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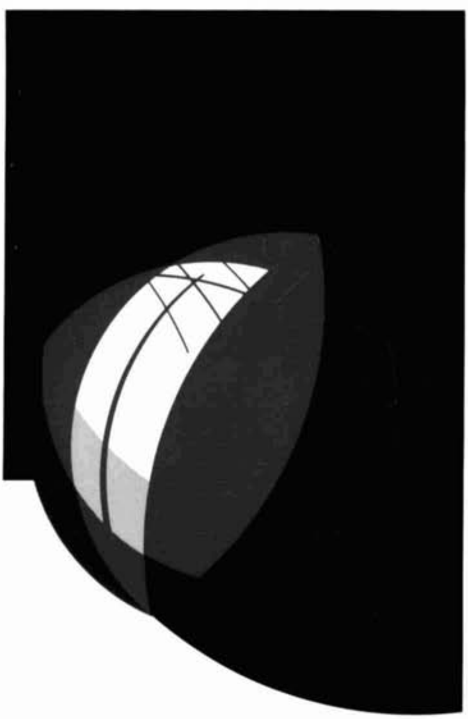

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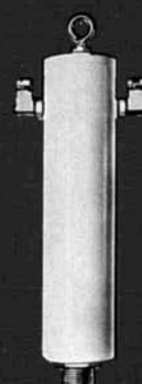

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contents

a second look

by jim fisk

Periodically, a story about the first ham station makes the rounds. It goes something like this: In Boston, just before 1910, there were three young wireless operators, Albert S. Hyman, Robert Almy and Reginald Murray. These young men put together a small wireless station, and since there were no licensing regulations at the time, they decided to call it the Hyman-Almