

The stripline interdigital filter is a useful alternative to the popular coaxial interdigital filter. The stripline construction is easier for amateurs lacking a lathe and is mechanically stable since the fingers are an integral part of the frame. Future work will involve using double-clad printed circuit board stock for the center conductor to get an even more simple and rugged filter.

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ham radio

the antenna-transmission line analog

a key to designing and understanding antennas

A practical, non-mathematical discussion of a technique used by professional engineers to design and analyze antenna performance — it is equally applicable to amateur antennas

Mr. Boyer is a prominent antenna consulting engineer who holds twelve patents in antennas, wave filters, and radar targets. Probably best known to amateurs is his low-profile DDRR antenna, which was selected by an international board as one of the 100 most significant inventions of 1963. He has served as an expert consultant to all branches of the U.S. armed forces, and to NATO, NASA, and the Institute for Defense Analysis. Mr. Boyer was previously licensed as W8PVL and now is W6UYH.

At the present state of the electronics art, the antenna represents the most rewarding, fun-filled, and low-cost area remaining for amateur experimentation. In addition, there is always the challenging possibility of making a real contribution to technology. In all cases, experimentation with radiators will invariably result in improved on-the-air signals.

Many creative amateurs who begin to investigate antennas, however, become frustrated. They have no difficulty understanding certain principles given in elementary treatments of antenna theory, and such initial knowledge carries them through an early fun period of cut and try; but some of the results obtained from such experiments are confusing and demand explanations not found in non-professional books on antennas. If the amateur persists in his experimentation and becomes seriously interested, he finally gets to a point where he wants to know — *before* stringing up more wire or guying up more sections of metal tubing — answers to questions such as, "How do I tailor my antenna design so I can tune over the entire band without the vswr on my feedline climbing to magnitudes into which my rig refuses to load? How much coil reactance does it take to resonate my old 75-meter vertical antenna on the 160-meter band? How efficient is my 20-meter center-loaded whip on the station wagon?" Beyond this, many hams would like to try out their own ideas for antennas but want to know beforehand, with reasonable accuracy, how their brainchild is going to perform on the air.

Giving up on the elementary texts, some of these same amateurs turn to the professional antenna literature for help but are usually stopped cold. Unless they are already engineers by training, they are taken aback by pages covered with the esoteric symbols of the higher mathematics: Fourier series and transforms, Bessel functions, Legendre polynomials, and everywhere copious use of the integral and differential calculus. Few

By Joseph M. Boyer, Antenna Consulting Engineer, *17302 Yukon, Suite 63, Torrance, California 90504*

amateurs want to go back to college in order to pursue a hobby. Even in the communications industry, many engineers feel there is something mysterious and scary about antennas and leave their design to a handful of specialists.

There is, however, a relatively easy way to avoid the need for using high powered mathematics while getting good, workable answers of engineering accuracy to your antenna questions; answers which will permit you to design complex antennas and predict their performance *before* you build them. The key to all this goes by the rather complicated sounding name of *Antenna/ Transmission Line Analogue*, but the only complicated part about it is its name.

What the Transmission Line Analogue does is to permit an on-paper conversion of your own particular antenna or concept into an equivalent rf transmission line. Once done correctly, this analogue key cranks out answers like your own private computer terminal. You might suspect that because the analogue key dispenses with higher mathematics, it must be some inaccurate, slip-shod method shunned by the real professionals. This is far from true – the analogue key method is used daily by working professional antenna engineers to design commercial and military radiators of all types for use in the frequency spectrum from 10 kHz on up. In its most fundamental form it was used by the brilliant antenna theoretician Dr. Schelkunoff[1] in evolving his powerful mode theory of antennas. As you gain familiarity with the analogue key by usage, you will become positively ingenious in figuring out ways to extend its application into the most involved antenna situations.

To really use the analogue key effectively, however, you must first understand how and why it works and where it comes from. The material which follows may lead you back over some familiar territory, but the route is necessary to establish a certain basic way of thinking about antennas.

primary mode waves on cylindrical antennas

Fig. 1A shows a perfectly straight, center-fed cylindrical conductor magically suspended without support in free space. You may recognize it as the familiar doublet antenna, having a total length of $2h$ and a half length, h. Its uniform conductor diameter, d, is twice its radius, a. For the moment, forget just how long $2h$ or h is supposed to be in electrical degrees at the operating frequency f (hertz). This doublet antenna has two center input terminals labeled A and B. If you could connect a very accurate rf impedance bridge (complete with its own built-in signal source) directly to the antenna input terminals A - B, without disturbing the invisible fields surrounding it in space, the bridge would read out the doublet's input impedance $Z_{in(A,B)}$ in the form of two separate parts, R_{total} and jX. The R_{total} part is the real or resistive part of the entire complex impedance $Z_{in(A,B)}$; the X part with the lower-case j complex operator in front of it (as a label to make sure you separate it from the real part) is reactance. The bridge cannot tell you that the resistive part R_{total} is really made up of two separate resistive parts, or

$$R_{total} = R_r + R_l$$

The R_r is a resistance-like term called the radiation resistance which is a measure of how much wave energy is lost from the antenna by radiation per *rf cycle*.

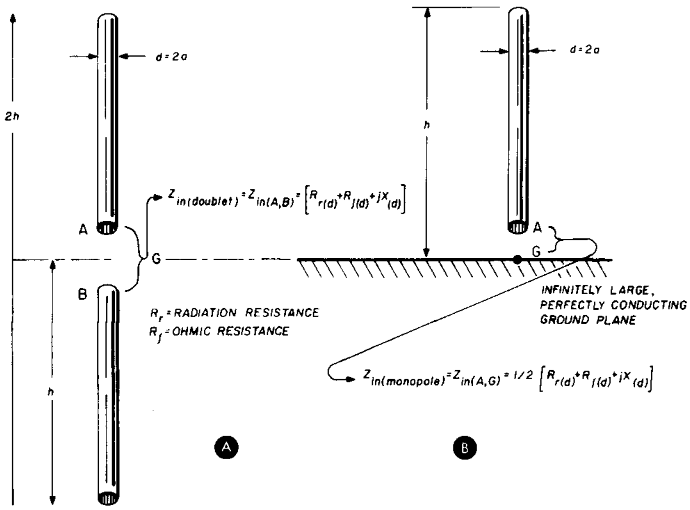

fig. 1. Center-fed, cylindrical doublet antenna in free space (A), and its feedpoint impedance. Equivalent monopole antenna of equal half-length, h, and equal radius, a, operated against an infinitely large, perfectly conducting ground plane is shown at (B).

No one wants the R_l ohmic loss resistance part of R_{total}. It just causes some of your input power to terminals A - B to be converted into heat, yet it is always present in real-world antenna elements. One of the battles in antenna design is to keep the ohmic part as small as possible in ways which will be discussed later. In any event, carefully log the input impedance,

$$\begin{aligned} Z_{in(A,B)} &= R_{total(d)} + jX_{(d)} \\ &= R_{r(d)} + R_{l(d)} + jX_{(d)} \ ohms \end{aligned}$$

which you have measured for this particular *doublet* antenna of specific half length h, and particular conductor radius a, at some exact rf frequency f (hertz).

Now imagine that the half of the antenna connected to input terminal B suddenly disappears, leaving only the other half-length element suspended in free space. A very thin, infinitely large, perfectly conducting metal sheet is then placed exactly through the mid-way point between the former input terminals, the sheet forming a plane lying at a right angle to the remaining antenna half (**fig. 1B**). The remaining antenna terminal A is now spaced a small distance above the metal plane. By connecting a ground lug to the metal plate at a point directly below terminal A, you now have a monopole antenna (half antenna) operating over a perfect ground plane. Label the newly installed ground terminal with the letter G. The remaining half of the former doublet is still as before: Same length h, same conductor radius a. With the rf bridge reconnected to the new input terminals A, G take a reading of the monopole input impedance over perfect ground. You find the new rf input impedance to be

$$\begin{aligned} Z_{in(m)A,G} &= \tfrac{1}{2}\,[Z_{in(A,B)}] \\ &= \tfrac{1}{2}\,[R_{total(d)} + jX_{(d)}] \\ &= \tfrac{1}{2}\,[R_{r(d)} + R_{l(d)} + jX_{(d)}]\ ohms \end{aligned}$$

or,

$$\begin{aligned} Z_{in(m)A,G} &= R_{total(m)} + jX_{(m)} \\ &= R_{r(m)} + R_{l(m)} + jX_{(m)}\ ohms \end{aligned}$$

As a result of this first experiment, you write yourself a rather formal note, "The complex input impedance $Z_{in(m)A,G} = R_{r(m)} + R_{l(m)} + jX_{(m)}$ of a cylindrical monopole antenna of conductor length h and conductor radius a, erected normal to an infinitely large, perfectly conducting ground plane, is exactly *one-half* the complex input impedance $Z_{in(A,B)} = R_{r(d)} + R_{l(d)} + jX_{(d)}$ of a full doublet antenna of identical half-length h and conductor radius a in free space, when both antennas are measured at the same radio frequency."

With this important experiment out of the way, step back some distance from the monopole antenna erected over the perfect ground plane so that you can inspect its entire length, h. Now, really using your imagination, assume that you own a very special pair of eye glasses which permit you to actually "see" electric field lines of force E, and magnetic field lines of force H. Carefully watching the monopole antenna, again turn on the rf generator so that it supplies energy at frequency f to the monopole input terminals A - G. **Fig. 2A** is an attempt to show what you would "see."

At the instant you closed the switch (t=1), a small expanding surface like a bubble would appear around the antenna base. Its surface would be covered with dotted E lines pointing radially outward from the surface of the monopole conductor element, with each E line gracefully arching over so that it pointed directly down at right angles to the flat surface of the ground plane. At the same instant you would preceive dashed circles of magnetic field lines of H form concentrically around the monopole antenna element, each growing larger in diameter with the passage of time. Let time suddenly freeze at this point so that you can closely inspect the initial wave surface.

The surface of the "bubble" is a wave front. As this is the first wave to be introduced to the monopole antenna, it's called a *precursor*; a sort of "scout wave" sent out to explore the electrical nature of the yet-unknown antenna to determine — at this one particular frequency — just exactly how the waves to follow will have to finally arrange themselves to be in agreement with certain natural laws.

One of these natural laws dictates that the electric lines of force always point precisely at right angles into the surface of a good conductor such as the antenna element, *and* also point precisely into the flat perfectly conducting surface of the groundplane. Another thing: The "antenna" is the total *combination* of the monopole conductor and the ground-plane surface; the wave front or antenna field is not *in* the antenna conductors, but instead fills the space surrounding the monopole conductor element and the ground-plane surface. The antenna conductor monopole element (or each half of the doublet in free space), together with the ground plane, compose the "nature" of the antenna, and are called the antenna *boundaries*. These boundaries are what the first precursor wave is trying to explore, for they *alone* will determine what finally happens later in time.

Now unfreeze time and let the wave front expand and climb higher up the antenna. Again freeze time at t=2. Now you will notice a very interesting effect: In order to span the increasing distance along the arc between the monopole conductor surface and the ground plane, the E lines get longer and longer as they climb up the monopole. You will also notice that the brightness of the E line arcs closest to the base of the antenna are less intense than those stretching over to ground from higher up on the antenna. Conversely, a fixed radial distance from the antenna, the magnetic field circles around the monopole conductor are intensely bright and glowing around the antenna base, but are less bright as they form around higher parts of the monopole conductor. Clearly, electric field intensity is *increasing* with height up the antenna; magnetic field intensity is *decreasing* with height.

Antenna specialists use the ratio of the magnitude of the electric line of force, E, to that of the magnetic field line, H, (at any point in space) to define what is called wave impedance, Z_W. This is comparable to what the electronics engineer merely calls impedance when he is dealing with the ratio of voltage, V, to current, I, in circuits physically small in terms of the wavelength of rf energy circulating within them. In contrast, antennas are "big" circuits in terms of the operating wavelength. As a consequence, their fields extend out to great distances around the antenna and the field — rather than voltage or current on conductors — is of first importance.

With this idea of wave impedance in mind, a re-examination of the monopole antenna field discloses that the wave impedance equal to E/H must be increasing in the wavefront as it expands higher and higher up the antenna (because E is increasing and H is decreasing). The wave impedance is small in magnitude near the input terminals A - G, but increases steadily with antenna height, h.

This brings up one important way of thinking about *all* antennas: Antennas attempt to perform an impedance-matching function, providing an impedance match between their input terminals and that of the surrounding space by using their conductors as wave impedance "transformers." In this picture, space itself is a common "master" transmission line connecting your antenna with every other antenna in the universe. Such a space transmission line has its own characteristic impedance, Z_s, and possesses an infinite number of input and output terminals. For the moment, however, just assume that this space transmission line surrounds your antenna; it wants to accept the rf energy you are putting into the input terminals of the antenna, but will only accept your wave energy as radiation when certain precise *boundary conditions* are met. The antenna tries to accomplish this feat of getting its waves off into space, but must wait to see what the precursor wave "says" after exploring the antenna.

Let time again unfreeze, and watch the wave front expand through t=3 to the moment of truth at t=4. The wave front has been racing (when you permitted it to) up the antenna at almost, but not quite, the speed of light (3 x 10^8 meters/second). Things appear to be going smoothly so far as the precursor wave is concerned, with those antenna boundaries changing in a nice, gentle fashion. Then *crash*! As if it had smashed head-on into a wall, the precursor wave finds the end of the monopole conductor. To the scout wave this is like an electromagnetic explosion. Let's freeze time again just at the instant this explosion occurs. What happened? To find out, we have to take a cut through the antenna field from a point looking directly down on the monopole conductor. Such a field cut is denoted by the dashed line x-y in view t=4 of **fig. 2**. The downward looking view is seen in **fig. 3**.

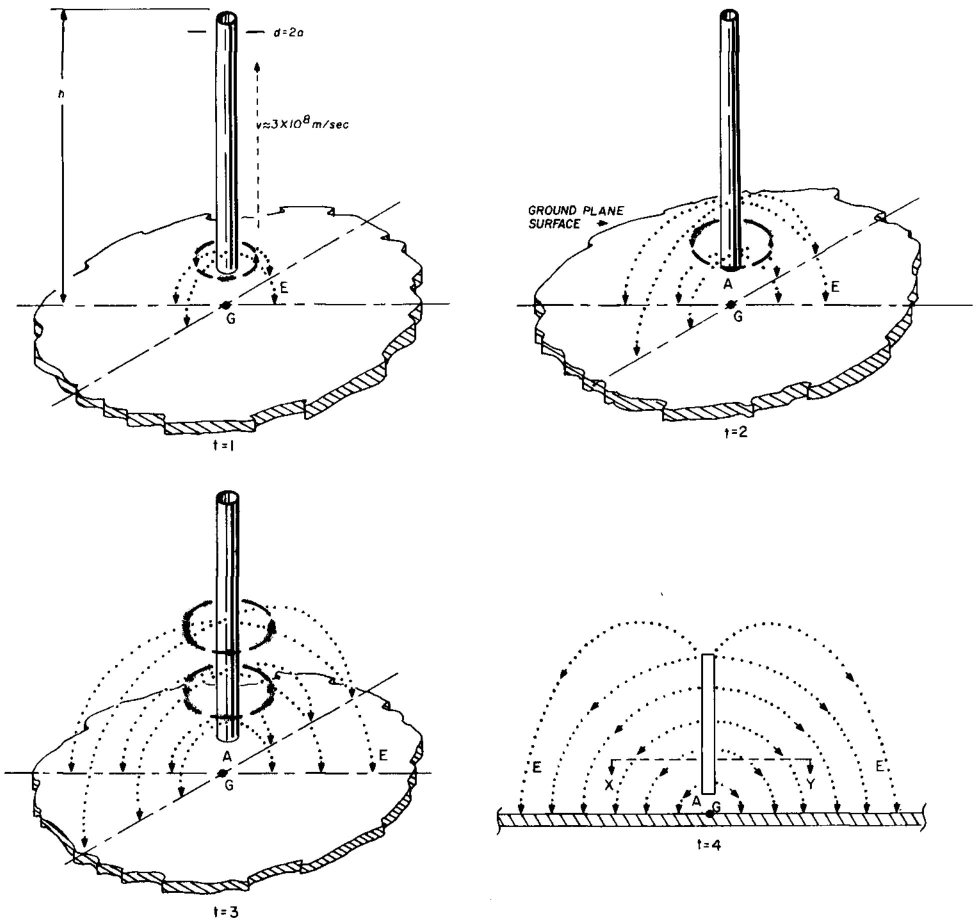

fig. 2. Precursor TEM wavefront moving up a cylindrical monopole antenna and outward on ground plane G for times t=1, t=2, t=3, and t=4. End reflection occurs at t=4. Top view of the E and H lines at t=4 are shown in fig. 3.

There are those outward pointing radial electric field lines, E, and the closed circles of magnetic lines, H. Waves in which the electric field lines lie at right angles to the magnetic field lines entirely in the plane of the wavefront, so that there are no electric or magnetic field components in the direction of wave propagation, are called *transverse electromagnetic* or TEM modes in wave shorthand. **Fig. 3B** shows the wave front inside an ordinary coaxial transmission line with air as an insulator. The similarity of the waves on the antenna and those in the non-radiating coaxial transmission line is no coincidence. Both are type TEM mode waves. In a TEM mode wave, the electric field lines *must end* on the surface of conductors; the monopole antenna conductor and the ground plane serve the same boundary purpose as the inner conductor and inside surface of the shield in the coaxial transmission line. In both cases, the wave fronts are guided in the empty space by the ending of the electric lines onto the oppositely polarized conducting surfaces. The TEM mode waves are held to these boundaries in the same manner as a spider with sticky feet when running around his web.

Now let's try to think the way an antenna theoretician does. Here we have these two classes of waves: TEM mode waves on the antenna in which all electric lines must end on conductors, and space or radiation waves. In free space (say halfway to the planet Mars) there are no electrical conducting surfaces upon which the electric field lines in space waves can end. But we already know that radio waves *can* propagate through free space. That must mean that the kind of waves which can exist as radiation must — regardless of wavefront geometry — contain electric lines which *close on themselves* or form loops the way the magnetic field closes on itself around our antenna. This idea turns out to be correct. There is an infinite variety of space wave modes, but *none* of them includes the TEM mode wave. Therefore, a pure TEM mode *cannot* make the transfer from the antenna boundaries into free space. Free space is an incompatible boundary condition for the TEM mode, precursor wave. Such incompatibility constitutes a huge *impedance mismatch* to the guided TEM mode wave at the *end* of the antenna.

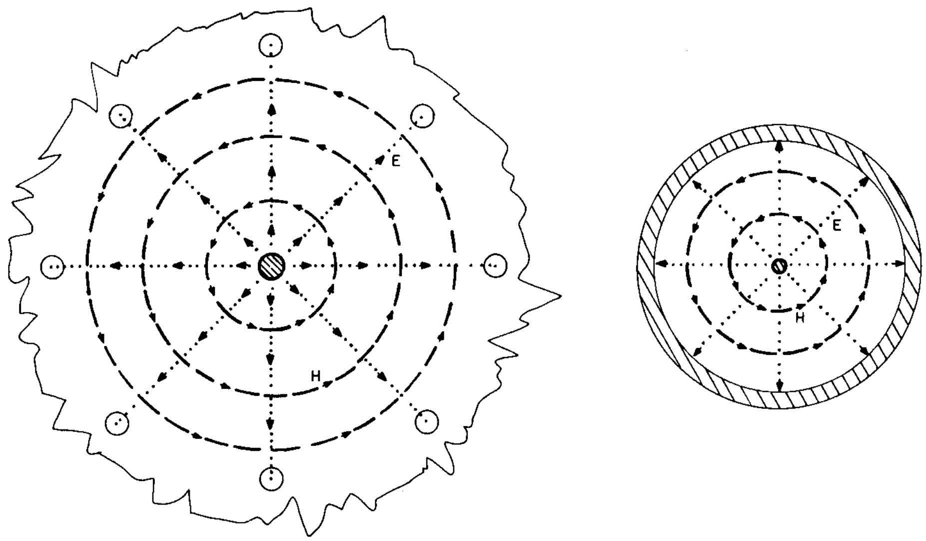

fig. 3. View looking down on the antenna field at x-y plane indicated in fig. 2. Note similarity to the E and H lines of the wavefront in a coaxial transmission line to right; both fields are type TEM.

Faced by a large mismatch at the top of the antenna, the TEM mode wave does what all waves do when faced by a mismatch on a transmission line: it is reflected and starts back down the antenna toward the input terminals. In doing so, however, the scout wave encounters other TEM mode waves coming up the antenna in the opposite direction. This kind of situation, with coherent waves moving in opposite directions, always produces the same phenomenon: Standing waves. Note, however, that these standing waves exist in space along the entire length of the antenna and are *not* to be confused with similar standing waves which can form in an antenna feedline because of an impedance mismatch between the antenna input impedance and the line's characteristic impedance.

It should be noted that the TEM scout wave is *totally* reflected from the open end of the antenna; absolutely no wave energy got away into space as radiation. When a total reflection occurs, the only thing that prevents the space standing wave from reaching a vswr of infinity to one is the very small ohmic resistance of the highly conducting antenna element. Obviously, if that situation continued, antennas, as we know them would not exist. Fortunately, for radio amateurs and antenna men, it doesn't. Nature has arranged things so that as the downward moving scout wave continues to interfere with more and more TEM mode waves coming up the antenna, a wave conversion results; some of the original TEM mode energy is transformed into new, higher order mode waves — wave types which possess closed E and H line geometry and which can make the transfer from the antenna to space. As this converted wave energy (a surprisingly small amount of the total) begins to leave the antenna, energy loss causes the near-to-infinite vswr of the space standing wave to drop to a more reasonable magnitude. The antenna has now reached its steady state of operation.

Here, now, is our antenna: It is "ringing" like a sort of electromagnetic bell as the waves (not charges) race out along the length of the antenna, smash into the top end impedance discontinuity, then race back down the antenna, performing the mode change and supporting the existence of the space standing wave. Each rf cycle produces a small loss of energy to free space as radiation. The actual amount of radiation loss per rf cycle is dependent upon the length of the antenna (h or $2h$) in electrical degrees at the operating frequency, and the conductor geometry (which, for a monopole, includes the ground plane).

Do I hear you say that this picture sounds very much like one describing the way an open-ended (open-circuited) rf transmission line operates? Let's examine that idea! We saw that the wave impedance, $Z_W = E/H$, was not uniform along the length of the antenna. We also compared a cross section of the antenna field (precursor or scout wave) to the field inside a coaxial transmission line and found them to be the same. Now, even elementary books tell us that, in a lossless transmission line, the ratio of *distributed* series inductance to *distributed* shunt capacitance per unit length solely determines the characteristic impedance of the line as

$$Z_o = \sqrt{L/C}$$

(Advanced textbooks go on to say that the wave impedance, Z_W, of the TEM mode wavefront propagating down the transmission line is also a function of this same L/C ratio in the transmission line. Standard types of rf transmission lines, however, have uniform characteristic impedance so therefore they must have a *constant* L to C ratio per unit length.

Such reasoning makes it clear that the cylindrical antenna – when viewed as an rf transmission line – must possess a *variable* ratio of L to C along its length. To reinforce this idea, make another mental experiment: cut out a short section from the monopole antenna conductor. Measure the shunt capacitance to ground of this short conductor section, first at the antenna tip height, then at the antenna midheight, and finally, at the base just above the ground plane. Intuition tells us that the shunt capacitance to ground of the conductor section will be maximum at the antenna base, less at the midheight, and least at the top of the monopole. **Fig. 4** shows this same "measurement" result for the case of a doublet antenna. If capacitance to ground (or to the other side of a doublet) varies with position, obviously the L/C ratio cannot be a constant – and that says that the antenna characteristic impedance must also be non-uniform.

But, couldn't we just take this non-uniform characteristic impedance of the antenna and use it as a transmission line model of the antenna in the analogue key method? Yes, but this approach would be a bit messy to put into practical use. Calculations for non-uniform impedance transmission lines are more laborious than those related to lines with uniform impedance. Let's try again. If you have a quantity which changes in some smooth way over a given distance such as h, it's possible to take its mean or average value. In high class mathematics, this is called the integral of something (in this case, Z_o) over the length, h. That is actually what is done: You take this mean or average value of Z_o for the antenna length, h, and then use this *average* Z_o as the *uniform* Z_o of your analogue transmission line representing your antenna.

It sounds neat, except that getting this mean Z_o for an antenna is not an easy task. It was solved back in the 1920s at great calculation labor using a dc potential method. Fortunately, later work by Dr. Schelkunoff of Bell Telephone Laboratories has given us some simple formulas to determine the average characteristic impedance of certain kinds of antenna conductor geometry. These conductor geometries include those most often used by amateurs, and professionals alike.

Before presenting these simple formulas, let's make sure we have the concept of the analogue transmission line idea clearly in mind so it may be used with confidence in our experiments on paper with antennas.

antenna into transmission line

An rf transmission line constructed from extremely high conductivity elements of copper or aluminum, with only air as insulation, represents a very low electrical loss system. Electromagnetic waves moving down such a line stay almost perfectly constant in strength or amplitude even when traveling over long distances in electrical degrees of line length. This constancy of amplitude means that you can use simple trigonometric functions such as the sine, cosine, tangent, or cotangent of the line length in electrical degrees to accurately represent the behavior of waves on low loss line.

On the other hand, if you used poor conductors such as steel or lead to build an rf transmission line, the resistance of the conductors would rob energy from the waves moving down the line and convert it into heat; as a consequence of this energy loss, the wave amplitude would decay or decrease in strength with electrical distance traveled. To represent decaying waves you have to use mathematical functions which also decay in amplitude with electrical distance: Hyperbolic functions. Finally, in correctly representing radiation loss you come up against the cosine and sine integral calculus functions. Not only are these, valuable as they are, a little tacky to use in an amateur technical journal, but I promised at the beginning that only simple math would be needed.

A decision to stick to the use of simple, everyday trig functions means that we must use a lossless equivalent transmission line to represent the antenna. Knowing that real antennas have loss, hopefully the good kind of loss called radiation resistance, how can a lossless transmission line model of the antenna give us accurate answers when solving real antenna problems? Recall the TEM wave mode which did not radiate? That TEM mode would be a uniform amplitude wave representing the major portion of the rf energy oscillating (standing) in the antenna region. If we used only this non-radiating mode wave in the analogue line representing the antenna, the answers would describe *only* the reactive behavior of the antenna at its input terminals. The real or resistive part of Z_{in} would be missing in the answer because it is radiation energy loss which the TEM mode cannot account for in antenna systems. Is that bad? Certainly not! One of the most important things you want to find out when exploring your antenna ideas on paper is how the reactance at the input terminals will change as you move your transmitting frequency over an amateur band, what jX will do if you use a loading coil in the antenna, or how jX changes from one amateur band to another.

The real or resistive part of the antenna's input impedance (which is related to radiation resistance) changes very *slowly* with frequency; the reactive part, however, varies at a much greater rate with changes in operating frequency. *The rate at which the reactive part of antenna input impedance changes with frequency is* governed by the antenna's characteristic impedance when viewed as a transmission line. Does this mean that you just forget all about the real part of Z_{in}? No, not at all. You'll end up with a complete answer for $R_A + jX_A$ alright, but you will obtain the real part, R_A, the lazy man's way: By looking up the antenna's radiation resistance, R_r, as a function of its electrical length h (or $2h$) at the operating frequency, using published graphs of this data. Then you'll add the radiation resistance to the reactive part obtained from the analogue key transmission line key model. It's that simple if a) you are using high conductivity antenna conductors such as copper or aluminum, and b) are feeding the antenna at a current maximum point such as the base of a monopole or the center of a doublet. In the rare case where you are not feeding at a current maximum point on the antenna, then you transfer the R_r value you looked up to the actual feedpoint by a method to be given later.

Incidentally, in deference to Dr. Schelkunoff, it is only fair to mention that in the equivalent transmission line method he evolved, both the real and reactive parts of the total complex input impedance are obtained by using a special lumped "load impedance," determined by separate calculations, which is placed across the end or "output terminals" of the paper analogue transmission line representing the antenna. This more sophisticated technique, however, demands use of advanced forms of mathematics.[2]

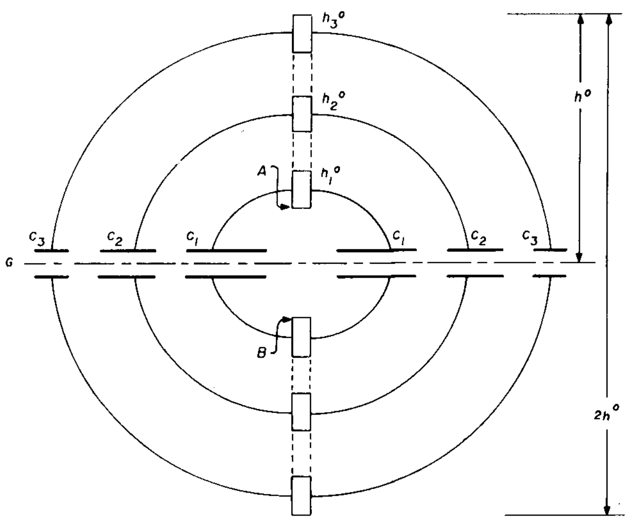

fig. 4. Variation in shunt capacitance between equal length conductor sections located at the input terminals, mid half-length, and tips of a doublet antenna. Passing the ground plane through G gives the same effect for an equivalent monopole. For a monopole, shunt capacitances double in magnitude.

characteristic impedance of cylindrical antennas

Schelkunoff gives the mean characteristic impedance of a doublet antenna with cylindrical conductor elements as,

$$K_A = 120\ (\log_e \frac{2h}{a} - 1)\ ohms \qquad \textbf{(1)}$$

Then, recalling the first experiment where you found that a monopole antenna of length h over a perfect ground plane had an input impedance exactly one-half that of a doublet antenna in free space of half length h, and the *same* conductor radius a, the mean characteristic impedance of a monopole antenna over ground is

$$K_m = 60\ (\log_e \frac{2h}{a} - 1)\ ohms \qquad \textbf{(2)}$$

Don't let the $\log_e$ part bother you. If your pocket electronic calculator doesn't give the natural $\log_e$ of a number directly, or if you only have tables of the common $\log_{10}$, then

$$\log_e \frac{2(h)}{a} = (2.3026) \times \log_{10} \frac{2(h)}{a}$$

The notation K_A and K_m is used to denote the mean antenna characteristic impedance instead of, say, Z_{oA} or Z_{om}. This avoids any confusion with the Z_o of a standard transmission line used to feed the antenna.

coming up

In the second part of this article I will describe use of the transmission line key method to solve a number of different antenna problems faced by the radio amateur. These will include the design of monopole and doublet antennas capable of being operated over the entire frequency width of an amateur band while keeping the vswr in the feedline down to a specified maximum value into which modern transmitters will load full power. I will also discuss base, center, and higher position coil loading of electrically short monopole and doublet antennas for maximum efficiency. Finally, I will show you how to "dissect" an antenna of your own design into parts to detemine if it will operate as you wish.

Each example will be carefully worked out in full detail (no steps omitted) so you can easily follow the solution and not get lost. In this way you will be able to quickly translate the analogue method to your own problems for any antenna on any amateur band. In the meantime, if you are totally unfamiliar or a bit rusty in the use of elementary plane vectors (phasors) to represent a complex ac impedance, $R + jX$, I suggest you visit the library and get a copy of *Basic Mathematics For Electronics*,[3] or its much earlier version, *Mathematics For Radiomen and Electricians*, by N. Cooke. Cooke was able to teach tens of thousands of Navy gobs to easily master basic ac math on a crash basis during WW II. You will also find him easy to follow and understand.

The difference between an amateur and a professional in a given field of science should not be one of knowledge, but only that the amateur is rewarded in pleasure and the professional in coin of the realm.

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novel indicator circuit

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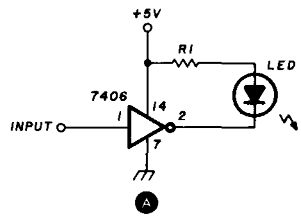
Since light-emitting diodes have become available at reasonable prices I find myself using more and more of these devices to monitor the internal circuitry in my equipment from the front panel. LEDs make it easy and economical to do this. I think you gain a better understanding of the internal functions of your equipment through the use of monitor or indicator points. When trouble occurs in the equipment it can often be diagnosed from the front panel with these indicating devices. They can pay their own way both in operator satisfaction and in ease of maintenance, not to mention their esthetic value.

I recently began using a fairly new device, the MV5491 LED manufactured by Monsanto. This device is quite different from the LEDs I had utilized in the past and has proven to be not only interesting, but very practical. Before describing this device I would like to review the standard LED drive circuits I had been using in most of my equipment to acquaint fellow amateurs who, as of yet, have not put these devices to work in their own equipment.

a standard driver circuit

The standard LED driver circuit that I have been using for the past few years is shown in **fig. 1**. This configuration uses one-sixth of a TTL hex-inverter buffer, type SN7406, to drive a LED. The SN7406 has an open collector output with the LED and associated current limiting resistor serving as the collector load for the output transistor of the IC. In this circuit, a positive



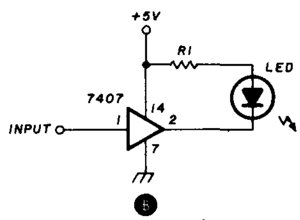


fig. 1. Basic LED drivers configured for normal and inverted inputs.

input will provide a low output at the collector, forward biasing the LED and causing it to illuminate. Resistor R1 limits the current through the LED to a safe value as specified by the LED manufacturer. This is a simple, inexpensive circuit and six LEDs can be driven from a

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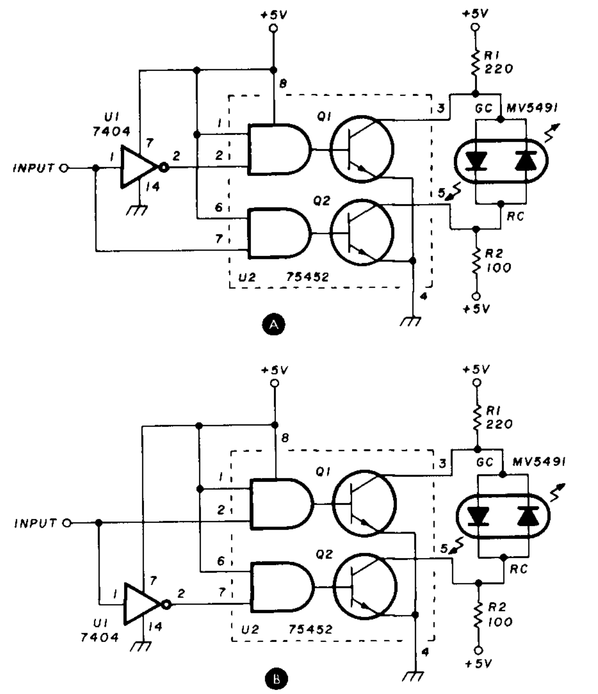

fig. 2. Two-state LED drivers. The circuit shown in (A) will provide a red indication with a positive input, (B) a green indication.

single SN7406. This makes for a very low component count and is easily added to existing equipment or new designs. The only calculation required is the value of R1, the current-limiting resistor. To calculate this we must know the voltage and current ratings of the LED. Red LEDs are usually rated at 1.65 volts and about 20 mA. Subtracting the LED voltage and the saturated SN7406 collector voltage from the source voltage, 1.65 + 0.35 from 5.00, we come up with a difference of 3.0 volts which must be dissipated across R1, at the rated current of 20 mA. Using Ohm's law we can calculate the resistor value to be 150 ohms. For the wattage requirement we can use the I^2R formula to arrive at 0.06 watts, so a ¼-watt resistor will suffice.

This circuit works well for source voltages up to about 30 volts, and of course the value of the current limiting resistor must be selected for different source voltages. Most of the LEDs have a maximum current rating of 40 to 50 millamperes and I find that approximately half the rated current will yield adequate light output and assure long life for the indicator. **Fig. 1B** is the same basic circuit but it will light the LED with a low or false input. This is accomplished by using a hex buffer, SN7407, rather than the SN7406. The SN7407 does not contain the inverter function found in the SN7406, but it is pin for pin compatible and the voltage and current ratings are alike. As can be seen in **fig. 1**, you can drive the LED with a true or a false signal, depending on the driver you select. As you get more involved in digital equipment, and I am sure we all will be more involved in this technology in the not-too-distant future, our use of circuits of this type will increase and a working knowledge of them will become more important.

The Monsanto MV5491 is in a standard package like most other LEDs, but is comprised of two light-emitting diodes connected inversely in parallel. One of the parallel diodes is red and the other is green, so steering current in one direction will yield a green light and reversing the current flow will yield a red indication. Stopping the current flow will turn both diodes off, furnishing three distinct indications from one device. The colors are a very vivid red and green, not like some of the earlier single diode devices that were red no matter what the label said!

The red diode section of the MV5491 is rated at 1.65 volts, with a test current of 20 milliamperes and a maximum current of 70 mA. The green diode section is rated at 3.0 volts, with a test current of 20 milliamperes and a maximum current of 50 mA.

The unit comes complete with mounting ring and requires a quarter-inch (6.5mm) diameter mounting hole. When drilling the mounting holes for these and other LEDs in this type package, do not bevel the edges of the holes as this tends to negate the positive effect of the plastic mounting hardware furnished with the LEDs.

the driver circuit

The project I was working on when I came across this new device used a single five-volt source, so it took a little head scratching to figure out a scheme for driving this new LED. I began scratching out diagrams on the top of the work bench and after an hour or so had an LED that would change color as the input was toggled. I was working on a piece of digital gear so I wanted the driver circuit to be ICs rather than discrete components and eventually came up with the circuit shown in **fig. 2.** This circuit worked out fine and I had an indicator that was green when the input was normal but promptly turned red when an off-normal condition was encountered.

The indicator circuit uses a single SN75452 IC driver and one-sixth of a hex inverter, SN7404, and two

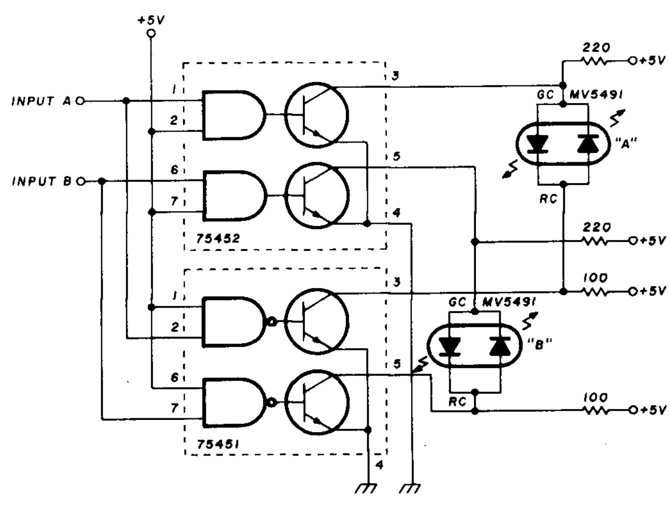

fig. 3. Two indicators using common drivers and inverters.

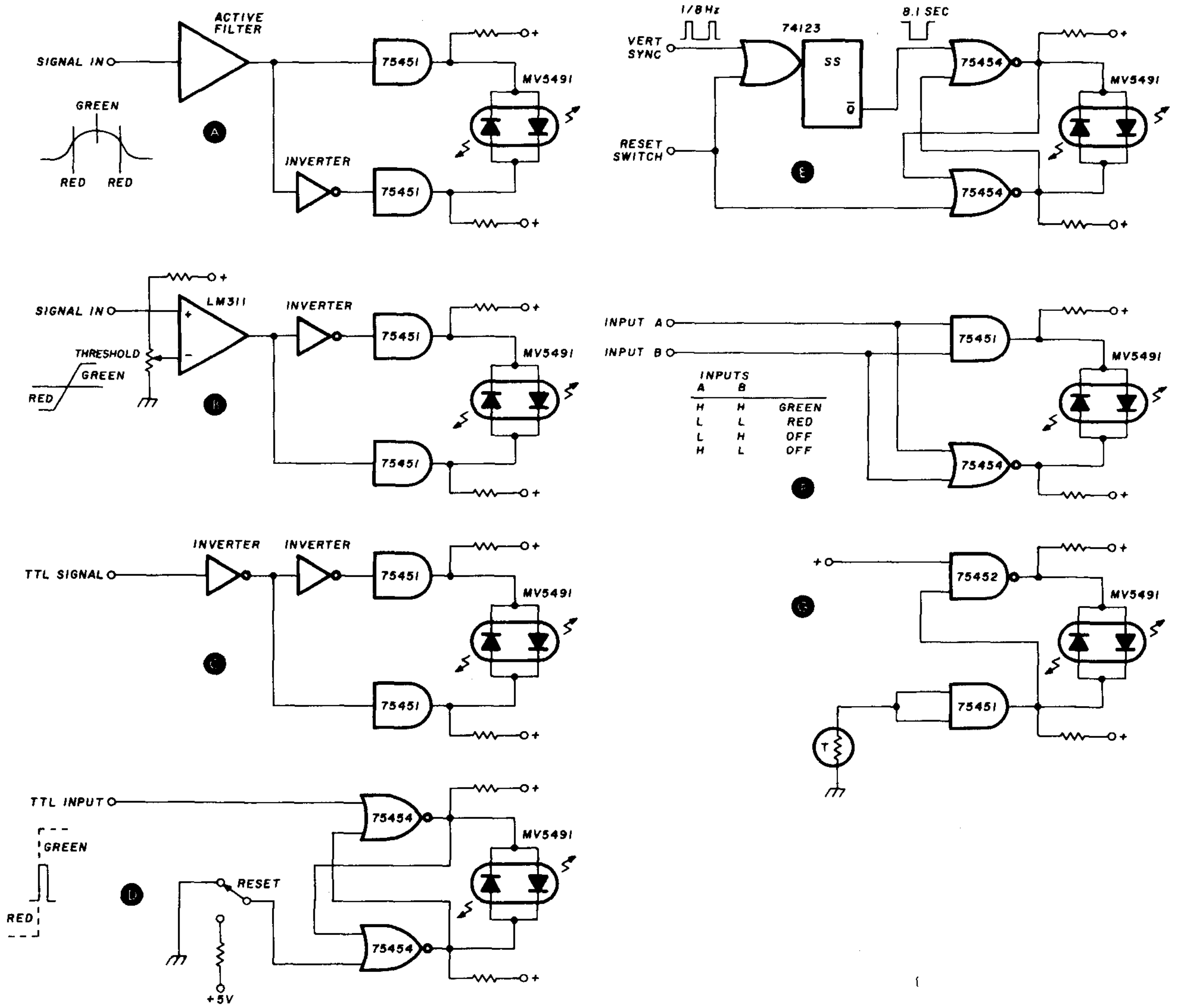

fig. 4. Applications of the MV5491 LED. (A) shows how the device is used to indicate the presence of a signal in the desired frequency range while (B) uses a comparator to indicate the crossing of a specific level. (C) and (D) can be used as logic probes. (E) is used as a vertical sync indicator for SSTV. (F) is a coincidence detector that indicates equal signals, both high and low, or no signal. (G) is used to indicate when a certain temperature has been reached.

resistors. It is much like the circuit described in **fig. 1**, but produces a more profound effect when activated. When this circuit, depicted in **fig. 2A**, receives an active or positive input, transistor Q2 of U2 is in conduction. This condition brings the red cathode of the LED to ground potential. At the same time, the Q1 section of U2 is cut-off. This brings the green cathode and red anode to a positive potential. With this condition we have the red diode of the LED forward biased (lighted) and the green diode reverse biased (extinguished).

The current through the red diode will be limited by resistor R1. A negative input to the circuit will cause the Q1 section to conduct and the Q2 section to be cutoff. This will forward bias the green diode and reverse bias the red diode. Resistor R2 will limit the current through the green diode. In this configuration, a high or positive input to the circuit will give a red indication and a low input will yield a green indication. If the opposite function is desired, green indication on a positive input, the inverter SN7404, should be placed in series with the Q2 input as per **fig. 2B**.

Do not try transposing the LED as the voltage and current specs of the two parallel diodes are different. This is the reason why the current-limiting resistors, R1 and R2, have different values. With the values shown, the red diode is forward biased at approximately 1.65 volts at 15 milliamperes with R1 setting this parameter. The green diode is set at approximately 3 volts and 20 mA by the value of R2. The resistor values are calculated in the same manner as described for the basic circuit of **fig. 1**, with the exception of resistor wattage ratings.

When Q1 or Q2 is conducting, the entire source voltage is dissipated across the resistors. The maximum dissipation in this case is about ¼ watt, so ½-watt resistors should be installed. In digital circuits where both Q and $\bar{Q}$ signals are available, such as the output of a flip-flop, the inverter portion of the circuit can be deleted and the out-of-phase signals fed to the driver circuits.

If your application requires more than one indicator circuit, the configuration shown in **fig. 3** will reduce the component count. In this configuration two different type drivers are used, one noninverting and one inverting type. These are both in mini-dip packages and will fit in a single dip socket or equivalent space on a pc board. This circuit will drive two MV5491 LEDs and requires only four external resistors. A close inspection of the circuit diagram will show that it uses the internal function of the driver IC to perform the function accomplished by the hex inverter in earlier circuit descriptions. If a number of these dual diode indicators are to be used, this is an excellent choice for a driver circuit.

applications

I am sure that there are numerous applications for the MV5491 LED, probably as many as there are amateurs still in the homebrew business. I have sketched out a few ideas that I hope to put to use in the future and possibly interest others in their applications.

Fig. 4A uses an active filter preceding the indicator circuit to form a level detector for signals at the desired frequency. This circuit should furnish a green indication on resonance and a red signal at either side of resonance. A circuit or circuits of this type might prove useful for SSTV, RTTY, or subaudio-tone indication for control purposes on fm. This is the same basic circuit described earlier which uses one drive IC and one-sixth of a hex-inverter IC.

Fig. 4B is a take-off on the tuning indicator but would provide an indication with an input change of only a few millivolts, depending on the type of comparator. When used with the LM 311 it will provide excellent performance and can be operated from a single 5-volt power source.

Fig. 4C is a TTL level detector which can be used with TTL, DTL, and RTL logic in the form of a logic probe for trouble-shooting. It will furnish a green indication on a high or plus signal and a red indication on low or false signals. The voltage to operate the circuit can be taken from the equipment under test.

Fig. 4D is a TTL pulse catcher. Again, it can be used as a logic probe for IC circuitry. In this circuit the driver circuit is cross-coupled to form a latch or memory element. A positive-going pulse sets the latch to indicate the presence of a pulse and the latch is reset manually with the reset switch. It should also work well for locating intermittents, such as glitches. You can leave the probe connected to the circuit under test and return to the problem later to see if the latch has been set by an unwanted signal. A circuit such as this one can at times, be more valuable than a scope.

Fig. 4E is a vertical sync indicator for SSTV. In this application we can trigger a retriggerable single-shot with the one-eighth hertz vertical-sync pulse. The absence of a vertical sync pulse will allow the single-shot to time out

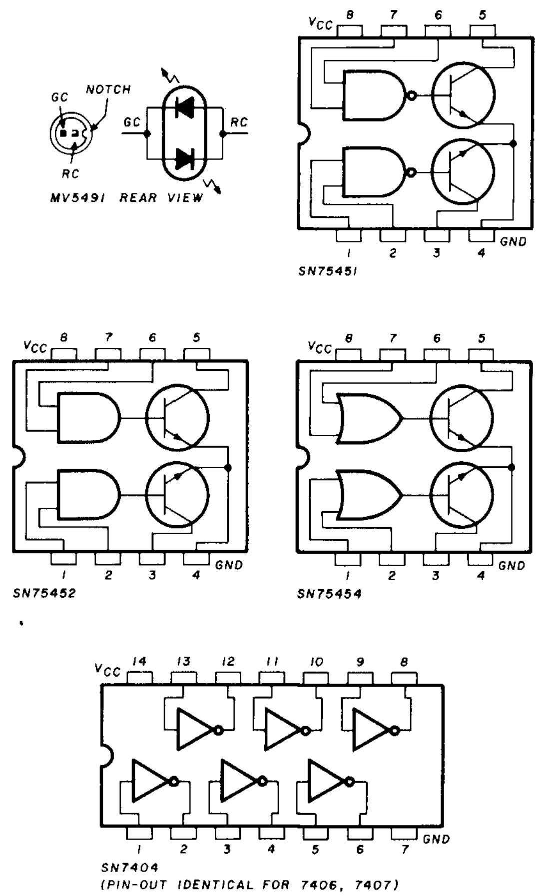

fig. 5. Device pin-outs for the integrated circuits.

and change the state of the LED from green to red. The vertical reset switch is then used to restart the vertical sweep and reset the LED to its green state.

Fig. 4F is a coincidence detector with three states. If inputs A and B are both high, the indication will be green. If inputs A and B are both low, the indication will be red. If the inputs are out of phase, one high and one low, the indicator will be in the off or third state. This would provide a good indicator for complex logic circuit monitoring.

Fig. 4G could be used as a temperature-monitoring device with the set points adjusted by a trimming resistor shunted across the thermistor. For photometric application, the thermistor could be replaced by a photocell. A comparator could be utilized before the TTL gate, for increased sensitivity.

I think this is just the beginning for applications of this unique device and I hope this article will stir the interest of other amateurs and further the interest in a return to more homebrew activity. I have enjoyed working with this device and the highlight was the look on a friend's face as the LED changed from red to green. I really had him going until I explained the device to him. This little LED can add the *bells and whistles* touch so often found in today's commercial equipment to your own homebrew efforts.

ham radio

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TS-520 TS-820

HF

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Features

Following are a few of the TS-820's many exciting features

SPEECH PROCESSER • An HF circuit provides quick time constant compression using a true RF compressor as opposed to an IF clipper. Amount of compression is adjustable to the desired level by a convenient front panel control.

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

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

TS-820 Specifications

FREQUENCY RANGE: 1.8-29.7 MHz (160 - 10 meters)
MODES: USB, LSB, CW, FSK
INPUT POWER: 200W PEP on SSB
160 W DC on CW
100 W DC on FSK
ANTENNA IMPEDANCE: 50-75 ohms, unbalanced
CARRIER SUPPRESSION: Better than 40 dB
SIDEBAND SUPPRESSION: Better than 50 dB
SPURIOUS RADIATION: Greater than -60 dB (Harmonics more than -40 dB)
RECEIVER SENSITIVITY: Better than 0.25uV
RECEIVER SELECTIVITY:
SSB 2.4 kHz (-6 dB)
4.4 kHz (-60 dB)
CW* 0.5 kHz (-6 dB)
1.8 kHz (-60 dB)
*(with optional CW filter installed)
IMAGE RATIO: 160-15 meters: Better than 60 dB
10 meters: Better than 50 dB
IF REJECTION: Better than 80 dB
POWER REQUIREMENTS: 120/220 VAC, 50/60 Hz, 13.8 VDC (with optional DS-1A DC-DC converter)
POWER CONSUMPTION: Transmit: 280 Watts
Receive: 26 Watts (heaters off)
DIMENSIONS: 13-1/8" W x 6" H x 13-3/16" D
WEIGHT: 35.2 lbs (16 kg)

DG 1, digital readout optional

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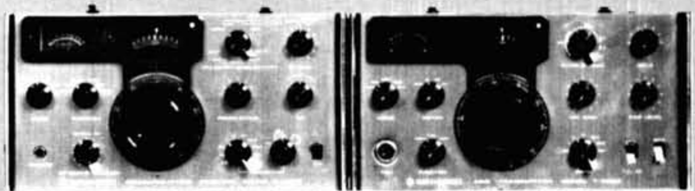
TV-506 TV-502 VFO-520 TS-520 SP-520

As a TS-520 owner, you go on the air with a sense of pride and confidence. Thousands of these precision-built beauties are in operation all over the world . . . in ham shacks, field day sites, in DX and contest stations and in countless mobile installations. No other rig has ever offered the performance, dependability, versatility and value that is built into every KENWOOD TS-520.

You have certainly heard the TS-520's clean signal on the air and have probably heard a lot of glowing praise by other hams. So if you don't already own a 520, maybe it's time you did.

MODES: USB, LSB, CW
POWER: 200 watts PEP input on SSB, 160 watts DC input on CW
ANTENNA IMPEDANCE: 50-75 Ohms, unbalanced
CARRIER SUPPRESSION: Better than −45 dB
UNWANTED SIDEBAND SUPPRESSION: Better than −40 dB
HARMONIC RADIATION: Better than −40 dB
AF RESPONSE: 400 to 2600 Hz (−6 dB)
AUDIO INPUT SENSITIVITY: 0.25μV for 10 dB (S+N)/N
SELECTIVITY: SSB 2.4 kHz (−6 dB), 4.4 kHz (−60 dB), CW 0.5 kHz (−6 dB), 1.5 kHz (−60 dB) (with accessory filter)
FREQUENCY STABILITY: 100 Hz per 30 minutes after warmup
IMAGE RATIO: Better than 50 dB
IF REJECTION: Better than 50 dB
TUBE & SEMICONDUCTOR COMPLEMENT: 3 tubes (2 x S-2001, 12BY7A), 1 IC, 18 FET, 44 transistors, 84 diodes
DIMENSIONS: 13.1" W x 5.9" H x 13.2" D
WEIGHT: 35.2 lbs.



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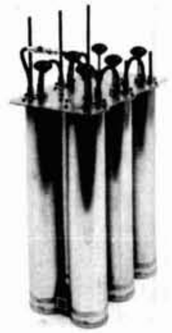

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More Details? CHECK—OFF Page 126

medical data relay

via Oscar satellite

A look at the techniques for the transmission of high-speed data through Oscar

The Oscar series of satellites have suddenly provided the amateur radio operator with a whole new field of expertise to learn and experiment with. Now we have reliable and widespread vhf and uhf coverage with minor atmospheric noise. Part of the justification behind the Oscar program is that we, as amateurs, use the satellites to explore and experiment with previously unavailable technologies. Many experiments have involved propagation and its related effects. RTTY, slow-scan TV, mobile operation, and even microcomputer access have all been tried. But, a largely unexplored area involved the use of the satellite to relay data unrelated to amateur radio. Achieving this result could have strong implications in the chosen field.

The practice of medicine today is absorbing new technologies as fast as they appear. This oftentimes carries a requirement that man and medicine be able to efficiently exchange information. It soon became apparent that the data transmission system I was developing at the University of Arizona Hospital could be adapted for use through Oscar.

It was originally proposed by 4X4MH to send an electrocardiogram (EKG) through Oscar 6. In the latter part of 1975 this was attempted by W6CG and W7VEW. Success was claimed but nobody knew the parameters necessary to decode the data. It had been an experiment involving "black boxes." At that point it became obvious that the handling of this kind of data was new to amateurs and new techniques had to be learned. I decided to stay with the EKG because it represents a commonly known and needed piece of physiological data. In addition, its bandwidth is wider than most other physiological signals. The ability to successfully relay an EKG would mean that most other types of medical data could be handled as well. Two techniques will be discussed and their merits examined.

To begin with, it's necessary to understand a little about the EKG itself. This is the controlling signal that tells your heart when and how to beat. To satisfactorily represent an EKG for most purposes requires a bandwidth of 0.5 to 50 Hz. Ideally, the lower response should extend to dc. In addition, this signal must be made compatible with amateur ssb transmitters and tape

By David Nelson, K7RGE, 4733 East Linden, Tucson, Arizona 85712

recorders if Oscar is to be used in the link. Remember that recorders and transmitters have a lower audio cutoff frequency that may vary from 50 to 300 Hz. Especially in a transmitter, the passband linearity of the audio circuits may vary a few dB. Obviously, the data must be encoded into a new format so that it can be transmitted through the system. Any form of a-m is clearly out. This then leaves fm, which has several desirable characteristics. The first is that amplitude variations due to nonlinear circuits, and a host of other possibilities, disappear. Amateurs experimenting with sstv discovered this in the late 1950s. Second, the use of fm permits the use of a tape recorder to serve as an interim storage device. This is necessary since the source of the EKG was anything but close to the transmitting site. Separate

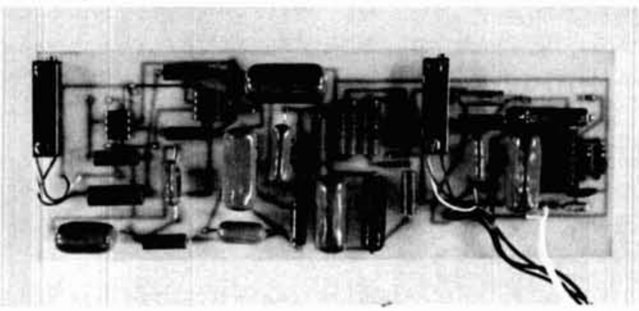

fig. 2. The completed demodulator assembled on a printed-circuit board. The layout of the board is not critical.

scheduling of the various phases of the experiment was then possible. However, the use of a tape recorder introduces some gremlins because the mechanical assembly that moves the tape can introduce errors that are difficult or impossible to eliminate. This and other considerations come into play when designing a system with 1 per cent accuracy.

analog fm

This approach is the easiest to understand and put to practical use, but it also has its shortcomings. The principle is to use a voltage-controlled oscillator (vco), with its input driven by the data source, as a source of fm. All that remains is to choose an appropriate carrier frequency and set limits on the deviation. At the receiving end, an fm demodulator will transform the varying frequency into a similarly varying voltage. If all goes well, this will provide a faithful reproduction of our data.

system problems

Since a tape recorder is being used, we must contend with a certain amount of instability in the speed of the tape drive. This speed fluctuation, known as flutter, will modulate the material recorded on the tape and add undesirable components. Rerecording on different machines adds to the problem. If the basic drive speed of the two is slightly different, a dc offset will be introduced. Unless we have a grounded second fm channel available for reference, and the ability to simultaneously transmit the two, the degradation caused by flutter cannot be overcome; it can only be minimized by using quality tape machines and maximizing the deviation of the vco.

It is appropriate now to look ahead and anticipate that Doppler shift and receiver tuning errors will be present. As such, it is unwise to make the vco deviation equal to the maximum that the receiving demodulator can accommodate as there would be no room left for error. With all these factors in mind, I settled on a 1-kHz carrier with ±40 per cent deviation for full-scale input to the vco. The vco I used was a Hewlett-Packard 3310A function generator set to 1 kHz and modulated by the EKG source.

Another problem is noise and here we can only maximize the received signal-to-noise ratio. However, this will not completely delete the effect of noise introduced by the satellite's transponder and atmospherics, especially on 10 meters. Noise causes a directly visible reduction in the resolution of the data.

The receiving demodulator is worthy of a little more attention. In **fig. 1**, you can see that the audio signal is fed to the input of a 565 phase locked loop. The error voltage of the loop (pin 7) contains the data being

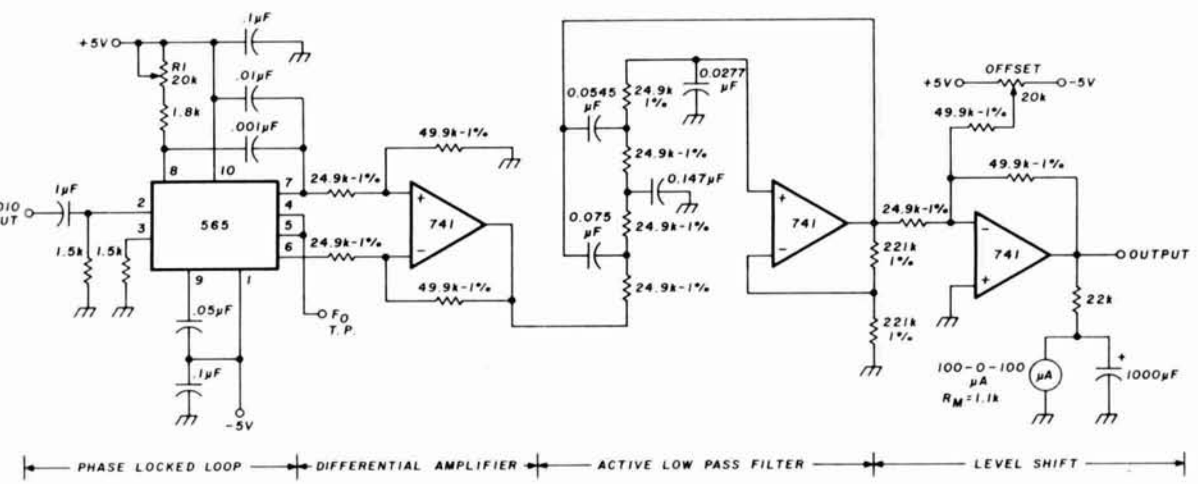

fig. 1. Schematic diagram of the analog fm demodulator. The frequency of the phase-locked loop can be changed by adjusting R1.

sought, along with the undesirable dc and ac components. Following the loop is a differential amplifier. Its purpose is to remove the large intrinsic dc component of the error signal and amplify the rest by a factor of two. Following the amplifier is a four-pole, active-RC, low-pass filter, also with a gain of two. This filter serves to eliminate the high-frequency ac components that are a product of the loop's internal detection process. It also very accurately determines the bandwidth of the demodulator. In this case, the Butterworth filter is scaled for a cutoff frequency of 100 Hz. This response is flat in the passband, down by −3 dB at the cutoff frequency, and rolls off at −6 dB per octave per pole or −24 dB per octave for this filter. The oddball capacitor values are a consequence of an accurate filter design. Labeled capacitor values should not be trusted. I used a pair of 10 per cent capacitors (no ceramics) for each of the four and measured each pair on a bridge to obtain values within one per cent. Lesser accuracy can be tolerated for less accurate results. Following the filter is an amplifier to scale and shift the output to a reasonable value.

demodulator alignment

To align the demodulator, apply ±5 volts and set the potentiometer to give a loop frequency of 1 kHz at pins 4 and 5. The input should be grounded at this time. Then the potentiometer on the output amplifier is adjusted to give zero output voltage. If the device is now working properly, the loop should lock onto and track any input signal that lies within ±50 to 60 per cent of 1 kHz. The input level can range from 20 mV to 2 volts rms. The output will swing positive for input frequencies above 1 kHz and negative for inputs below 1 kHz. The output-scale factor is approximately 0.5 volt per 10 per cent deviation and is independent of the loop frequency. Note that only one adjustment must be changed to accommodate different carrier frequencies. **Fig. 2** shows a photograph of the demodulator.

The 1 per cent resistors used here serve more to guarantee close resistance ratios than absolute values. Readers familiar with op-amp design will note that most of the values in the individual sections can be changed, except for the filter ladder, as long as the resistance ratios are accurately maintained. This ability lends itself to the use of available units rather than pursuing absolute values.

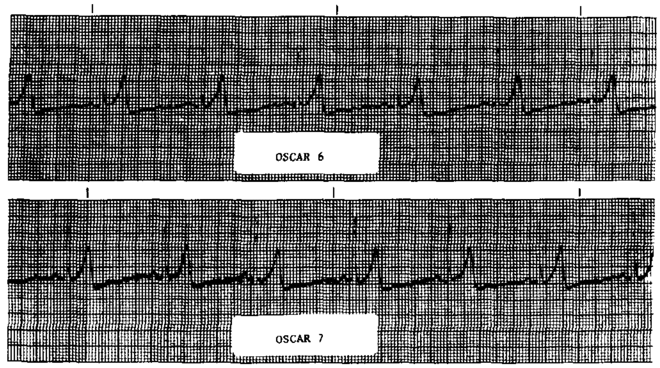

fig. 3. Analog electrocardiograms that have been transmitted through Oscar 6 and 7. Note the less accurate response of the Oscar 7 Mode A version.

This demodulator can also be made to serve as a tuning indicator during the test transmission if you have some prior knowledge of the data. The EKG waveform stays mostly near zero and deviates rapidly when it does so. If a zero-center microammeter is bypassed with a capacitor to provide a one to two second time-constant with the meter resistance, the meter will respond to very slow changes such as Doppler shift but will show little response to the faster EKG data. This yields an indicator that tells how to correct the receiver tuning to offset the Doppler shift. All you have to do is keep the meter centered during the transmission by adjusting the receiver tuning. Incorrect tuning will deflect the meter to one side or the other. This method is considerably more accurate than tuning by ear because changes of 10 per cent are quite visible.

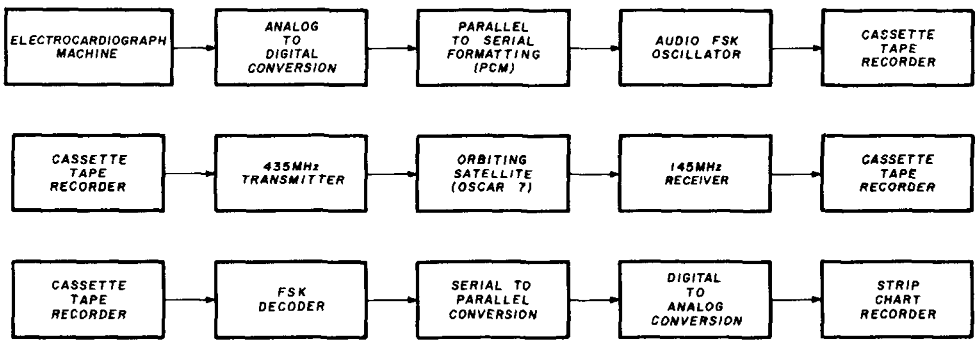

fig. 4. A flow chart for the PCM type system. This method allows the process to be done at different times and locations.

Another method of correcting for Doppler involves the use of a second non-modulated audio carrier sufficiently removed from the first so it doesn't interact. The tuning indicator is locked onto this new carrier. However, this may exceed the available audio bandwidth and requires bandpass filters.

In September, 1975, I conducted a test of this

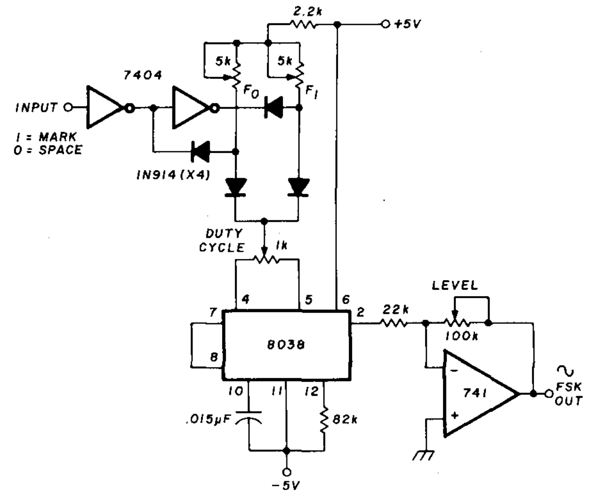

fig. 5. The fsk oscillator. The separate timing resistors set the mark and space frequencies. A sinusoidal waveform is synthesized internally.

method with W6ELT transmitting a cassette tape through Oscar 6 and Oscar 7, mode A. Shortly thereafter we tried another transmission to both myself and the radio club at the National Institute of Health in Bethesda, Maryland. The results are shown in **fig. 3.** Note the difference in the results between Oscar 6 and 7. This is due to the stronger signal available from Oscar 6. Although perfectly copiable signals were present, there is still a residual amount of noise present. This is best seen in the baseline of the EKG. In addition, even though the Doppler was manually compensated for, this method would have been useless if the data had been of a completely unknown form. Obviously, there is room for improvement in many areas.

going digital

The quickest way to solve many of the problems is to go to a digital method of encoding data. Not only can better resolution be achieved, but the effects of noise, Doppler, and tape flutter can be simultaneously eliminated. However, you must pay the price in terms of system complexity and cost. But if truly accurate, reproducible results are required, this is the way to go. This method is commonly known as pulse code modulation (pcm).

The principles are to convert the data into a digital word and then handle the bits serially, one at a time. Unlike analog techniques, digital signals can be regenerated in the presence of noise. The resolution of the system is determined by the number of bits that represents the data. I chose eight bits which gives a resolution of 0.4 per cent of full scale. This method of handling the data contains a number of parallels to RTTY and some of the techniques can be directly applied. The conversion from analog to digital is accomplished by a commercial module appropriately called an analog-to-digital (A/D) converter. For the purposes of this discussion it is not necessary to understand the inner workings of this and other modules used. It is only necessary to understand their function.

Following the conversion to digital, the parallel eight-bit word presented by the A/D converter is transformed into a serial bit stream by a rather amazing IC called a universal asynchronous receiver transmitter, or UAR/T. This IC consists of two halves. The transmitter

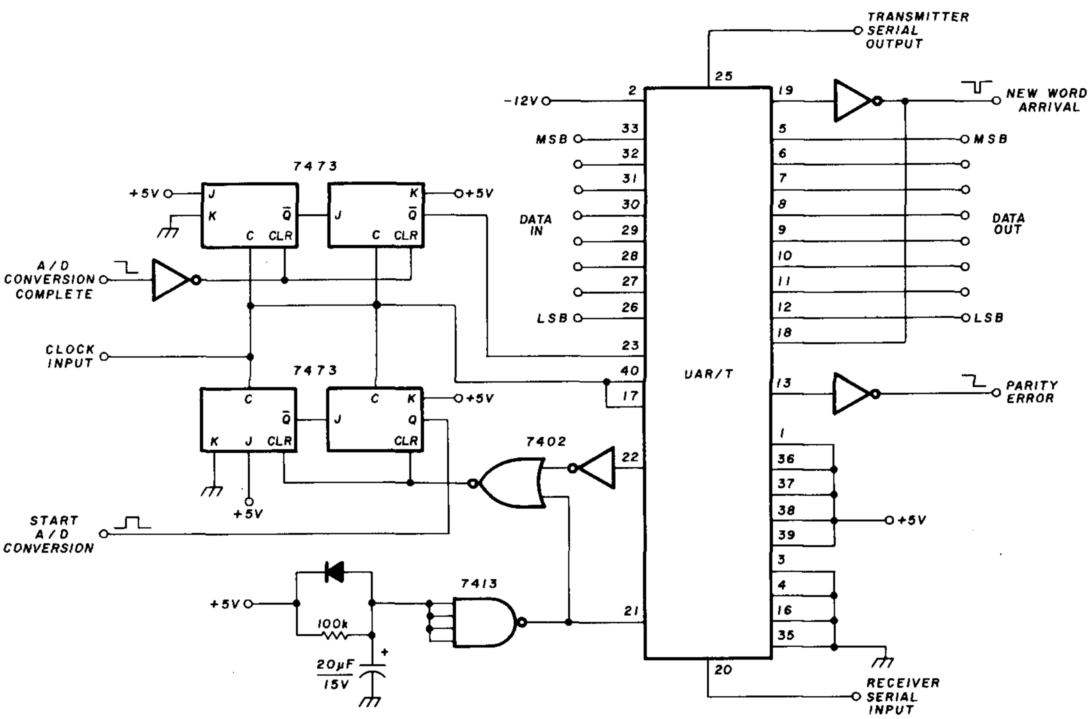

fig. 6. Schematic diagram of the UAR/T data conversion system. This method performs all conversions between serial and parallel data.

half takes a parallel input and formats it into a serial asynchronous code with start, stop, and parity bits added to the data. The receiving half does just the opposite. It accepts a serial input, strips off the bookkeeping bits, checks for errors, and outputs the digital word in parallel. Controlling all of these operations is an external clock which determines how fast the system runs. Except for the clock, all of this is nicely packaged into a single 40-pin IC.

The serial-bit stream from the UAR/T then controls an fsk oscillator. This oscillator is switched between two discrete audio frequencies and is an audio representation of the digital signal. These audio signals are then placed on a cassette tape to be transmitted later through the satellite. A block diagram of the system is shown in **fig. 4**. Each of the three lines takes place at a different time due to equipment locations and schedules.

The fsk generator is fairly straightforward and is shown in **fig. 5**. Basically an 8038 IC function generator is switched between two adjustable trimmer resistors giving independently adjustable *mark* and *space* frequencies. The output is phase coherent although switching does not necessarily take place at the zero crossing points of the sine wave.

UAR/T description

As mentioned before, it is the UAR/T's job to perform the parallel-to-serial and the serial-to-parallel conversions. This IC does each job independently, simultaneously, and at different rates if desired. In my case this last capability was not necessary. As shown in **fig. 6**, the IC is programmed for an eight bit word, two stop bits, and even parity by setting pins 36-39 permanently high. The parity bit is included for error-detecting purposes and is used later to give an indication of the validity of the received data. The input data word is applied, in parallel, to pins 26 to 33 by the A/D converter and a pulse is generated by the flip-flops which loads this word into the UAR/T's internal register. The UAR/T begins clocking out this word serially immediately. As soon as the UAR/T is ready to receive another word, it signals by setting pin 22 high. This creates a pulse which triggers the A/D converter. Double buffering allows the UAR/T to be sending one word while the A/D converter is busy loading a new word. Thus the rate at which information can be sent is determined by the rate at which the UAR/T can send out the bits.

Sampling theorem states that at least two samples per second must be taken per hertz of data bandwidth. Thus, in order to represent an EKG which contains components up to 50 Hz, the analog EKG data must be converted to digital at least 100 times a second. Since each conversion produces an eight-bit word and the UAR/T adds four bits of bookkeeping data per word, we must transmit the bits at the rate of 1200 per second to process the EKG. It would be nice to go faster, but bandwidth requirements become more restrictive when it comes to the fsk; to send 100 data points per second, the UAR/T must be clocked at 16 times the baud rate (1200 baud). Thus a 19.2 kHz clock is required. This clock should be crystal derived and applied to both the transmit and receive halves of the chip (pins 40 and 17).

The receiving portion is equally as simple. The serial bit stream is applied to pin 20. The received word is presented on pins 5-12 and pin 19 goes high to signal its arrival. This control signal is also applied to the UAR/T through pin 18 to signal the removal of the data. When pin 18 is set low, it resets pin 19 low. However, this does not happen instantly and internal propagation delays turn the pin 19 signal into a short low-going strobe. If needed, this may be used to signal the arrival of a new data word.

The rest of the pins are control functions which should be wired as shown. Pin 21 must be strobed high when power is first applied to clear all internal registers and ready the chip for operation. Anyone desiring to build a modern RTTY system using surplus keyboards will find the UAR/T very handy indeed. Several manu-

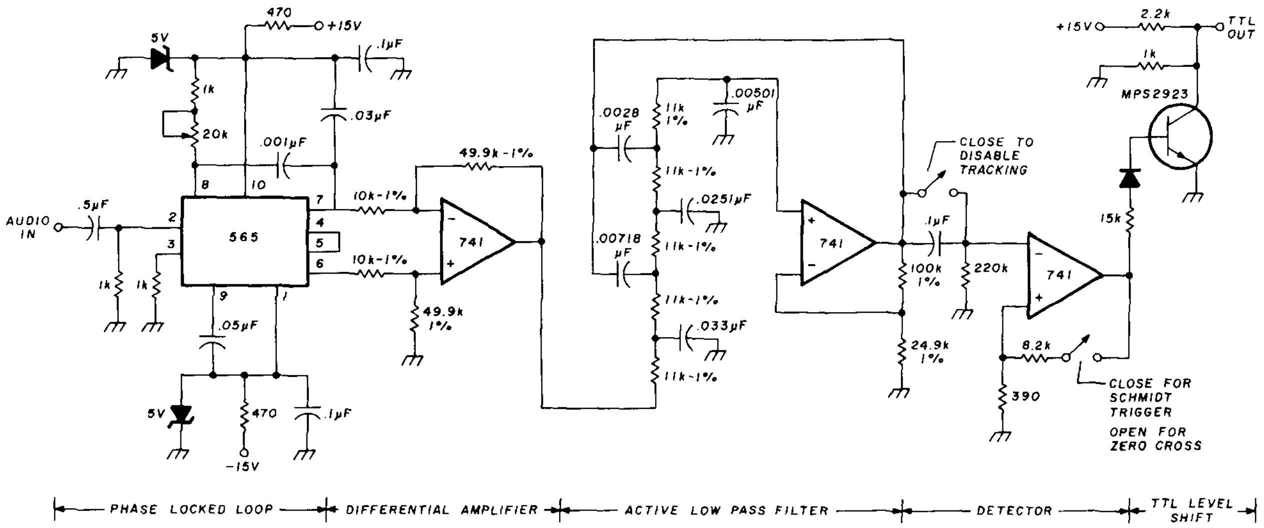

fig. 7. The schematic diagram for the fsk demodulator. An active RC filter allows reception of the 1200-baud data.

facturers, including General Instruments, Texas Instruments, American Microsystems, and National Semiconductor, produce compatible ICs. The National MM5303N was used in the system I built.

digital decoding

Receiving and decoding the fsk signal involves the same process as receiving RTTY. The only differences are that the system is running much faster and that the keying is a change in audio frequency presented to the transmitter rather than a change in its vfo. **Fig. 7** shows the diagram of a suitable fsk demodulator. Note the lack of bandpass filters on the input. Since these tests were done on Oscar's experimental orbits, a clear channel was available. This made it easy to control Doppler shift. Once again, a phase-locked loop tracks the input signal frequency and provides an appropriate error signal. Following the differential amplifier, the error signal is applied to a five-pole, Butterworth lowpass filter which has a cutoff frequency at 1500 Hz. The output of the filter resembles a digital signal but shifts up and down from zero depending on Doppler and the receiver tuning. Therefore, a zero-crossing detector is useless at this point. The easiest way to remove the dc offset is by capacitor coupling. The time constant associated with this capacitor is chosen to block slow changes such as Doppler and slow receiver tuning changes, but to pass the faster incoming data. Now, you can use either a zero-crossing detector or preferably, a Schmidt trigger.

The next stage can serve as either type of detector. By adjusting the feedback resistance on the op amp, the hysteresis can be tailored to the amount desired. More resistance results in less hysteresis. This method of detection allows any combination of *mark-space* frequencies

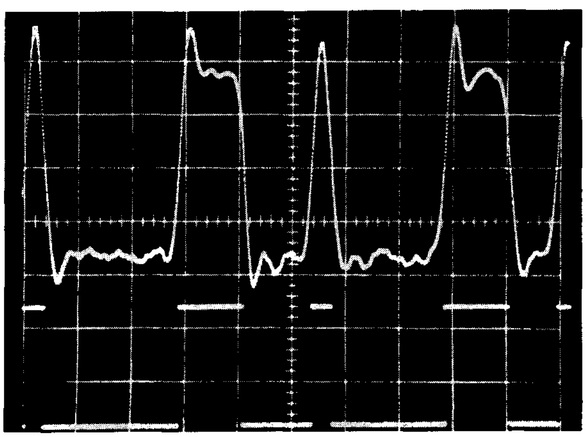

fig. 8. Oscilloscope traces of the detector input (top) and the TTL output (bottom). The time base is 2 ms per division.

to be used. As long as these frequencies do not drift too near the passband of the lowpass filter, any shift that exceeds the effective modulation introduced by the noise will produce a waveform that the detector can convert into a usable digital signal. The fact that these frequencies may drift is of no consequence. If, for some reason, it is desirable to disable this tracking feature, simply short the coupling capacitor in the detector as shown.

This signal is next converted into a TTL compatible level. The net result is a decoder that does not care what

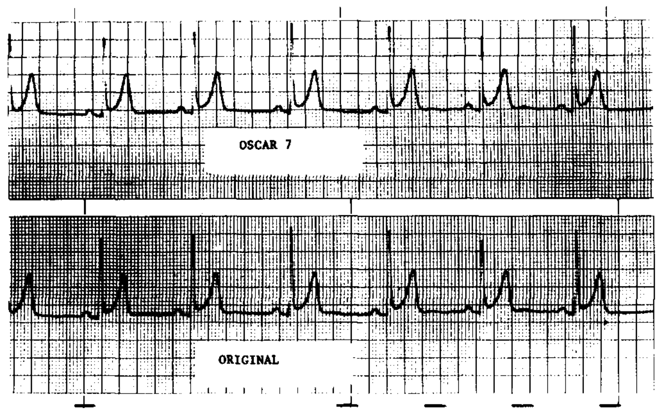

fig. 9. Samples of the digitized EKG data as transmitted through Oscar 7 Mode B.

the *mark-space* frequencies are, what their deviation is, and is not affected by drift, except by going too far down. On a test bench this circuit reliably decodes any shift above 50 Hz, with an input frequency of at least 1900 Hz and at any speed to 1200 baud.

Fig. 8 shows a dual-trace presentation of data as received through Oscar. The upper display is the input to the detector and the lower display is the digital output. The vertical scale represents shifts of 250 Hz per division; 850-Hz shift being used at this time. The waveform represents part of a digital word being received; note the noise-induced ripple. Any deviation exceeding this margin gives good data. This display shows how much deviation is needed for the noise level present. My first test used 170-Hz shift which proved to be insufficient as the noise margin was over 200 Hz. The next test, using 850-Hz shift succeeded. Subsequent tests are being conducted to determine the minimum necessary shift. Obviously the bandwidth presented to the phase-locked loop depends on the receiver and there is noise contribution from every bit of it. In my case the receiver was a Drake 2B with the selectivity set to 3.6 kHz.

Received signal strength also has a great deal of influence with respect to the noise. The use of bandpass filters on the input of the decoder could drastically cut this noise bandwidth but that would require the receiver operator to exactly track the Doppler shift by hand. In addition, the group-delay distortion of any such filters must be closely controlled since at 1200 baud you only have a couple of cycles of one frequency present assuming a 2400 Hz *mark*. To my way of thinking, it was the job of the decoder to eliminate the Doppler and not mine, so I just left the bandwidth wide open and used a wider shift. Only occasional retuning was necessary during the pass to compensate for the Doppler.

Following this decoder, the data goes to the receiving half of the UAR/T where the serial-to-parallel transformation takes place. The parallel digital word presented at its output goes to a digital-to-analog converter which feeds a fourpole, lowpass filter set to 50 Hz. The output of this filter is the reconstructed analog data. **Fig. 9** shows the result of the tests. The bottom strip is the original data and the top strip shows the same data after passing through Oscar 7 mode B. Out of 60 seconds of data, 5 parity errors were found. Although the same EKG was not used for both techniques, the improvement is obvious. Note the lack of degradation due to Doppler shift and noise. Doctors who examined the strips were unable to distinguish between the original and the satellite relayed versions.

While not optimum, this technique demonstrates what can be done with an ordinary audio channel. For more permanent experimentation I would recommend that you replace the 8038 fsk generator with a crystal-controlled programmable system, using digital counters, and synthesize a sine wave from a BCD-to-decimal decoder such as the 7442. This would give absolutely stable and accurate *mark-space* frequencies. Some input bandpass filters could be added to the demodulator. This system could also be adapted to multichannel use. It would require splitting the data word into two parts and sending each half, along with its appropriate address, as one word through the UAR/T. A substantially higher bit rate would then be required.

summing up

So what has been the benefit of all this? Most obvious is that a satellite may be used to relay high-quality physiological data. This has definite use for regions where traditional land-based communications are marginal or non-existent. Small Alaskan towns are a prime example. Currently the federal government is involved in projects of exactly this nature. For amateurs this marks the first time that the FCC has granted permission to use an eight-level teleprinter code. At the time these experiments were being formulated, such permission did not exist and AMSAT was kind enough to pass my request along to the FCC. This opens the field for the use of surplus ASCII terminals and I anticipate that we will begin to see video terminals interconnected via amateur radio. As the cost of television typewriters continues to drop, this will make it easy for anyone to implement a low-cost RTTY station that can run at 25 characters per second. Although this experiment used an EKG, it could just as easily been text coded into ASCII and sent at 100 characters/second. The ability to handle bulky information rapidly can benefit users such as RTTY traffic nets, W1AW bulletins, and satellite telemetry of internal operations. The list of possibilities is limited only by one's imagination.

I need to acknowledge the help given me by Amsat in relaying my request for authority to use an eight-level code to the FCC, and to W6CG and W6ELT who helped control the satellite and transmit the tapes through it. I would be happy to answer anyone's questions upon receipt of a self-addressed, stamped envelope.

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the solution

Being a classical music fan and having a deep interest in audio, I came up with a technique that should attract any phone man. The audio system in many receivers leaves a lot to be desired. Exceptions to this were Karl Pierson and his KP-81 with high and lowpass audio filters. The Hallicrafters SX88 also had selectable audio response curves. Both had 10 watts of clean audio.

What I've done on all my receivers (except the KP-81 and SX88) is add a 20-watt hi-fi amplifier. Connected to the output of your receiver's detector (**fig. 1**), you can select either the normal receiver audio, the outboard amplifier, or both. Further, the use of a good speaker, preferably one of the good high-efficiency hi-fi speakers, means you don't have to increase the volume as much. A large speaker moves a lot of air and it "fills" the room. I often run upstairs and can hear the bigger speaker better upstairs as compared to a smaller one.

audio emphasis

The ultimate in audio control is to use an audio amplifier with a Sound Effects Amplifier (SEA), as used by JVC in their audio systems. I find it very helpful in shaping the audio characteristics of various signals under different interference and voice conditions.

Radio Shack has an excellent SEA unit called a Stereo Frequency Equalizer that permits you to emphasize or attenuate any part of the audio response curve by 10 dB. The change in voice characteristics is useful under various receiving conditions and will also

By Ken Judge Glanzer, K7GCO, 202 South 124th Street, Seattle, Washington 98168

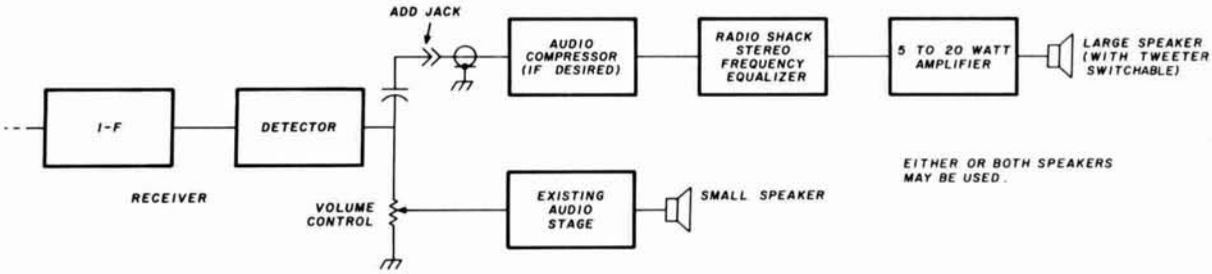


fig. 1. The large speaker may be connected to the receiver through other pieces of audio processing equipment. This allows you to adjust for the best audio reception under various receiving conditions.

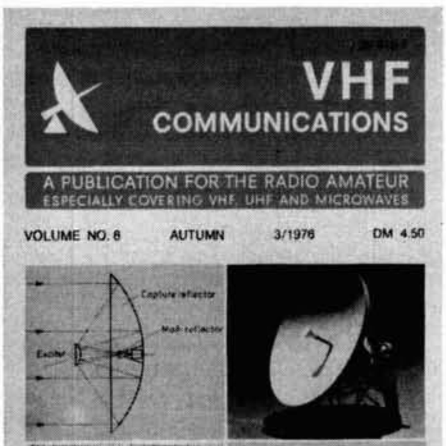
The Stereo Frequency Equalizer can be used to shape the audio output of most receivers. They will provide attenuation or emphasis of specific frequency segments.

act as an audio filter for CW. An equalizer could be used between the microphone and transmitter to provide the best audio for rag chewing or DXing. One idea is to record your normal and altered voice and then listen to the tape directly, not over the air.

circuit modifications

Minor modifications can be made to many receivers by removing capacitors that are across the collector to base junction of transistors in the audio amplifier stage. In some cases there may be a capacitor across the primary of the output transformer. These capacitors usually attenuate the high frequencies. Typically the first loss in hearing occurs at the higher frequencies. This can be relieved by increasing the high frequency response. Sometimes the substitution of a different tube, transistor, or output transformer will also help to alleviate the problem.

ham radio



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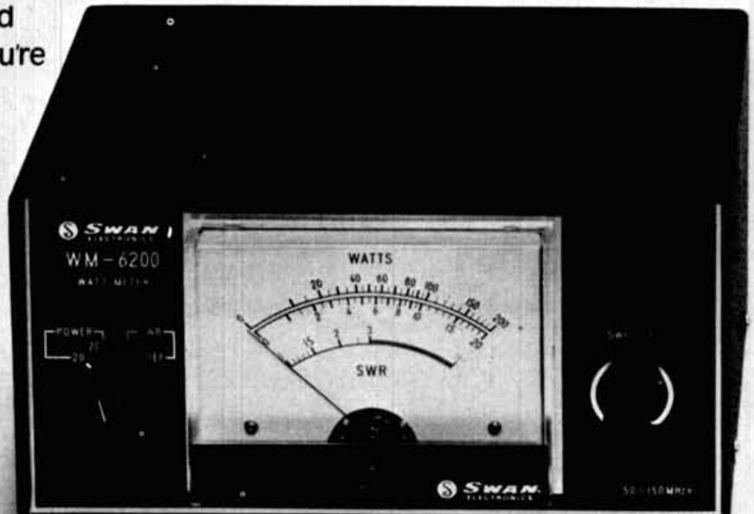

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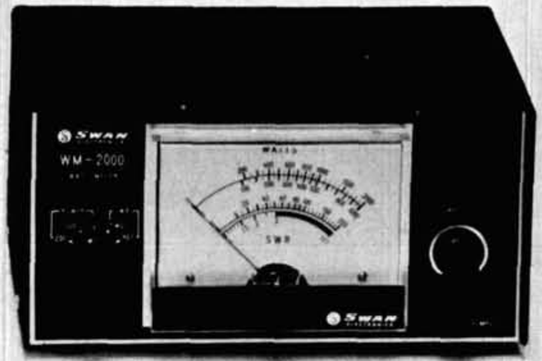

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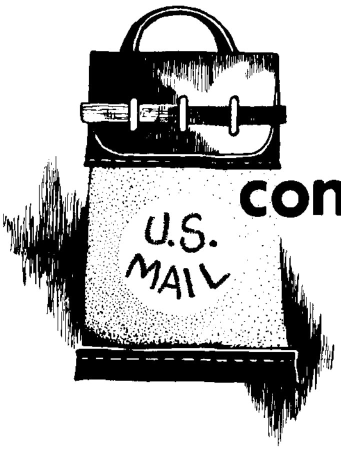


comments

RC notch filter

Dear HR:

A subscriber of your magazine recently showed me the article on "Hall's" notch filter in the September, 1975, issue* and I was very pleased with your enthusiastic presentation of the circuit. Obviously, I too feel that it has not had the popularity it deserves! Since you solicited comments on it and similar circuits, I would like to make a few comments which may be of some interest.

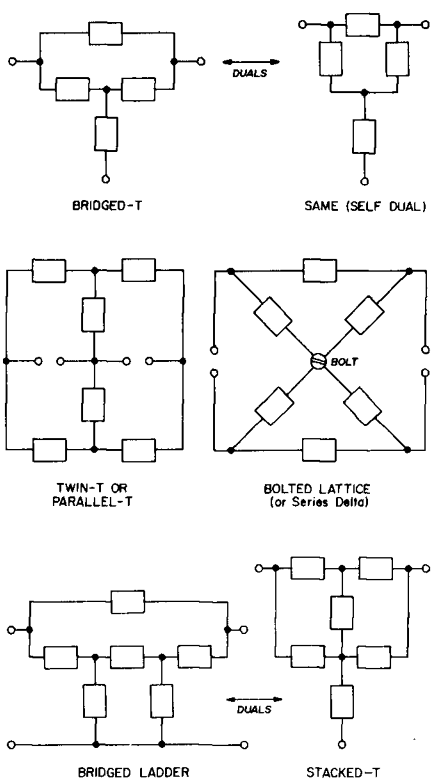


fig. 1. Popular ac null circuits (left) and their duals (right). The bridged-ladder was the circuit discussed in WA5SNZ's article in the September, 1975 issue of the magazine.

First, just for the record, the circuit has had some mention. The most complete description was by Leon Grillet *(91^c Congress des Societes Savantes*, Rennes, 1966, Tome II) which is in French, except for the algebra. You mentioned the article by Glasgal who referenced another *EEE* article by B.M. Van Emden (May, 1964). It has also appeared in a book, *Alternating Current Bridge Methods* (Hague and Foord, Pitman, London). This is Foord's revision of Hague's classic, the bible of ac bridge designers (my speciality). It was also briefly described by Penn and Grillet in letters (see *Electronic Engineering*, 1964). Others have mentioned it but only as introductory references before they described new circuits.

The name "bridged-differentiator" describes more than the network's typology which I've called a "bridged-ladder." It should be noted that the term "twin-T" has been used to describe measurement circuits with the same typology as the notch filter but with different circuit elements. Also, the dual† of a twin-T has been called a bolted-lattice (by an Englishman, I think) but could be called a "series-pi." The dual of the bridged-ladder might be called a "stacked—T" (see **fig. 1**).

The network has its advantages and disadvantages. The advantage in the General Radio null detector is not so much its economy on one potentiometer (which has a 46 dB exponential taper to get a log scale and is not inexpensive to make), but in its behavior. The obvious choice would have been a two-potentiometer Wien bridge, but when that circuit is adjusted in a high-gain feedback circuit, tracking unbalance as the wiper jumps from one wire to the next causes gain variations which make an indicating meter fluctuate widely. With one potentiometer, the gain may change in discrete steps as successive wires are contacted, but at

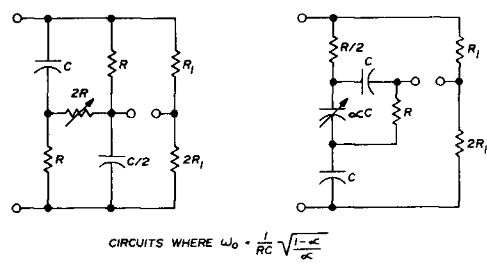


fig. 2. Bridge circuits with a tuning law which tunes from zero to infinity. Over a 10:1 frequency range, a linear variable component provides a good logarithmic dial.

least the peak is approached monotonically.

The main disadvantage of this circuit is that its Q is lower than that of a twin-T, like all null circuits adjusted by one component, Q varies over the

*Courtney Hall, WA5SNZ, "Tunable RC Notch Filter," *ham radio*, September, 1975, page 16.

†Dual networks are defined as a pair of networks such that their branches can be marked in a one-to-one correspondence so that any mesh (or loop) of one corresponds to a cut-set of the other.

tuning range. The null detector uses the transfer impedance to obtain a more constant Q. The result is not ideal but surprisingly good. As I remember it, we were about to abandon the network when we discovered this way of using it.

There are other circuits to be considered, particularly the one by Y.A. Andreyev mentioned in the *G-R Experimenter* article (and by Grillett and Penn). Its tuning law seems preferable. There are many bridge-type networks which use a single variable resistor or capacitor. My favorites are those in the *G-R Experimenter* article that have a $\sqrt{(1-\alpha)/\alpha}$ tuning law which tunes from zero to infinity (see **fig. 2**). Over a 10 to 1 frequency range, a linear variable component gives a pretty good logarithmic dial. I've always wanted to experiment more with these circuits and maybe your readers would have fun making oscillators with them.

Finally, I would like to note that I'm very glad to have been made aware of your magazine While I'm not an amateur, the few issues of *ham radio* I glanced through contained several articles that interested me. I plan to look through more and follow it in the future.

Henry P. Hall
Senior Principal Engineer
Gen Rad

50-MHz frequency counter

Dear HR:

I am writing concerning WB2DFA's article on the "50-MHz Frequency Counter" in the January, 1976, issue of *ham radio*. The schematic for the counter circuit contains three errors in the crystal-oscillator section: the inputs to the SN7400 NAND gates are left floating. For the oscillator to work properly there must be a definite *high* on the unconnected input pins. Admittedly, the oscillator may work if the inputs drift high because of internal transistor action, but for reliable and stable operation there should be a *high* on the unconnected inputs to the NAND gates. This problem may be corrected in one of two ways:

1. Connect input pin 2 of U1A to pin 1, pin 4 of U1B to pin 5, and pin 10 of U1C to pin 9; or

2. Supply $+V_{cc}$ to the unconnected inputs through 2200- to 6800-ohm resistors (any resistor between these two values will work).

Both methods make the NAND gates function as inverters but method 2 decreases the effective fan-in of the chip so it presents less of a circuit load.

This should solve any problems readers may have with intermittent or malfunctioning oscillators in the frequency counter.

James R. Aiello
JRA Systems
St. Clair Shores, Michigan

Although the inputs to the NAND gates should be pulled high rather than being allowed to float for true digital operation, we are not concerned with noise immunity in this case. In fact, it's less noise immunity that kicks the oscillator into action in the first place. Furthermore, tying both input pins together means that twice as much sinking current will be required to pull those inputs down, thus making the oscillator harder to start.

WB2WUF of Port Monmouth, New Jersey, built the counter almost a year ago, and in a recent conversation with him he reported that everything is still working okay. I have never had any problem with oscillator start-up either, and suggest that the circuit be left as is.

James W. Pollock, WB2DFA

telephone system precautions

Dear HR:

The telephone companies have rules against interconnects for a very good reason: they want the system to go on *working*.

In a recent issue of *ham radio** you published a schematic which showed a tap on a phone line which was directly grounded to the equipment. No doubt the author got away with it because he did not ground his equipment to the real world.

Although one wire of the phone pair into your house may appear to be at ground potential when first checked with a voltmeter, this is *not* a ground wire and the system must be isolated from ground to work properly. The phone will have all kinds of buzz and crosstalk if one lead is even bypassed to ground because phone pairs are *balanced* audio circuits.

A typical input to the switchgear in the phone office looks like this:

The transformer picks off the audio in a fully balanced manner, with the capacitor providing the tie at the center for audio. Another winding on the transformer passes the audio on into the system.

The relay detects the presence of dc current in the phone (off-hook and dial pulses), again in a balanced manner with two coils. Adding another ground at the user's end shorts one winding of each.

From there on, things get quite complicated because the relay is not always the same relay. A different physical relay may be in the circuit for *each digit* of the dialed number, with the dc provided during the talking phase being many steps removed (and perhaps many miles away) from the relay monitoring the line when on-hook. A ground on one lead causes great confusion in all this.

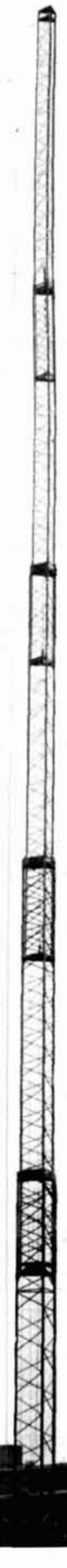
If you must attach to the system, there is a simple method of keeping out of trouble. *Always* present to the phone line a *transformer winding with a blocking capacitor in series.* The winding impedance can be anywhere from 50 to 2000 ohms. The capacitor should be no larger than 0.2 μF and able to stand 90 Vac ringing voltage with 48 Vdc superimposed (i.e. at least a 200 V paper capacitor).

On the other side of this transformer, you can do pretty much what you like as long as you don't drive audio into it at such levels that things are louder than normal speech in the normal telephone.

For a simple ring-detector, as was desired in the *ham radio* article, an optical coupler works admirably. Place a 0.1 μF in series with it to block dc from the phone system and put a diode across the optical coupler to pass the reversed part of the ringing cycle.

N. J. Thompson, KH6FOX
Honolulu, Hawaii

*Robert Shriner, WA0UZO, "Automatic Telephone Controller for Your Repeater," *ham radio*, November, 1974, page 44.



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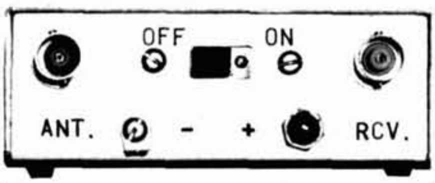

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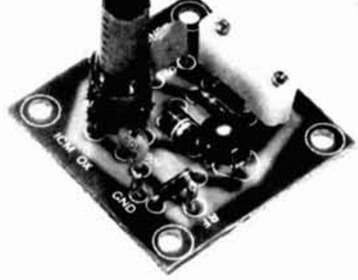

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Crystal controlled transistor type. 3 to 20 MHz, OX-Lo, Cat. No. 035100. 20 to 60 MHz, OX-Hi, Cat. No. 035101
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$3.95 ea.

MXX-1 TRANSISTOR RF MIXER
A single tuned circuit intended for signal conversion in the 30 to 170 MHz range. Harmonics of the OX or OF-1 oscillator are used for injection in the 60 to 179 MHz range. 3 to 20 MHz, Lo Kit, Cat. No. 035105. 20 to 170 MHz, Hi Kit, Cat. No. 035106
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PAX-1 TRANSISTOR RF POWER AMP
A single tuned output amplifier designed to follow the OX or OF-1 oscillator. Outputs up to 200 mw, depending on frequency and voltage. Amplifier can be amplitude modulated. 3 to 30 MHz, Cat. No. 035104
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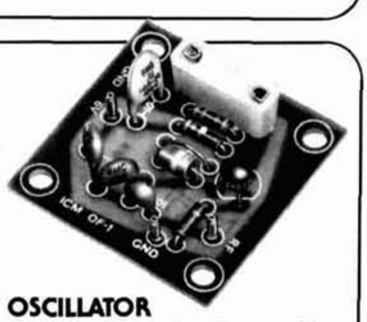

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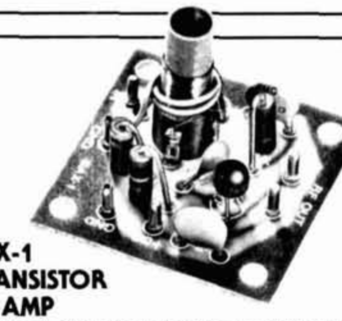

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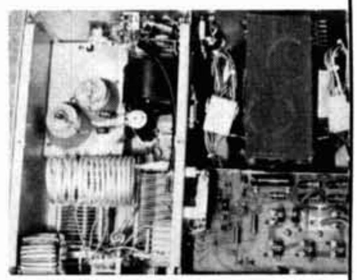

ALPHA 76 is first of all power — brute, uncompromising power. It easily handles up to 2.5 KW PEP or more — a full 1000 watts of average d-c input — with no need for a key-down time limit! That's *true* 'continuous duty' [or 'CCS' — Continuous Commercial Service]. No other linear amplifier, except another **ALPHA,** delivers such power capability in a self-contained desk top cabinet.

ALPHA 76 is also the only true [CCS] full KW linear that includes 160 meter amateur band coverage and costs under $2000!

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ETO's dedication to your convenience shows up in features like the plug-in transformer for easy handling and UPS shipping; the remarkably compact one-cubic-foot size for a linear of this power capability; full amateur band coverage including 160 meters; and the built in rf directional wattmeter.

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FREQUENCY COVERAGE: Tunable 1.8-2.0 and 3-30 MHz; all bands 160-10 meters.

PLATE POWER INPUT: Up to 2.5 KW PEP-SSB, 1 KW average or key down; no time limit.

RF POWER OUTPUT: Typically 1.2-1.5 KW PEP-SSB voice, 600 watts continuous carrier.

INPUT & OUTPUT IMPEDANCES: 50 ohms nominal, unbalanced; VSWR 2:1 or less.

DRIVE POWER: Nominal 100 watts PEP, 50 watts carrier; compatible with all popular exciters and transceivers.

DISTORTION: Third order IM more than 30 db below 1 KW PEP output.

ALC: Negative-going, adjustable threshold.

TUBE COMPLEMENT: Two Eimac 8874 ceramic-metal grounded grid triodes.

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PRIMARY POWER: 240V/10A or 120V/20A nominal, 50-60 Hz, 1Ø. Automatic step-start.

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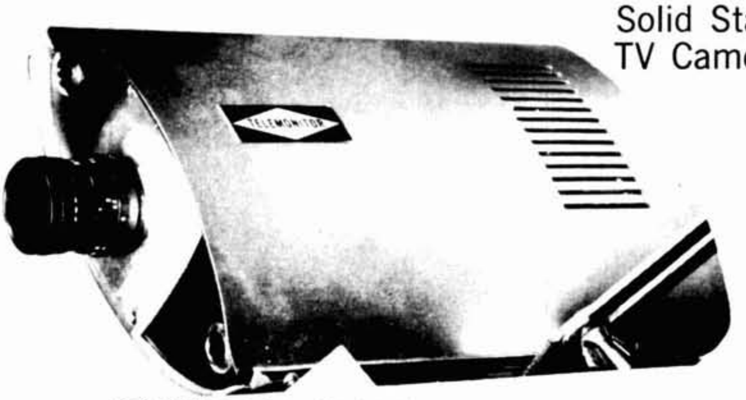

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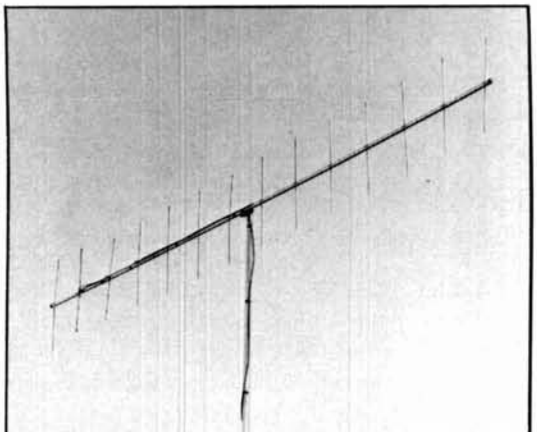

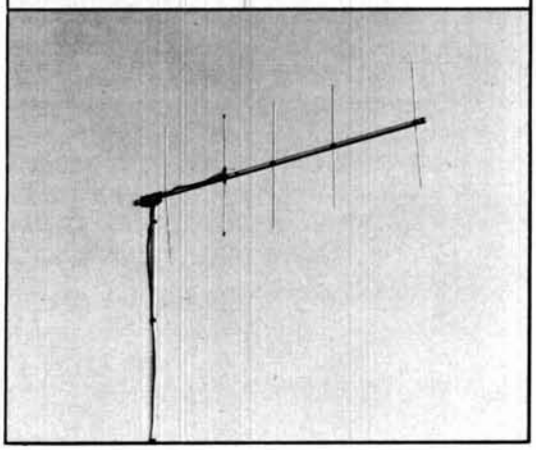

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SPECIFICATIONS	**214**	**208**	**205**	**203**
Mechanical				
Boom length	186″	148 ¾″	75″	43½″
Longest element	39½″	40¼″	39 ⅝″	40¼″
Turning radius	95″	75 ⅛″	73″	43½″
Wind survival	80 mph	80 mph	80 mph	80 mph
Mast diameter	1¼-1⅝″ O.D.	1¼-1⅝″ O.D.	1¼-1⅝″ O.D.	1¼-1⅝″ O.D.
Boom diameter	1¼″ O.D.	1¼″ O.D.	1¼″ O.D.	1¼″ O.D.
Wind load area	1.65 ft^2 max.	1.26 ft^2 max.	.740 ft^2 max.	.496 ft^2 max.
Net weight	5.5 lbs	4.1 lbs	2.9 lbs	2.2 lbs
Electrical				
Forward gain	13.0 dBd*	11.8 dBd*	9.1 dBd*	6.1 dBd*
Front-to-back ratio	20 dB	20 dB	20 dB	20 dB
Maximum SWR	2:1	2:1	2:1	2:1
Band width	2 MHz	2 MHz	4 MHz	4 MHz
Maximum power	250/500 PEP	250/500 PEP	250/500 PEP	250/500 PEP
Impedance w/balun	52 ohms	52 ohms	52 ohms	52 ohms
½ power beam width	35° vertical 35° horizontal	43° vertical 36° horizontal	60° vertical 45° horizontal	95° vertical 60° horizontal
Stacking distance	82″ min.	82″ min.	82″ min.	82″ min.

*Hy-Gain antennas are gain rated against a standard dipole antenna (dBd) instead of a theoretical isotropic source (dBi). This is a more honest and realistic means of comparing forward gain.

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High Pass Filters for TV Sets

provide more than 40 dB attenuation at 52 MHz and lower. Protect the TV set from amateur transmitters 6-160 meters.

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For 300 ohm twin lead

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Drake TV-42-LP **$14.60 ea.**

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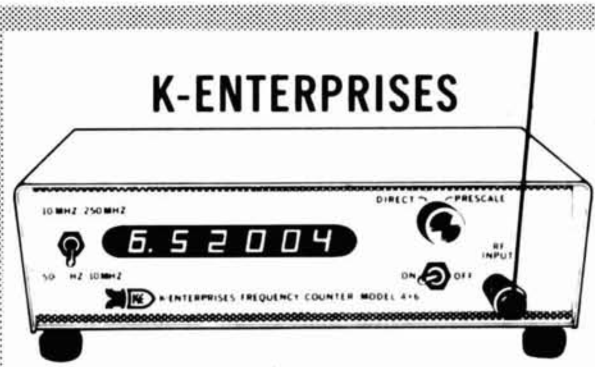

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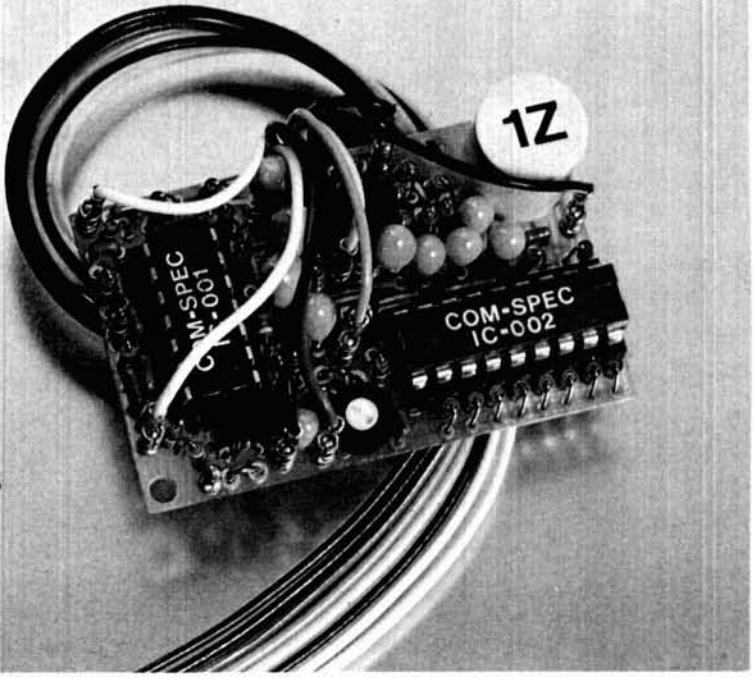

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A. ACCURACY. If it's not accurate, it's not useful.

The Davis Frequency Counter was designed for maximum accuracy from 10 HZ to 500 MHZ. Accuracy is achieved with a precision built-in TCXO (temperature compensated x-tal oscillator) time-base, accurate to ± .0002% (2 parts per million) over a temperature range of 15°C to 55°C. Also essential is the input sensitivity. The Davis Frequency Counter incorporates a high-sensitivity input amplifier that extends the useful range to measure low level signals.

THE DAVIS FREQUENCY COUNTER ALSO FEATURES:

- 8-Digit Display (for more accuracy)
- Large .3" LED Readouts
- Automatic Decimal Point Placement
- Resolution to 1Hz
- Automatic Input Limiting (eliminates input level adjustment)
- Input Diode Protected (protects circuitry against overload)
- Selectable Gate Times, I ms (for rapid reading) and I sec. (for maximum accuracy) Provision for 10 sec. (for maximum accuracy at low frequencies)
- Plug-In Time-Base (for future options)
- Plug-In Prescaler (for future options)
- Automatic Self-Check

SPECIFICATIONS

Frequency Range	10 Hz to 500 MHz
Input Z	1 Meg/20 ph to 50 MH z 50 ohm above 50 MHz
Sensitivity	10 Hz to 25 MHz 10 mv 25 MHz to 50 MHz 30 mv 50 MHz to 500 MHz 150 mv
Time Base Stability	STD ± 2ppm 15° to 55°C (± .0002%) Optional ± .5ppm 0° to 60°C (± .00005%)
Power	117v 50/60 Hz 15W
Size	8.8" x 8" x 2.8" (223.5 x 203.2 x 71.12 mm)
Weight	3 lbs. 10 oz. (1.64 kg.)

500 MHz Kit (TCXO standard) **$249.95**
(for preassembled and calibrated TCXO add $5.00)
Complete Kit: Includes all parts, drilled and plated PC boards, cabinet, switches, hardware and a complete instruction manual with calibrating instructions. Approx. 8 hrs. to assemble.
All parts are guaranteed for 90 days.
Factory service available for $25.

500 MHz Factory Assembled **$349.95**
Factory assembled units are tested and calibrated to specifications, and are guaranteed for 1 year.

Shipping Charges **$2.00**

Instruction and Calibrating Manual **$3.00**
(refundable with purchase)

OPTIONS

Crystal Oven Option (plug-in)
± .5PPM (± .00005%), 0° to 60°C
(increases accuracy over wider temperature range) **$49.95**

Larger (.46") Digit Option
Available on factory assembled unit only **$10.00**

Preamp Probe
20 dB gain range 10 MHz to 500 MHz
(increases sensitivity of counter 10X) **$49.95**

12 volt DC Option **$10.00**

10 Sec. Time-Base Switch Option **$10.00**

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The one 2M rig with everything fully synthesized plus full VFO FM/SSB/CW/AM - 10W Power Out. °PLUS - a self contained 29.5MHz receiver for OSCAR 6/7A. **$795** — IN STOCK

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The only 432 MHz SSB/CW synthesized transceiver on the market. °Excellent for OSCAR 7/Mode B **$495** in stock

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FOR 28, 50, 144, 220 & 432 MHz
PERFECT FOR TS-700A, FT221, ETC.

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$100-100 FT. SPECIALS

KLM KR400 Rotor + 100 ft. rotor cable	**$100.00**
CDE Big Talk + 100 ft. rotor cable	**$100.00**

CABLE 5/32", 6-strand, soft-drawn guy cable. For mast or light tower, **3¢ foot.**

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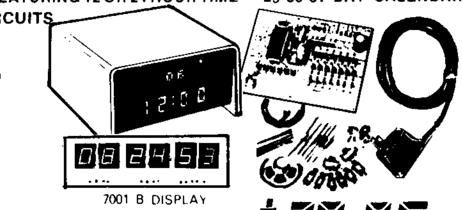
More Details? CHECK—OFF Page 126

6-DIGIT LED CLOCK CALENDAR KIT

DATE-TIME-SNOOZE ALARM & MORE... KIT 7001

OUR TOP OF THE LINE KIT FOR THE BUILDER THAT WANTS THE BEST. A TOTAL PACKAGE, FEATURING 12 OR 24 HOUR TIME — 29-30-31 DAY CALENDAR WITH ALARM, SNOOZE AND AUX. TIMER CIRCUITS

Will alternate time (8 seconds) and date (2 seconds) or may be wired for time or date display only, with other functions on demand. Has built-in oscillator for battery back-up. A loud 24 hour alarm with a repeatable 10 minute snooze alarm, alarm set & timer set indicators. Includes 110 VAC/60Hz power pack with cord and top quality components through-out.



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KIT-7001B WITH 6-.4" DIGITS $39.95
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KITS ARE COMPLETE (LESS CABINET) WITH PC BOARDS, POWER SUPPLY, IC & SOCKET, 16 TRANSISTORS, 9 SWITCHES AND ALL REQUIRED PARTS. ALL 7001 KITS FIT CABINET I AND ACCEPT (OPTIONAL) QUARTZ CRYSTAL TIME BASE KIT # TB-1

$39.95 ea.

6 DIGIT LED CLOCK KIT #850-4

12/24 HR. OPERATION BIG .4" DIGITS - 50/60 HZ OPERATION.

KIT INCLUDES
- INSTRUCTIONS
- QUALITY COMPONENTS
- 50 or 60 Hz OPERATION
- 12 or 24 HR OPERATION

6-LED Readouts (FND-359 Red, com. cathode)
1-MM5314 Clock Chip (24 pin)
13-Transistors
3-Switches
6-Capacitors
5-Diodes
9-Resistors
24-Molex pins for IC socket

LARGE .4" DIGITS! ORDER KIT #850-4 AN INCREDIBLE VALUE!

$11.95 ea. QTY. 1-5
$10.95 ea. QTY. 6-11
$9.95 ea QTY. 12 OR MORE

"Kit #850-4 will furnish a complete set of clock components as listed. The only additional items required are a 7-12 VAC transformer, a circuit board and a cabinet, if desired."

PRINTED CIRCUIT BOARD FOR KIT #850-4, SCREEN PRINTED DRILLED AND SOLDER PLATED FIBERGLASS $2.95
MINI-BRITE RED LED'S (FOR COLON IN CLOCK DISPLAY) .... Pkg. of 5-$1.00
MOLDED PLUG TRANSFORMER 115/10 VAC (WITH CORD) $2.50

NOTE: Entire Clock may be assembled on one PC Board or Board may be cut to remote display.
Kit #850-4 will fit Plexiglas Cabinet II.

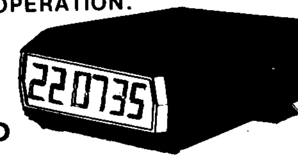
MOBILE LED CLOCK

12 OR 24-HOUR OPERATION

MODEL #2001

12 VOLT AC or DC POWERED FOR FIXED OR MOBILE OPERATION.

SIX LARGE .4" DIGITS!



Approx. Size: 1¾"H x 4"W x 4½"D

- 6 JUMBO .4" RED LED'S BEHIND RED FILTER LENS WITH CHROME RIM
- SET TIME FROM FRONT VIA HIDDEN SWITCHES • 12/24-Hr. TIME FORMAT
- STYLISH CHARCOAL GRAY CASE OF MOLDED HIGH TEMP. PLASTIC
- BRIDGE POWER INPUT CIRCUITRY — TWO WIRE NO POLARITY HOOK-UP
- OPTIONAL CONNECTION TO BLANK DISPLAY (Use When Key Off in Car, Etc.)
- TOP QUALITY PC BOARDS & COMPONENTS - EXCELLENT INSTRUCTIONS
- MOUNTING BRACKET INCLUDED

KIT #2001 COMPLETE KIT (Less 9V. Battery) 29.95 EA. 3 OR MORE $27.95 EA. 115 VAC Power Pack #AC-1 $2.50 EA.

ASSEMBLED UNITS ORDER #2001 WT WIRED & TESTED (LESS 9V. BATTERY) $39.95 EA. 3 OR MORE $37.95 EA. Assembled Units May Be Mixed With Kits for Qty. Price
WIRED FOR 12 HR. OP. IF NOT SPECIFIED OTHERWISE

JUMBO DIGIT CLOCK KIT

A complete Kit (less Cabinet) featuring: six .5" digits, MM5314 IC, 12/24 Hr. time, 50/60 HZ., Plug-Transformer, Line Cord, Switches, and all Parts. (Ideal Fit in Cabinet II) Kit #5314-5 .... $19.95 ea 2/$38.

JUMBO DIGIT CONVERSION KIT — $9.95 ea.

Convert small digit LED clock to large .5" displays. Kit includes 6 - .5" LED's, Multiplex PC Board & easy hook-up info.
Kit #JD-1CC For common Cathode
Kit #JD-1CA for common Anode

PRINTED CIRCUIT BOARDS for CT-7001 Kits sold separately with assembly info. PC Boards are drilled Fiberglass, solder plated and screened with component layout.

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3"x6"x1/8" 95¢ ea. 4/$3

60 HZ. XTAL TIME BASE
Will enable Digital Clock Kits or Clock-Calendar Kits to operate from 12V DC. 1"x2" PC Board
Power Req: 5-15V (2.5 MA. TYP.)
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Clear Plexiglas Stand

Kit #850-4 CP
- 6 Big .4" digits
- 12 or 24 hr. time
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- Plug transformer
- all parts included

Plexiglas is Pre-cut & drilled

Size: 6"H, 4⅓"W, 3"D

A SUPER LOOKING CLOCK!

$23.50 ea. 2/$45.

7-SEG LED

COMMON CATHODE

	COLOR	HT.	DEC PT.	PR EA.
FND-359	RED	.4"	RHDP	$ .95
FND-503	RED	.5"	RHDP	$1.35
DL-750	RED	.6"	LHDP	$2.95
XAN-654	GREEN	.6"	NDP	$1.95
XAN-664	RED	.6"	NDP	$1.95

COMMON ANODE

DL-747	RED	.6"	LHDP	$1.95
XAN-72	RED	.3"	LHDP	$1.25
MAN-72	RED	.3"	LHDP	$1.25
XAN-81	YELLOW	.3"	RHDP	$1.75
XAN-351	GREEN	.3"	RHDP	$1.50
XAN-361	RED	.3"	RHDP	$1.50
XAN-362	RED	.3"	LHDP	$1.50
XAN-662	RED	.6"	NDP	$1.95
XAN-692	RED	.6"	NDP	$1.95

SET OF 6 FND-359 WITH MULTIPLEX PC BOARD - 6.95

Fairchild Super Digit FND-359
.4" Char. Ht. 7 segment LED RED Com. Cath. Direct pin replacement for popular FND-70.
95¢ ea, 10/$8.50
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Form Inexpensive Sockets
100 for $1.25
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74S04	.55
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74S09	.55
74S10	.40
74S20	.50
74S22	.45
74S40	.45
74S50	.45
74S51	.55
74S60	.85
74S64	.55
74S74	.85
74S75	1.75
74S78	1.50
74S86	.95
74S107	.95
74S112	.95
74S113	1.40
74S114	.95
74S133	.75
74S134	.75
74S138	1.75
74S139	1.50
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MM5312	$ 4.95
MM5314	3.95
MM5375AB	3.95
CT-7001	7.95
CT-7002	13.95
50380	3.95
MM5369	2.50

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7448	.95
75491	.65
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LM309K	TO-3	1.25
7805	TAB	.95
7812	TAB	1.25
7812	TO-3	1.50
7815	TO-3	1.25
7815	TAB	1.25
78L15	TO-5	.75
7824	TO-3	1.25
723	DIP	.75
723	TO-5	.75

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REG. SLIDE DPDT	6/$1.
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8	$ .25	$ .22	$ .20
14	.25	.22	.20
16	.28	.25	.23
18	.31	.28	.26
24	.50	.45	.40
28	.60	.55	.50
40	.75	.70	.65

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2N3415	TO-98	5/$1.00
2N3704	TO 92	5/$1.00
2N4400	TO-92	5/$1.00
2N4125	TO-92	5/$1.00
2N4249	TO-92	5/$1.00
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741 TO-5
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And behold there was a great earthquake for the Angel of the Lord descended from Heaven, and came and rolled back the stone from the door.

And the Angel said to the women, "Fear not, for I know that ye seek Jesus who was crucified. He is not here; for he is risen as he said. Come see the place where the Lord lay."

Then the eleven disciples went to Galilee . . . and when they saw him they worshipped him: but some doubted. And Jesus came and spoke to them saying, "All power is given to me in Heaven and in earth. Go ye therefore and teach all nations baptizing them in the name of the Father, Son and Holy Spirit, and lo, I am with you always, even to the end of the world."

Matthew 28, 2-20

We would like to share the message and joy of Christ risen this Easter.

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- **Quality for an Economy Price**
- **Solid State Construction**
- **Linear Switch (FM/SSB)**
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Model	Input	Output	Typical	Frequency	Price
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702B	1W- 5W	60W-80W	1W in/70W out	143-149MHz	$169.00

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The broad band frequency range means that your amplifier is immediately ready to use. No tuning is required for the entire 2-meter band and adjacent MARS channels on TPL's new *Econo-line*.

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Export: EMEC Inc., 2350 South 30th Avenue, Hallandale, Fla. 33009

More Details? CHECK—OFF Page 126

PREASSEMBLED **COMPLETE KIT**

Crescomm Frequency Counters Features:

1 High Sensitivity VHF pre-scaler (built-in). 100mv RMS @ 50Ω @ 300MHZ. Frequency range DC through 600MHZ Model 600.

2 Excellent temperature compensation crystal controlled time base, yielding ± 1 part/ 10^6 stability per hour after 10 min warm-up, $\pm$ 10PPM worst case, from 0^o to $+ 55^oC$!, at 100Hz @ 450MHZ is attainable, typically if calibrated to WWVL. This is approximately 2 parts in 10^7!

3 7 digit display, resolution 100Hz with 10m Sec. gate interval, pre-scaled! 10Hz resolution with 1 Sec. pre-scaled, 1Hz resolution with 10 Sec. gate interval pre-scaled!

4 Built-in 5VDC regulator; input to 3-terminal regulator is accessible for use with 12VDC out-board PS. **You can use this counter mobile.**

5 Easy 6 hour assembly; all circuit broad tracks are pre-tinned, have drilled holes and are plated through! All I.C. sockets included - Makes trouble shooting a breeze.

6 Cabinet, plexy window and all necessary components included for easy, trouble-free assembly.

7 90-Day full coverage warranty.

Optional accessories TCXO time base yielding ½ PPM stability. $79.95

Optional 10 Sec. time gate resolution to 1Hz. $15.00. (Free with purchase of above TCXO)

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More Details? CHECK—OFF Page 126

Counter Preamp/Probe

Sniffs out weak RF

Measuring oscillator frequency on modern communications equipment is tricky business. Circuit density and accessibility problems can be frustrating if you try using clip leads, loops, or coils. You might short out something or pull the oscillator off frequency with close coupling and be unable to get a reading because there just isn't enough signal to drive the counter. Also the counter input cable itself may load the oscillator even though the counter is high impedance. The COUNTER PREAMP is designed to solve these problems. It has 20 dB of gain which increases the sensitivity of your counter ten times. The low capacity insulated probe can pick up the signal just by holding it near the oscillator crystal, coil, or any active component. Sometimes it is possible to read the oscillator thru a plastic case. The preamp has BNC connectors on both ends and can be used as an in-line preamp for scopes, detectors, RF meters, etc., as well as counters. Input is high Z and output is low Z to drive 50 ohm cable. Another serious problem when alligning receivers is that many signal generators shift frequency when the attenuator is moved from high output to drive the counter to low output to make adjustments or check receiver sensitivity. The preamp will give 20 dB of isolation and eliminate the error. Customers have commented that having used the preamp they are now unable to get along without it. Many have made repeat purchases. The preamps are battery powered with three pencells at 25 ma. Output level is 200 to 300 millivolts RMS. One year warranty, money back guarantee. Shipped with probe, less batteries, postpaid cont. USA. Foreign add $5.00. Call 213/342-2714 for COD or save money by sending check to Pagel Electronics.

VHF Preamp, 100 KHz to 200 MHz $35.00
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ELECTRONIC COUNTERS

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Features 3 Regulated power Supplies. 3 Output wave forms. 8 digital level switches. 2 no bounce pulser switches. 8 LEDs with drivers. 1 AP Super strip. Easily constructed. Designed by RETS Electronic Schools.
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Clock Kit (complete less case) **$12.95**
Clock Cabinet **$6.50**
or $4.50 with purchase of clock kit.
Function Generator Kit **$10.95**

FREQUENCY COUNTER
7 Digit 0-300 MHz Freq. Counter **$99.00**
7 Digit 0-500 MHz Freq. Counter **$139.00**
8 Digit 0-30 MHz Counter with options $109.00
0-300 MHz Prescaler for 8 Digit Counter $19.95
0-300 MHz Prescaler with Preamp **$29.95**
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Cabinet accessory package available for all of the above **$24.95**
Anyone of the above kits available pre-assembled for an additional $50.00. Allow 3-4 weeks on assembled units.

Please add $1.00 Shipping/Handling on any order under $15.00
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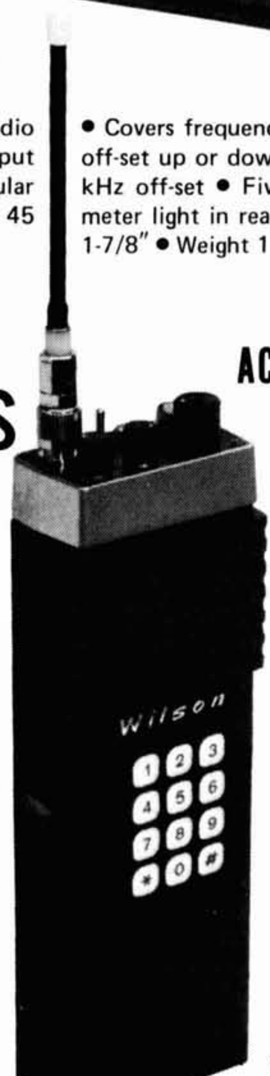

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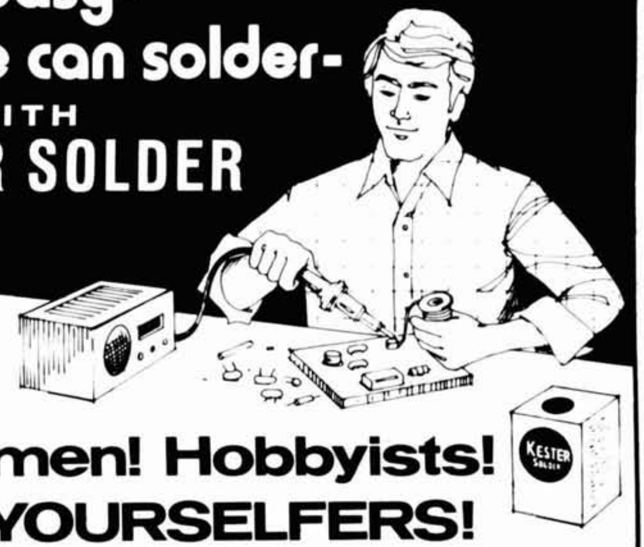

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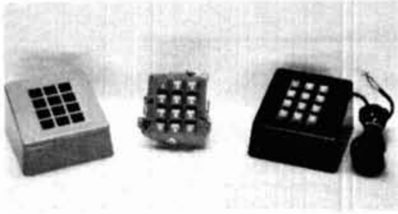

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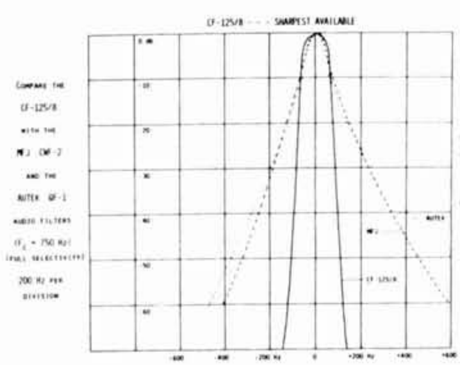

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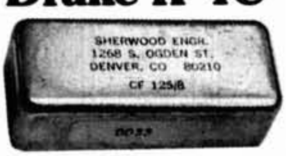

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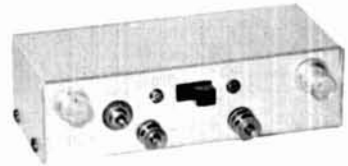

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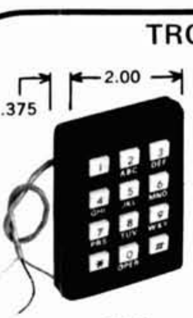

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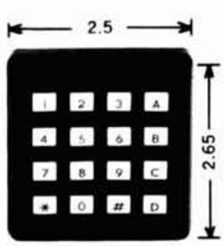

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BARREL KIT #182 JUMBO RED LEDS 15 for $1.98 — 100% material, user cancellation from factory dumps, 3V 10 mils. For 100's of projects, red lens. Cat. No. 4H3369

BARREL KIT #167 ROMS-RAMS-REGISTERS 1-lb. $1.98 — with approx count of 80 pcs. You get more. Mostly marked, mostly National. Some 40 pins. 4H3064

BARREL KIT #166 LONG LEAD DISCS 4 ozs. $1.98 — No counting makes it so you get 150-pc. kit (approx). All marked, 100%. Cat. No. 4H3089

BARREL KIT #163 MINI TRIM POTS 30 for $1.98 1 oz. — Asst. values 100 to 1 meg What a buy. Single turn, ¼ W. Wt. 6 oz. 4H3345

BARREL KIT #5 SCRS, TRIACS QUADRACS 40 for $1.98 — Raw factory stock! All the 10 amp types. Cat.No. 4H2419 Untested.

BARREL KIT #159 MODULAR SWITCHES 25 for $1.98 — Centralab switches. TV-makers excess. Dpdt. 6pdt, etc. Brand new. Cat. No. 4H3150

BARREL KIT #158 MAGNIFIED MAN-3's 12 for $1.98 — MAN-3, 7-seg readout, with built-on magnifier. Factory discontinued line. 100% material. Cat. No. 4H3325 1 oz.

BARREL KIT #157 MOLEX CONNECTORS 75 for $1.98 12 ozs. — Nylon, white cable connectors, asst. factory over-run. NO PICKEE! Mixed in barrels. Cat. No. 4H3324

BARREL KIT #144 RCA PHONO PLUGS 40 for $1.98 — 1,000,000 RCA phono plugs for this one. You hi-fi-ers know wut they are . . . 100% material. 4H3293

BARREL KIT #154 CLOCK CHIPS 20 for $1.98 — We gathered an assortment of clock chip, alarm, calendar, beepers, who knows, all' mixed. Cat. No. 4H3308

Back By Popular Demand 1¢

BARREL KIT #143 75-PC TRANSISTORS $1.98 — 100% material, TO-92, factory discontinued lines, pnp's mixed with npns, 2N4400-1-2-3-4. 4H3290

BARREL KIT #138 PANEL SWITCHES 30 for $1.98 — Did you hear of OAK? Another eqpt maker barrelled all types of rotaries, electric, slides, etc. 4H3268

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☐ SN7403N	.16	.17	☐ SN7473N	.36	.37	☐ SN74154N	1.03	1.04
☐ SN7404N	.16	.17	☐ SN3474N	.31	.32	☐ SN74155N	.70	.71
☐ SN7405N	.18	.19	☐ SN7475N	.49	.50	☐ SN74157N	.70	.71
☐ SN7406N	.20	.21	☐ SN7476N	.31	.32	☐ SN74158N	.85	.86
☐ SN7407N	.25	.26	☐ SN7478N	.79	.80	☐ SN74160N	.88	.89
☐ SN7409N	.18	.19	☐ SN7483N	.69	.70	☐ SN74161N	.88	.89
☐ SN7410N	.16	.17	☐ SN7489N	1.95	1.96	☐ SN74164N	.79	.80
☐ SN7411N	.25	.26	☐ SN7490N	.44	.45	☐ SN74165N	.99	1.00
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☐ SN7421N	.29	.30	☐ SN7494N	.78	.79	☐ SN74176N	.79	.80
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☐ SN7438N	.26	.27	☐ SN74107N	.33	.34	☐ SN74190N	1.15	1.16
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☐ SN7442N	.53	.54	☐ SN74114N	.65	.66	☐ SN74193N	.83	.84
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BARREL KIT #101 RESISTOR SPECIAL 200 for $1.98 — Includes: ⅛, ¼, ½, 1, 2-watters, carbon. 8 oz. 100% good. 4H3054

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BARREL KIT #91 SILVER MICAS 100 for $1.98 — Axial, red case, variety of physical sizes & values. Cat.No. 4H3018

BARREL KIT #88 LITRONICS LED READOUTS 10 for $1.98 — 747's, 727's, singles, triples, etc, .33 to 0.6. Bot from factory, all mixed; have fun! No. 4H2861

BARREL KIT #87 NATIONAL IC BONANZA 100 for $1.98 — Types 8000, 7400 series, DTLs, ROMs, registers, clock & calc. chips, linears, etc. Cat.No. 4H2860 Untested.

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More Details? CHECK—OFF Page 126

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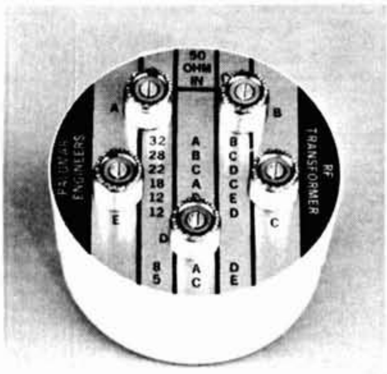

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FEATURES:

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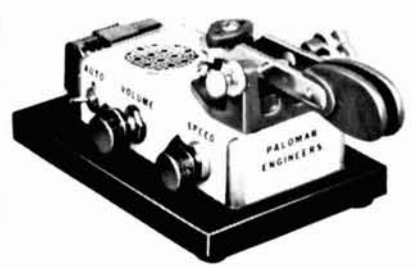

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Coming Events

W6LS 12th Los Angeles Amateur Radio Convention. Saturday and Sunday, May 21 & 22. 2814 Empire Ave., Burbank, CA 91605.

FIRST ANNUAL UNIVERSITY OF PITTSBURGH HAMFEST, March 26, 1977, 10 AM to 6 PM at Pitt Student Union on 5th Avenue near Cathedral of Learning. For more information contact W3YI, Pittsburgh Amateur Radio Association, Box 304, Schenley Hall, University of Pittsburgh, Pittsburgh, PA 15260 or phone 412-624-7768.

THE ROCK RIVER RADIO CLUB HAMFEST is April 24, 1977, at Amboy, Illinois, Lee County at the 4H Center, Routes 30 and 52. Same place as last year. Tickets $1.00 advance. $2.00 at gate. Camper parking available at a nominal fee. Write Carl Karlson, W9ECF, Nachusa, Illinois 61057. Indoor and outdoor facilities.

COME TO CANADA this summer for Ontario Hamfest 77. July 8-10 1977, sponsored by Burlington Amateur Radio Club. Weekend camping, fleamarket, auction, many displays. Write Box 836 Burlington Ont. L7R3Y7 for descriptive brochure.

KANSAS CITY: Eighth Annual Norhtwest Missouri Hamfest, April 23, 24, 1977 at Exhibit Hall 2, Municipal Airport. Forums, swap tables, commercial exhibits, contests, YL-XYL program, free parking. Banquet Saturday evening at world famous Gold Buffet with ARRL President Harry Dannals as guest speaker. Preregistration, $2.00; door, $2.50; preregistration with banquet, $8.00. Info: PHD, P.O. Box 11, Liberty, MO 64068.

VHF—UHF ENTHUSIASTS. 22nd annual West Coast VHF Conference at the Miramar Hotel on the beach in Santa Barbara, CA, May 13-15, 1977. Technical and operating-oriented sessions for beginners and advanced VHFers, plus the traditional receiver noise figure and antenna gain measurements. Registration at 6 PM Friday (May 13). Full day of tech sessions start 9 AM Saturday. Saturday evening, Paul Schuch (WA6UAM) will coordinate noise figure measurements, and a VHF Contest Forum will be led by Wayne Overback (K6YNB). Preregistration $2 until April 30, $3 after. Registration forms, hotel info and details from Dr. Overbeck, Communication Division, Pepperdine University, Malibu, CA 90265.

26th DAYTON HAMVENTION at Hara Arena April 29, 30, May 1, 1977. Technical forums, exhibits, and huge flea market. Program brochures mailed March 7th., to those registered within past three years. For accommodations or advance flyer, write Hamvention, P.O. Box 44, Dayton, Ohio 45401.

ANNOUNCING The first "International" PAN-AMERICAN Ham/Exposition Jamboree, October 29-30, 1977. Write in for regular show mailings. For further information: Broward Amateur Radio Club, Capt. S.F. "Red" Crise (Show Chairman), WA4ZRW, 3701 State Road 84, Ft. Lauderdale, FL 33312.

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200 Volt 100 Amp D08	8.50
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MPSA14	.90	8080A	$29.95
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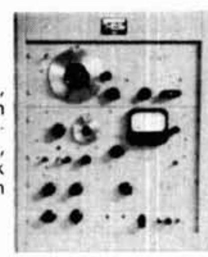

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738AR Provides DC and 400Hz AC at levels from 300uv to 300v in steps of 0.3, 0.5, 1, 2, and 3 with multipliers of .001, .01, .1, 1, 10, and 100v. Acc: 0.1% DC, 0.25% AC. Also provides 10-step measurement to check meter tracking.

739AR Companion unit providing 300kHz to 10mHz signals for RF meter checking. Acc: 0.5%.

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GRAY Electronics

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DIODES/ZENERS

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1N4004	400v	1A	.08
1N4005	600v	1A	.08
1N4007	1000v	1A	.15
1N4148	75v	10mA	.03
1N753A	6.2v	z	.25
1N758A	10v	z	.25
1N759A	12v	z	.25
1N4733	5.1v	z	.25
1N5243	13v	z	.25
1N5244B	14v	z	.25
1N5245B	15v	z	.25

SOCKETS/BRIDGES

8-pin	pcb	.25	ww	.45
14-pin	pcb	.25	ww	.40
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4010	.45
4011	.20
4012	.20
4013	.40
4014	1.10
4015	.95
4016	.35
4017	1.10
4018	1.10
4019	.70
4020	.85
4021	1.35
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4028	.95
4030	.45
4033	1.95
4034	2.45
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4040	1.35
4041	.69
4042	.95
4043	1.25
4044	.95
4046	1.50
4049	.80
4050	.70
4066	1.35
4069	.40
4071	.35
4082	.45

— T T L —

7400	.15
7401	.15
7402	.20
7403	.25
7404	.15
7405	.25
7406	.35
7407	.55
7408	.25
7409	.15
7410	.15
7411	.25
7412	.30
7413	.65
7414	1.10
7416	.25
7417	.50
7420	.15
7426	.40
7427	.45
7430	.15
7432	.45
7437	.45
7438	.35
7440	.25
7441	1.15
7442	.65
7443	.95
7444	.55
7445	.95
7446	.95
7447	.95
7448	1.20
7450	.25
7451	.25
7453	.25
7454	.25
7460	.40
7470	.45
7472	.45
7473	.35
7474	.40
7475	.45
7476	.20
7480	.65
7481	.99
7483	1.00
7485	1.05
7486	.40
7489	2.50
7490	.55
7491	1.15
7492	.95
7493	.45
7494	1.25
7495	.85
7496	.95
74100	1.85
74107	.45
74121	.40
74122	.55
74123	.55
74125	.45
74132	1.35
74141	1.30
74150	1.00
74151	.95
74153	.95
74154	.75
74156	1.15
74157	.75
74161	1.25
74163	1.25
74164	.95
74165	1.50
74166	1.35
74175	.95
74176	1.25
74180	.85
74181	3.25
74182	.95
74190	1.75
74192	1.65
74193	.85
74194	1.45
74195	.95
74196	1.50
74197	1.25
74198	2.35
74367	.85
75108A	.35
75110	.35
75491	.50
75492	.50
74H00	.25
74H01	.25
74H04	.25
74H05	.25
74H11	.25
74H15	.30
74H20	.30
74H22	.40
74H30	.25
74H40	.25
74H51	.25
74H52	.15
74H53J	.25
74H55	.25
74H72	.55
74H101	.75
74H103	.75
74H106	.95
74L00	.35
74L02	.35
74L03	.30
74L10	.35
74L30	.45
74L47	1.95
74L55	.65
74L72	.45
74L75	.55
74S00	.55
74S02	.55
74S03	.50
74S04	.45
74S05	.45
74S08	.45
74S10	.45
74S11	.45
74S20	.50
74S40	.30
74S50	.35
74S51	.45
74S64	.30
74S74	.50
74S112	1.50
74S133	.45
74S140	.75
74S151A	.45
74S153	.45
74S158	.45
74S194	1.50
74S257 (8123)	.25
74LS00	.45
74LS01	.45
74LS02	.45
74LS04	.55
74LS08	.45
74LS09	.45
74LS10	.45
74LS11	.45
74LS20	.50
74LS21	.25
74LS22	.25
74LS32	.55
74LS37	.40
74LS40	.55
74LS42	1.75
74LS74	.95
74LS90	1.30
74LS93	1.00
74LS107	.95
74LS153	1.20
74LS157	.85
74LS164	1.90
74LS367	.85
74LS368	.70

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9309	.45
9322	1.10
9602	1.50

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LM320T12	1.65
LM320T15	1.65
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LM340T-18	1.65
LM340T-24	1.25
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LM340K-15	1.65
LM340K-18	1.65
LM340K-24	1.25
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LM711	.45
LM723	.45
LM725	1.95
LM739	1.50
LM741 8-14	.25
LM747	1.10
LM1307	1.25
LM1458	.95
LM3900	.65
LM75451	.65
NE555	.50
NE556	1.10
NE565	.95
NE566	1.75
NE567	1.35
SN72720	.35
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T-106	135			1.06	1.50
T-80	55	45		.80	.80
T-68	57	47	21	.68	.65
T-50	51	40	18	.50	.55
T-25	34	27	12	.25	.40

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CORE SIZE	MIX Q1 u=125 1-70 MHz	MIX Q2 u = 40 10-150 MHz	SIZE OD (in.)	PRICE USA $
F-240	1300	400	2.40	6.00
F-125	900	300	1.25	3.00
F-87	600	190	.87	2.05
F-50	500	190	.50	1.25
F-37	400	140	.37	1.25
F-23	190	60	.23	1.10

Chart shows uH per 100 turns.

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flea market

STARVED ROCK RADIO CLUB HAMFEST — June 5. S.A.S.E. after 4/1/77 for details. SRRC/W9MKS, RFD #1, Oglesby, Ill. 61348.

RADIO EXPO, CHICAGO. September 17, 18, at Lake County Illinois Fairgrounds. Manufacturers' exhibits, flea market, seminars, and door prizes. 4000 attended last year! Exhibitors are urged to reserve booth space now — call Doug Thornton, Workdays at (312) 595-0020.

NEW JERSEY Delaware Valley Radio Association (W2ZQ/WR2ADE) flea market and auction will be held on Sunday, May 1, 1977, 9 AM rain or shine at the Villa Victoria Academy in West Trenton, N. J. (The school is located adjacent to Rt. 29 near the junction of Rt. 29 and I-95.) Talk in on 07/67 and 146.52. Refreshments are available. Advance registration $1.00; or $1.50 at the gate. For additional information or tickets write: DVRA, P.O. Box 7024, West Trenton, New Jersey 08628, s.a.s.e. please.

HAMFEST! Indiana's friendliest and largest hamfest. Wabash county amateur radio club's 9th annual hamfest will be held Sunday, May 22, 1977, rain or shine, at the Wabash County 4-H fairgrounds in Wabash, Indiana. Large flea market (no table or set-up charge), technical forums, bingo for the xyl, plenty of free parking, lots of good food at reasonable prices. Only one ticket to buy this year. Admission is $2.00 for advance tickets, $2.50 at the gate. Children under 12 years old are admitted free. For more information or advanced tickets, write Bob Mitting, 663 Spring Street, Wabash, Indiana 46992.

MOULTRIE AMATEUR RADIO KLUB 16th. Annual hamfest the last sunday of April at Wyman Park, Sullivan, Ill. Heated indoor area and large outdoor parking area. No charge to vendors. For information write Mark Radio Klub, PO Box 327 Mattoon, Ill. 61938. Talk In 146.94.

KENTUCKY HAM-O-RAMA — Sunday, May 29 (Memorial Day Weekend) at Boone County Fairgrounds, Burlington, Kentucky. 10 minutes south of Cincinnati, 2 miles west of I-75 South, Burlington exit. Prizes, refreshments, exhibits, flea market. NKARC, Box 31, Ft. Mitchell, Kentucky 41017.

F.M. B*A*S*H*, DAYTON, OHIO, April 29, 1977, on the Friday night of the DAYTON HAMVENTION. This is a social evening for all hams and their friends from 8PM til midnight at the Dayton Biltmore Towers, First and Main Street. Admission is free. Sandwiches, beverages, snacks and C.O.D. bar will be available. Live entertainment by TV personality Rob Reider (WA8GFF) and his group. 11PM prize drawing featuring ICOM IC-245 and other prizes. See you where the action is!

POTOMAC AREA VHF SOCIETY sixth annual hamfest Saturday, May 7, 1977, from 8 a.m. to 5 p.m. at Frying Pan Park on West Ox Road in Herndon Virginia, approximately 15 miles west of Washington, D. C. Registration of $3 includes flea market or tail gate sales. Professional food and beverage catering, unlimited parking. Talk-in on 146.52 and 31-91. repeater. For information contact K3DUA or WA3NZL.

THE MESILLA VALLEY RADIO CLUB Sponsors Whitey's Bean Feed and Swap-Fest Sunday, April 24th, at 10:00a.m. Located near Las Cruces, New Mexico at La Mesa with talk-ins on 16-76, 04-64 and 3940 KC. Fun for all the family with big prizes, plenty of food and the usual beverage truck. All included for $5.00 for adults $1.75 for kid tickets. Eat, drink and win a prize with Whitney, K5ECQ as host. Free overnight parking at grounds so come for a spell. All correspondence should be made with Thomas B. Rapkoch Jr., 640 W. Las Cruces Ave., Las Cruces, New Mexico 88001.

THE CENTRAL MASS. AMATEUR RADIO ASSOC. auction and flea market April 15, 1977 at American Legion Post 341, 1023 Main Street, Worcester, Mass. Talk-in on 37/97. Doors open at 6 PM, auction starts 7 PM sharp. Flea market table (items $5 and under) rent will be $5. Items $5 to $100 auctioned with 15% commission. Area set aside for direct buyer-seller barter for items $100 and up again with 15% commission to CMARA.

SWAPFEST Sunday May 1, 1977, sponsored by the Brownfield Amateur Radio Club, Brownfield, Texas. Info from Earl Elrod, Box 821, Brownfield, TX 79316.

VACATIONLAND HAMFEST at the Erie County Fairgrounds, Sandusky, Ohio, May 22, 1977. Plenty of flea market tables available ($4 each), $1 for flea market (8 acres for trunk sales). Advance tickets $1.50, $2.00 at gate. Talk-in 52/52. More info from Erie Amateur Radio Society, P.O. Box 2037, Sandusky, OH 44870. Free transportation to Cedar Point during Hamfest.

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More Details? CHECK—OFF Page 126

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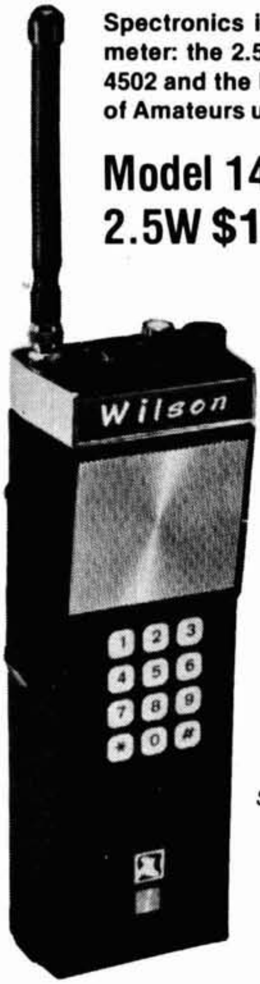

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- 10.7 IF and 455 kHz IF
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- S-Meter/Battery Indicator
- Size: 8 7/8 x 1 7/8 x 2 7/8
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- 10.7 and 455 kHz IF
- .3 Microvolt Sesnsitivity for 20 dB Quieting
- Weight: 1 lb. 4 oz. less Battery
- Battery Indicator
- Size: 8 7/8 x 1 3/4 x 2 7/8
- Switchable 1 & 2.5 Watts Output @ 12 VDC
- Current Drain: RX 25 MA, TX 400 MA
- Microswitch Speaker Mic
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Frequency Range 420 - 450 MHz

- 6 Channel Operation
- Individual Trimmers on all TX/RX Crystals
- All Crystals Plug In
- 12 kHz Ceramic Filter
- 21.4 and 455 kHz IF
- .3 Microvolt Sensitivity for 20 dB Quieting
- Weight: 1 lb. 4 oz. less Battery
- Battery Indicator
- Size: 8 7/8 x 1 3/4 x 2 7/8
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- Current Drain - RX 25 MA, TX 400 MA
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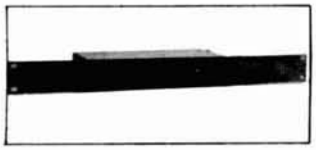

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flea market

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CADILLAC MICHIGAN 17th Annual Swap-Shop will be held Saturday, May 21st 1977 at the National Guard Armory, Cadillac, Michigan. Free parking, everyone welcome. Tickets $2.00. Talk-in on 146.37/97.

18TH ANNUAL STARC HAMFEST May 7, Binghamton, NY. Flea market, snack bar, tech talks, hourly door prizes. Admission $2.00, banquet reservation $6.00. Indoor exhibit reservation $5.00/table. Contact STARC, P.O. Box 11, Endicott, NY 13760.

ELMIRA, NEW YORK HAMFEST, Saturday, Sept. 24, 1977. Bigger and better than 1976. Plan on it.

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THE CHAMPAIGN/LOGAN AMATEUR RADIO CLUB will hold its annual flea market on May 15, 1977, at the West Liberty Lion's Park, West Liberty, Ohio. Free admission. Trunk sales and tables $1.00. Door prizes. Talk-in on 146.52.

THIRD ANNUAL NORTHWESTERN PENNSYLVANIA HAMFEST, May 7, Crawford County Fairgrounds, Meadville, Pa. Free admission. Flea Market begins at 10:00 AM $2 to display. Hourly door prizes, refreshments, commercial displays welcome. Indoors if rain. Talk-in 146.04/64 and 146.52. Details C.A.R.S. PO Box 653 Meadville, Pa. 16335.

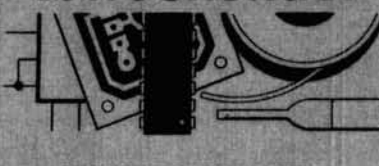
PENNSYLVANIA: Warminster Amateur Radio Club's "Hammart", Flea Market and Auction Sunday, May 15, from 9 to 4 at William Tennent Intermediate High School, Street Road (Route 132), 2 miles East of York Road (Route 263), Warminster, Bucks County, Pa. Registration $1.00, Tailgating $2.00 additional. Talk-in on 147.69-09; 146.16-76 and 146.52. Further information write to Horace Carter, K3ZAC, 38 Hickory Lane, Doylestown, Pa. 18901.

BIGGEST FEST ON THE GULF COAST. Mobile Amateur Radio Club hamfest and computerfest April 16 & 17. All the newest equipment on display. Computers, too. Swap and Shop all day Saturday from 9 to 5. Banquet at 7 PM. Doors open Sunday 9 AM. Prize drawing at 1 PM. Lots of prizes. Actives for ladies and children. Campsites available. Over 1500 people expected. For more info contact Marvin Uphaus, K4BVG, Tuttle Ave., Mobile, AL 36604.

W7DQ SKAGIT AMATEUR RADIO CLUB 24th annual convention April 22, 23 and 24 at the Bryant Grange Center, near Seattle, WA.

BIRMINGHAMFEST AMATEUR RADIO CONVENTION. WHEN: May 7 and 8, 1977. Where: Alabama State Fairgrounds, Birmingham, and Rodeway Inn — Oxmoor at I-65 and Oxmoor Road. Attractions: one of the country's largest flea markets, technical and operating forums, huge prize drawing, manufacturers and distributors displays, ladies and childrens activities. Booth Information: booth display area will be offered free of charge to bona fide distributors, manufacturers, publishers, etc., on a first-come, first-served basis. Others may rent space in inside or outside flea market areas at a small charge. Talk-In: 34/94, 3965 kHz. No admission charge. Prize ticket donations — $1. For booth display space, information, and reservations, write, Birminghamfest, P. O. Box 603, Birmingham, Alabama 35201.

the weekender



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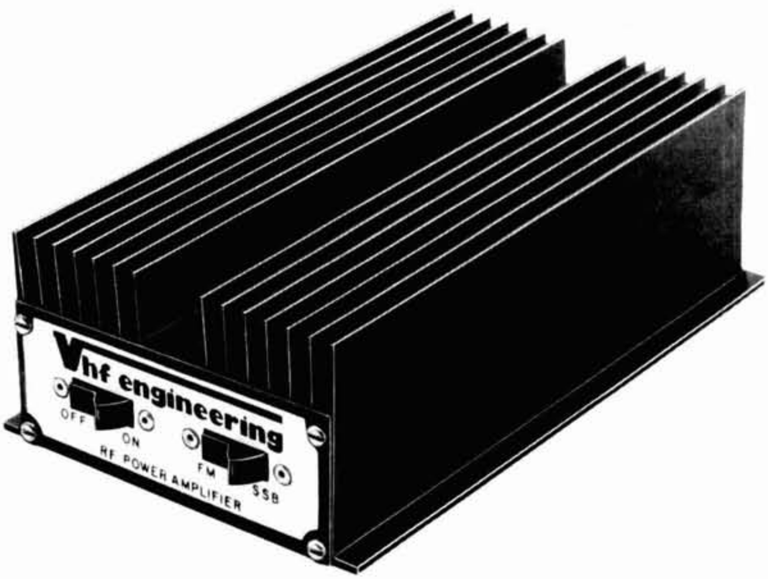

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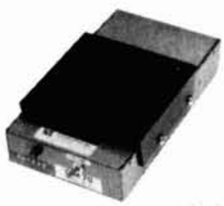

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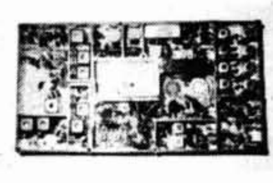

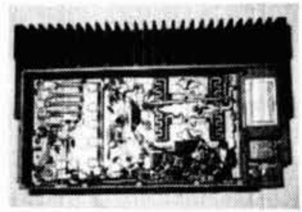

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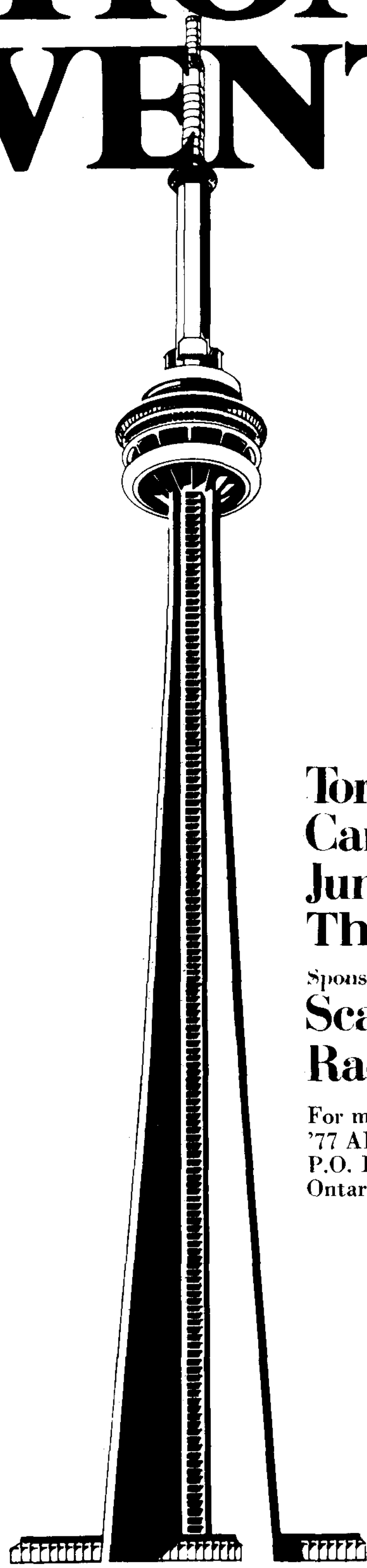

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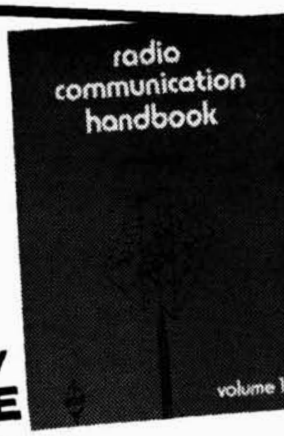

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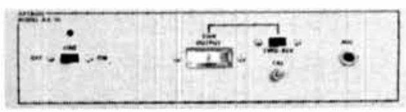

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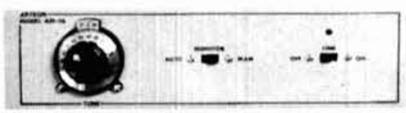

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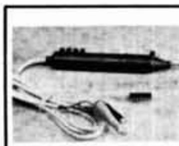

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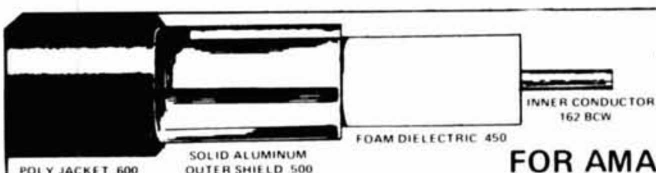

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SN7446N	.81	SN74144N	4.50	SN74200N	5.59
SN7447N	.69	SN74145N	1.15	SN74279N	.90
SN7448N	.79	SN74147N	2.35	SN74251N	1.79
SN7450N	.26	SN74148N	2.00	SN74284N	6.00
SN7451N	.27	SN74150N	1.00	SN74285N	6.00
SN7453N	.27	SN74151N	.79	SN74367N	.75
SN7454N	.20	SN74153N	.89		

MANY OTHERS AVAILABLE ON REQUEST
20% Discount for 100 Combined 7400's

CMOS

CD4000	.25	CD4035	1.85	74C04N	.75
CD4001	.25	CD4040	2.45	74C10N	.65
CD4002	.25	CD4042	1.90	74C20N	.65
CD4006	2.50	CD4044	1.50	74C30N	.65
CD4007	.25	CD4046	2.51	74C42N	2.15
CD4009	.59	CD4047	2.75	74C73N	1.50
CD4010	.59	CD4049	.79	74C74	1.15
CD4011	.25	CD4050	.79	74C90N	3.00
CD4012	.25	CD4051	2.95	74C95N	2.00
CD4013	.47	CD4053	2.95	74C107N	1.25
CD4016	.56	CD4060	3.25	74C151	2.90
CD4017	1.35	CD4066	1.75	74C154	4.00
CD4019	.55	CD4069	.45	74C157	2.15
CD4020	1.49	CD4071	.45	74C160	3.25
CD4022	1.25	CD4081	.45	74C161	3.25
CD4023	.25	CD4511	2.50	74C163	3.00
CD4024	1.50	CD4518	2.50	74C164	3.25
CD4025	.25	MC14566	3.00	74C173	2.60
CD4027	.69	74C00N	.39	74C193	2.75
CD4028	1.65	74C02N	.55	74C195	2.75
CD4029	2.90			MC4044	4.50
CD4030	.65			MC14016	.56

LINEAR

LM300H	.80			LM1310N	2.95
LM301H	.35			LM1351N	1.65
LM301CN	.35			LM1414N	1.75
LM302H	.75			LM1458C	.65
LM304H	1.00	LM370N	1.15	LM1496N	.95
LM305H	.95	LM373N	3.25	LM1556V	1.85
LM307CN	.35	LM377N	4.00	LM2111N	1.95
LM308H	1.00	LM380N	1.39	LM2901N	2.95
LM308CN	1.00	LM380CN	1.05	LM3065N	.69
LM309H	1.10	LM381N	1.79	LM3900N	.55
LM309K	.99	LM382N	1.79	LM3905N	.60
LM310CN	1.15	NE501K	8.00	LM3909	1.25
LM311H	.90	NE510A	6.00	LM5556N	1.85
LM311N	.90	NE531H	3.00	MC5558V	1.00
LM318CN	1.50	NE536T	6.00	LM7525N	.90
LM319N	1.30	NE540L	6.00	LM7535N	1.25
LM320K-5	1.35	NE550N	.79	8038B	4.95
LM320K-5.2	1.35	NE555V	.39	LM75450	.49
LM320K-12	1.35	NE560B	5.00	75451CN	.39
LM320K-15	1.35	NE561B	5.00	75452CN	.39
LM320T-5	1.75	NE562B	5.00	75453CN	.39
LM320T-5.2	1.75	NE565H	1.25	75454CN	.39
LM320T-8	1.75	NE565N	1.25	75491CN	.79
LM320T-12	1.75	NE566CN	1.25	75492CN	.89
LM320T-15	1.75	NE567H	1.95	75494CN	.89
LM320T-18	1.75	NE567V	1.50	**RCA LINEAR**	
LM320T-24	1.75	LM703CN	.45	CA3013	2.15
LM323K-5	9.95	LM709H	.29	CA3032	2.56
LM324N	1.80	LM709N	.29	CA3035	2.48
LM339N	1.70	LM710N	.79	CA3039	1.35
LM340K-5	1.95	LM711N	.39	CA3046	1.30
LM340K-6	1.95	LM723H	.55	CA3059	3.25
LM340K-8	1.95	LM723N	.55	CA3060	3.25
LM340K-12	1.95	LM733N	1.00	CA3080	.85
LM340K-15	1.95	LM739N	1.00	CA3081	2.00
LM340K-18	1.95	LM741CH	.35	CA3082	2.00
LM340K-24	1.95	LM741CN	.35	CA3083	1.60
LM340T-5	1.75	LM741-14N	.39	CA3086	.85
LM340T-6	1.75	LM747H	.79	CA3089	3.75
LM340T-8	1.75	LM747N	.79	CA3091	10.20
LM340T-12	1.75	LM748H	.39	CA3123	2.15
LM340T-15	1.75	LM748N	.39	CA3130	1.39
LM340T-18	1.75	LM1303N	.90	CA3140	1.25
LM340T-24	1.75	LM1304N	1.19	CA3600	1.75
LM350N	1.00	LM1305N	1.40	RC4194	5.95
LM351CN	.65	LM1307N	.85	RC4195	3.25

74LS00 TTL

74LS00	.29			74LS139	1.95
74LS02	.29			74LS151	1.55
74LS03	.29	74LS74	.49	74LS153	1.89
74LS04	.35	74LS75	.69	74LS157	1.55
74LS05	.29	74LS76	.49	74LS162	2.25
74LS08	.29	74LS83	1.75	74LS163	2.25
74LS10	.29	74LS85	2.49	74LS164	1.95
74LS13	.69	74LS86	.49	74LS175	1.95
74LS14	1.75	74LS90	1.25	74LS181	3.69
74LS20	.29	74LS92	1.25	74LS190	2.85
74LS26	.39	74LS93	1.25	74LS191	2.85
74LS27	.39	74LS95	1.95	74LS192	2.85
74LS28	.39	74LS96	1.89	74LS193	2.85
74LS30	.29	74LS107	.59	74LS194	1.89
74LS32	.39	74LS109	.59	74LS195	1.89
74LS40	.39	74LS112	.59	74LS257	1.75
74LS51	.29	74LS132	1.25	74LS260	.55
74LS55	.29	74LS136	.59	74LS279	.79
74LS73	.49	74LS138	1.89	74LS670	3.95

CLOCK CHIPS

MM5309	6 Digit, BCD Outputs, Reset PIN	$9.95
MM5311	6 Digit, BCD Outputs, 12 or 24 Hour	4.95
MM5312	4 Digit, BCD Outputs, 1 PPS Output	4.95
MM5314	6 Digit, 12 or 24 Hour, 50 or 60 Hz	4.95
MM5316	4 Digit, Alarm, 1 PPS Output	6.95
MM5318	Video Clock Chip, For Use With (MM5841 - $9.95)	9.95
CT7001	6 Digit, Calendar, Alarm, 12 or 24 Hour	5.95

DATA HANDBOOKS

7400	Pin out & Description of 5400/7400 ICS	$2.95
CMOS	Pin out & Description of 4000 Series ICS	$2.95
Linear	Pin out & Functional Description	$2.95

ALL THREE HANDBOOKS $6.95

FAIRCHILD TECHNOLOGY KITS FAIRCHILD

DIGITS

FTK0001	0.5" High Common Cathode Digit	$1.00
FTK0002	0.5" High Common Anode Digit	1.00
FTK0003	.357" High Common Cathode Digit	.75
FTK0004	0.8" High Common Cathode Digit	2.00
FTK0005	0.8" High Common Anode Digit	2.00

0.8" HIGH DISPLAY ARRAYS

FTK0010	12 Hour, 3½ Digit Clock Display	7.00
FTK0011	24 Hour, 4 Digit Clock Display	8.00

LED LAMPS

FTK0020	10 Red LED Lamps	1.00
FTK0021	5 Mixed Colored LED Lamps	1.00
FTK0022	10 LED Mounting Clips	1.00
FTK0023	5 Three Piece LED Mounting Adapters	1.00

PHOTO TRANSISTORS

FTK0030	5 Flat Lens Photo Transistors	1.00
FTK0031	5 Round Lens Photo Transistors	1.00
FTK0032	3 Flat Lens Photo Darlingtons	1.00
FTK0033	3 Round Lens Photo Darlingtons	1.00

PHOTO ARRAYS

FTK0040	9-Element Tape Reader Array	16.00
FTK0041	12-Element Card Reader Array	24.00
FTK0042	Reflective Opto Coupler	4.00

COUPLERS

FTK0050	3 General Purpose Opto Couplers	1.00
FTK0051	Darlington Opto Coupler	1.00

MOS CLOCK CIRCUITS

FTK0400	Digital Clock/Calendar Circuit (FCM7001)	7.00
FTK0401	Digital Clock/Calendar with BCD Outputs (FCM7002)	7.00
FTK0402	Direct Drive Digital Clock Circuit with AC Output (FCM3817A)	5.00
FTK0403	Direct Drive Digital Clock Circuit with DC Output (FCM3817D)	5.00
FTK0405	Direct Drive Digital Clock/Calendar Circuit (FCM7015)	6.00

KITS

FTK0106	Automobile Clock Kit	40.00

DISCRETE LEDS

.125" dia

XC209	Red	10/$1
XC209	Green	4/$1
XC209	Orange	4/$1
XC209	Yellow	4/$1

.190" dia

XC111	Red	10/$1
XC111	Green	4/$1
XC111	Yellow	4/$1
XC111	Orange	4/$1

.200" dia			.185" dia			.200" dia		
XC22	Red	10/$1	XC526	Red	10/$1	XC556	Red	10/$1
XC22	Green	4/$1	XC526	Green	4/$1	XC556	Green	4/$1
XC22	Yellow	4/$1	XC526	Yellow	4/$1	XC556	Yellow	4/$1
XC22	Orange	4/$1	XC526	Orange	4/$1	XC556	Orange	4/$1
SSL-22	RT	4/$1	XC526	Clear	4/$1	XC556	Clear	7/$1

.085" dia MV50 — 6/$1

INFRA-RED LED ¼" x ¼" x 1/16" Flat 5/$1.00

SPECIAL * — XC556 Red 100/$8.00 1000/$60.00 — SPECIAL *

DISPLAY LEDS

DL707 DL338

TYPE	POLARITY	HT		TYPE	POLARITY	HT	
MAN 1	Common Anode	270	2.95	MAN 3640	Common Cathode-orange	300	1.75
MAN 2	5 x 7 Dot Matrix	300	4.95	MAN 4710	Common Anode-Red	400	1.95
MAN 3	Common Cathode	125	3/1.00	DL701	Common Anode-red ±	300	.99
MAN 4	Common Cathode	187	1.95	DL704	Common Cathode	300	.99
MAN 7	Common Anode	300	1.25	DL707	Common Anode	300	.99
MAN 7G	Common Anode-green	300	1.95	DL 728	Common Cathode	500	1.95
MAN 7Y	Common Anode-yellow	300	1.95	DL741	Common Anode	600	1.50
MAN 52	Common Anode-green	300	.99	DL 747	Common Anode	600	2.25
MAN 64	Common Anode-red	400	.99	DL 750	Common Cathode	600	2.49
MAN 74	Common Cathode	300	1.50	DL 338	Common Cathode	110	.50
MAN 82	Common Anode-yellow	300	.99	FND70	Common Cathode	250	.75
MAN 84	Common Cathode-yellow	300	.99	FND503	Common Cathode	500	1.00
MAN 3620	Common Anode-orange	300	1.75	FND507	Common Anode	500	1.00

FACTORY REJECTS

ATARI GAME BOARDS

BOARD A — 8½" x 16" — **$6.95 ea.**
Over 60 each reusable IC's, Misc. Transistors, Resistors, Diodes, Caps, Crystals, Switch, etc.
ONLY 500 EA AVAILABLE

DELUXE BOARD B — 11½" x 18" — **$9.95 ea.**
Over 100 each reusable IC's, Misc. Transistors, Resistors, Diodes, Caps, Crystals, Switches, LEDS, etc.
ONLY 500 EA AVAILABLE

NEW AY-3-8500-1 TV GAME CHIP

- Score display
- Selectable angles, speed and bat size
- Auto or manual ball service
- Realism sounds
- 6 Games — tennis, hockey, soccer, squash, pelota, practice and two rifle games

28 Lead Dual in Line

The circuit is intended to be battery powered. Minimum number of external components required for complete system.

$24.95

IC SOLDERTAIL — LOW PROFILE (TIN) SOCKETS

	1-24	25-49	50-100		1-24	25-49	50-100
8 pin	$.17	.16	.15	24 pin	$.38	.37	.36
14 pin	.20	.19	.18	28 pin	.45	.44	.43
16 pin	.22	.21	.20	36 pin	.60	.59	.58
18 pin	.29	.28	.27	40 pin	.63	.62	.61
22 pin	.37	.36	.35				

SOLDERTAIL STANDARD (TIN)

14 pin	$.27	.25	.24	28 pin	$.99	.90	.81
16 pin	.30	.27	.25	36 pin	1.39	1.26	1.15
18 pin	.35	.32	.30	40 pin	1.59	1.45	1.30
24 pin	.49	.45	.42				

SOLDERTAIL STANDARD (GOLD)

8 pin	$.30	.27	.24	24 pin	$.70	.63	.57
14 pin	.35	.32	.29	28 pin	1.10	1.00	.90
16 pin	.38	.35	.32	36 pin	1.75	1.40	1.26
18 pin	.52	.47	.43	40 pin	1.75	1.59	1.45

WIRE WRAP SOCKETS (GOLD) LEVEL #3

10 pin	$.45	.41	.37	24 pin	$1.05	.95	.85
14 pin	.39	.38	.37	28 pin	1.40	1.25	1.10
16 pin	.43	.42	.41	36 pin	1.59	1.45	1.30
18 pin	.75	.68	.62	40 pin	1.75	1.55	1.40

Plastic Push Button Switch

- 18 AWG Solid Wire - 5" Long
- .50 (wide) X .60 (high) ⅜-27 Thread
- 8 AMP @ 14 Volt - 1 AMP @ 110 Volt

		1-9	10-Up
J-188-1	Push On-Push Off	.59	.49
J-188-2	Normally Open	.59	.49
J-188-3	Normally Closed	.59	.49

MINATURE TOGGLE SWITCH

JMT-221	DPDT	on/off/on	$1.95
JMT-223	DPDT	on/none/on	$1.75
JMT-121	SPDT	on/off/on	$1.50
JMT-123	SPDT	on/none/on	$1.25

CLIPLITE 8/$1.49

LED MOUNTING SYSTEM
use with XC556 LEDS
Specify Colors — Red - Green - Amber - Yellow

DIP SWITCH

These switches feature 7 SPST slide switches in a molded dip. They are ideally suited for microprocessor applications. **$1.95**

50 PCS. RESISTOR ASSORTMENTS $1.75 PER ASST.

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

ASST. 8R Includes Resistor Assortments 1-7 (350 PCS.) **$10.95 ea.**

$5.00 Minimum Order — U.S. Funds Only
California Residents — Add 6% Sales Tax
Spec Sheets - 25¢ — Send 35¢ Stamp for 1977A Catalog
Dealer Discount Available — Request Pricing

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PHONE ORDERS WELCOME — (415) 592-8097
All Advertised Prices Good Thru April

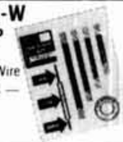
WIRE WRAP CENTER

HOBBY-WRAP TOOL-BW-630

- Battery Operated (Size C)
- Weighs ONLY 11 Ounces
- Wraps 30 AWG Wire onto Standard DIP Sockets (0.25 inch)
- Complete with built-in bit and sleeve

$34.95
(batteries not included)

WIRE-WRAP KIT — WK-2-W



WRAP • STRIP • UNWRAP

- Tool for 30 AWG Wire
- Roll of 50 Ft. White or Blue 30 AWG Wire
- 50 pcs. each 1", 2", 3" & 4" lengths — pre-stripped white wire

$11.95

WIRE WRAP TOOL WSU-30

WRAP • STRIP • UNWRAP - $5.95

WIRE WRAP WIRE — 30 AWG

25 ft. min. $1.25 50 ft. $1.95 100 ft. $2.95 1000 ft. $15.00
SPECIFY COLOR — White - Yellow - Red - Green - Blue - Black

CUTTER CRIMPER TOOL (CS-8)

- Plier Nose (serrated-jaw)
- Scissors Action Cutting
- 6 Bolt Cutters (4-40, 5-40, 6-32, 8-32, 10-32, 10-24)
- Crimp Stations (7mm Auto — 22-20 to 12-X0 elect.)
- "Up-Front" Wire Cutting
- Scissors Action Stripping (No. 22-20 to No. 10)
- Crimp Stations — insulated (2w-20 to 12-10 elect.)

Actual Size - 8¼" length **$8.50**

Permacel Electrical Tape

- ¾" (wide) X 66 ft (long) • All weather • Not import

$1.25 per roll — $9.95 per 10 roll package

ZENERS — DIODES — RECTIFIERS

TYPE	VOLTS	W	PRICE	TYPE	VOLTS	W	PRICE
1N746	3.3	400mm	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	8/1.00	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	6/1.00	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	1.50
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

SCR AND FW BRIDGE RECTIFIERS

C36D	15A @ 400V	SCR	$1.95
C38M	35A @ 200V	SCR	1.95
2N2328	1.6A @ 200V	SCR	.50
MDA 980-1	12A @ 50V	FW BRIDGE REC.	1.95
MDA 980-3	12A @ 200V	FW BRIDGE REC.	1.95

TRANSISTORS

MPS A05	5/$1.00			PN4249	4/$1.00
MPS A06	5/$1.00			PN4250	4/$1.00
2N2219A	3/$1.00	PN3567	3/$1.00	2N4400	4/$1.00
2N2221	4/$1.00	PN3568	4/$1.00	2N4401	4/$1.00
2N2222A	5/$1.00	PN3569	4/$1.00	2N4402	4/$1.00
2N2369	5/$1.00	2N3704	5/$1.00	2N4403	4/$1.00
2N2369A	4/$1.00	2N3705	5/$1.00	2N4409	5/$1.00
2N2484	4/$1.00	2N3706	5/$1.00	2N5086	4/$1.00
2N2906A	4/$1.00	2N3707	5/$1.00	2N5087	4/$1.00
2N2907A	5/$1.00	2N3711	5/$1.00	2N5088	4/$1.00
2N2925	5/$1.00	2N3724	$.65	2N5089	4/$1.00
2N3053	2/$1.00	2N3725	$1.00	2N5129	5/$1.00
2N3055	$.89	2N3903	5/$1.00	2N5138	5/$1.00
MJE3055	$1.00	2N3904	4/$1.00	2N5139	5/$1.00
MJE2955	$1.25	2N3905	4/$1.00	2N5209	5/$1.00
2N3392	5/$1.00	2N3906	4/$1.00	2N5951	5/$1.00
2N3398	5/$1.00	2N4013	3/$1.00	C106B1SCR	2/$1.00
		2N4014	3/$1.00	2N5432	$2.00
		2N4123	10/$1.00		

CAPACITOR CORNER

50 VOLT CERAMIC DISC CAPACITORS

	1-9	10-49	50-100		1-9	10-49	50-100
10 pf	.05	.04	.03	.001µF	.05	.04	.035
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
100 VOLT MYLAR FILM CAPACITORS							
.001mf	.12	.10	.07	.022mf	.13	.11	.08
.0022	.12	.10	.07	.047mf	.21	.17	.13
.0047mf	.12	.10	.07	.1mf	.27	.23	.17
.01mf	.12	.10	.07	.22mf	.33	.27	.22
+20% DIPPED TANTALUMS (SOLID) CAPACITORS							
.1/35V	.28	.23	.17	1.5/35V	.30	.26	.21
.15/35V	.28	.23	.17	2.2/25V	.31	.27	.22
.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
MINIATURE ALUMINUM ELECTROLYTIC CAPACITORS							
Axial Lead				**Radial Lead**			
.47/50V	.15	.13	.10	.47/25V	.15	.13	.10
1.0/50V	.16	.14	.11	.47/50V	.16	.14	.11
3.3/50V	.15	.13	.10	1.0/16V	.15	.13	.10
4.7/25V	.16	.14	.12	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	.11
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

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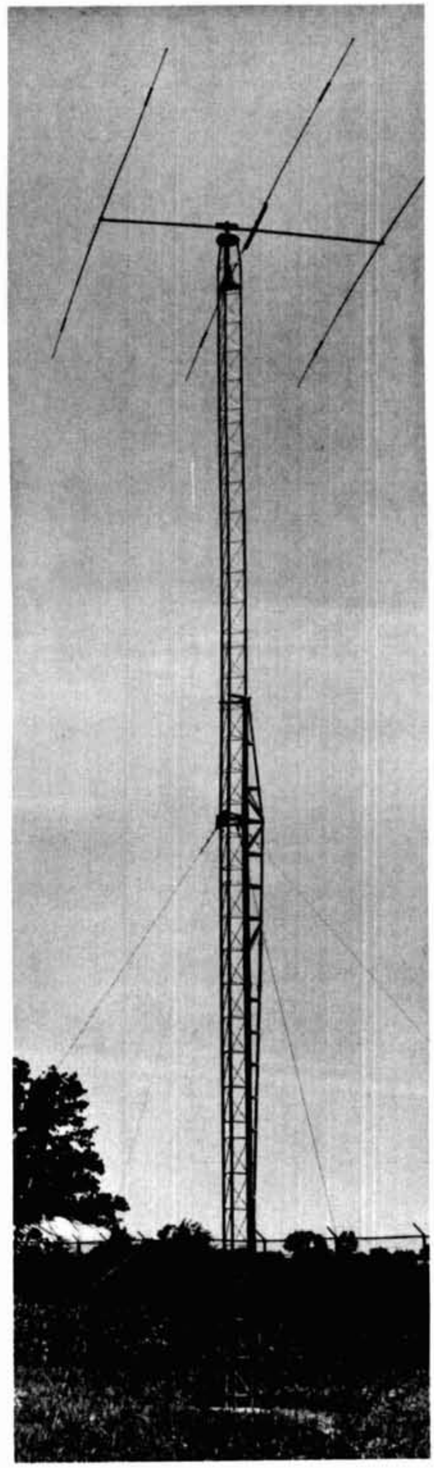

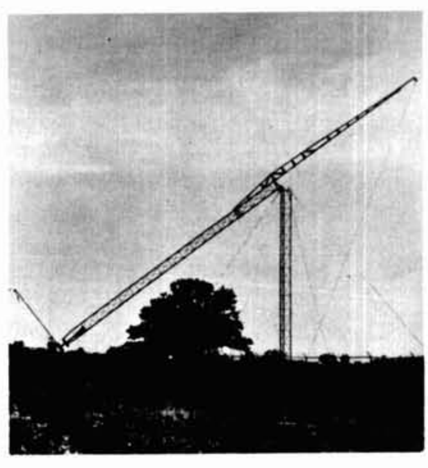

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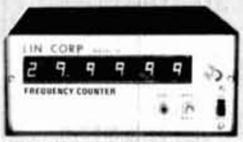

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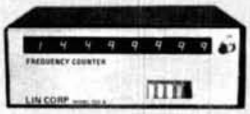

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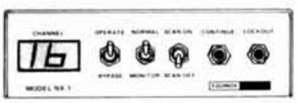

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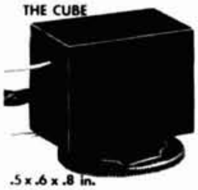

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11C70DC	600 MHz flip/flop with reset	12.30
11C83DC	1 GHz 248/256 Prescaler	29.20
11C90DC	650 MHz Prescaler Divide by 10/11	16.00
11C90DM	same as above except Mil. version	24.00
11C91DC	605 MHz Prescaler Divide by 5/6	16.00
11C91DM	same as above except Mil. version	24.00
95H90DC	350 MHz Prescaler Divide by 10/11	9.50
95H90DM	same as above except Mil. version	16.50
95H91DC	350 MHz Prescaler Divide by 5/6	9.50
95H91DM	same as above except Mil. version	16.50
T.I. TMS4060/C2107, 4K RAM		19.01

Batteries

NI-CAD's AA cells 1.25 volts at 500 mahr. $0.49
Gel-Cell 12 volts at 1.5 Amp Hr. #GC-1215 $19.95

Crystals

1.000000 MHz	4.95
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10 MHz	$4.95

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Motorola U43 GGT	$49.95
GE TPL	$99.95
GE MT-33	$39.95
Collins 26U	$150.00
CBS Labs Model IIIs, AUDIMAX	$775.00
GRC10 Radio Set	$169.95
Model T33BAT	$39.95 FOB Phoenix

Motorola MC14410CP CMOS tone Generator uses 1 MHz Crystal to produce standard dual frequency telephone dialing signal. Directly compatible with our 12 key Chomeric pads. Kit includes the following.

1 MC14410CP
1 Touch Tone Pad
1 1 MHz Crystal
1 Printed Circuit Board (From Ham Radio Sept. 1975)
And all other parts for assembly.

$19.95

Fairchild 95H90DC Prescaler divide by 10 to 350 MHz. Will take any 35 MHz Counter to 350 MHz. Kit includes the following.

1 95H90DC
1 2N5179
2 UG-88/u BNC's
1 Printed Circuit Board
And all other parts for assembly.

$29.95

Fairchild 11C90DC Prescaler divide by 10 to 650 MHz. Will take any 65 MHz Counter to 650 MHz or with a 82S90 it will divide by 10/100 to 650 MHz. This will take a 6.5 MHz counter to 650 MHz. Kit includes the following.

1 11C90DC
1 2N5179
2 UG-88/U
1 MC7805CP
1 Bridge
1 Printed Circuit Board and all other parts for assembly.
82S90 add $5.70 to total.

$59.95

Fairchild 3817 Clock Kit from Ham Radio, Feb. 1976, Pg. 26 — All parts included except transformer and case. 12 hour $24.95 24 hour $29.95

I.C.'s

8223B	$3.00
2102	$1.99

MuRata 10.7 MHz Ceramic Filters #SFW-10.7MA $3.95

Johanson and Johnson Trimmer Capacitors

1 to 14 pf.	$1.95
1 to 20 pf.	$1.95

Ferrite Beads 12 for .99 or 120 for 9.99

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2N3070	1.50	2N5460	.90	MFE3002	3.35
2N3436	2.25	2N5465	1.35	MPF102	.45
2N3458	1.30	2N5565	5.45	MPF121	1.50
2N3821	1.60	3N126	3.00	MPF4391	.80
2N3822	1.50	MFE2000	.90	U1282	2.50
2N4351	2.85	MFE2001	1.00	MMF5	5.00
2N4416	1.05	MFE2008	4.20	40673	1.39
2N4875	1.75	MFE2009	4.80	40674	1.49

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2E26	5.00	811	6.95	7984	4.95
3B28	4.00	811A	9.95	8072	32.00
4X150A	15.00	931A	11.95	8156	3.95
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F-93X	6.5v to 40v at 750 ma.	3.53
F-92A	6.5v to 40v at 1 amp	4.59
F-91X	6.5v to 40v at 300 ma.	2.72
N-51X	Isolation 115vac at 35va.	2.80
Model D-2	6.5v at 3.3 amps 6.5v at 3.3 amps	4.95
C-912-034	22vct at 200 ma. 11v at 250 ma.	1.49
BE-12433-001	30v at 15 ma.	.49
C-404-024	18vct at 400 ma.	1.49
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F-107Z	12V @ 4A or 24 V @ 2A	7.80
P6377	12v @ 4a or 24v @ 2a	6.31
P6378	12v @ 8a or 24v @ 4a	10.31
P8196	80vct @ 1.2a	6.28

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2N1692	15.00	2N3927	11.50	2N5636	11.95
2N1693	15.00	2N3948	2.00	2N5637	20.70
2N2631	4.20	2N3950	26.25	2N5643	20.70
2N2857	1.80	2N3961	6.60	2N5641	4.90
2N2876	12.35	2N4072	1.70	2N5643	20.70
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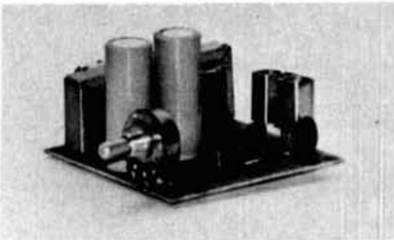

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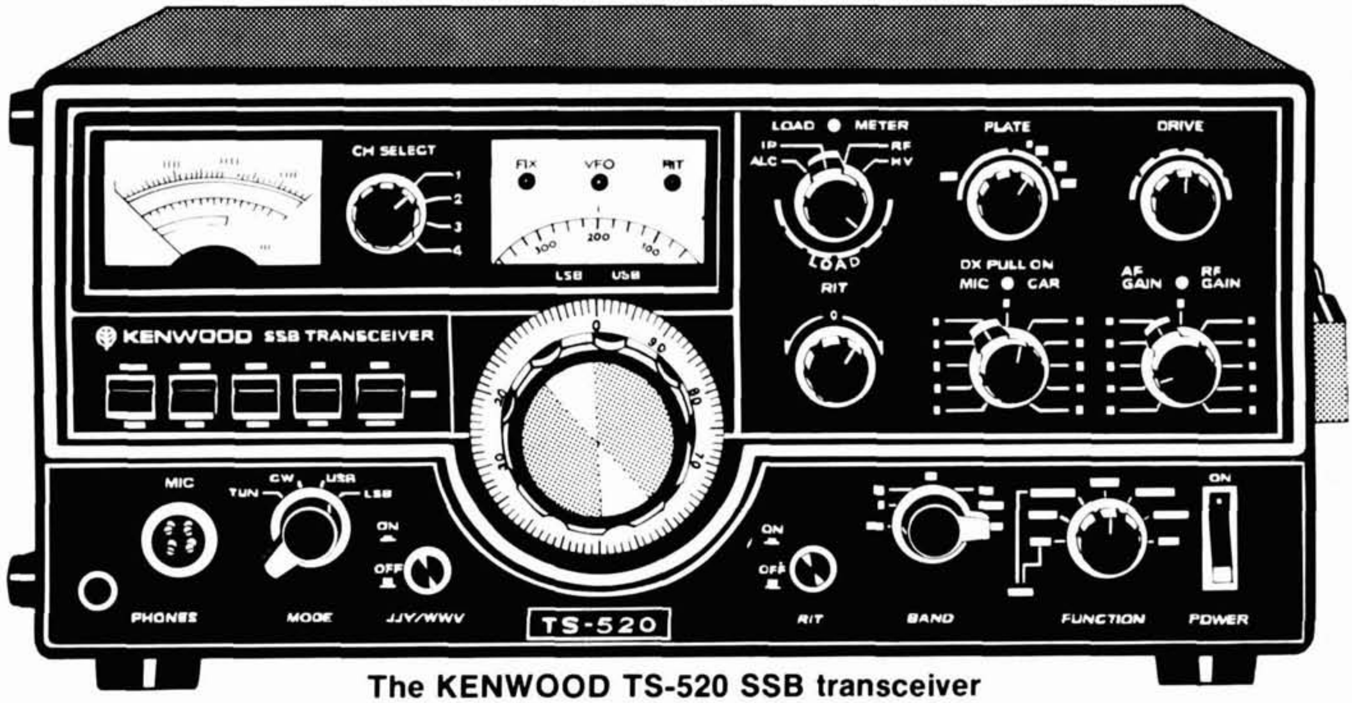

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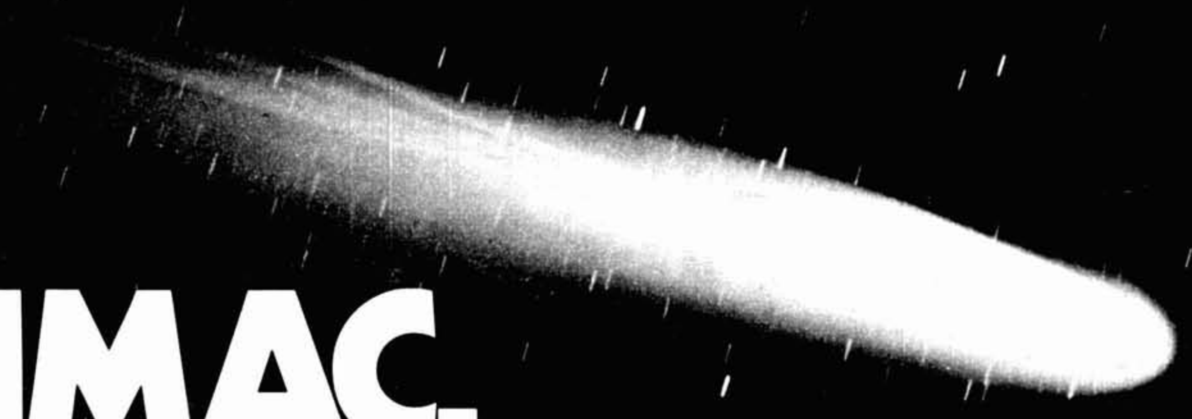

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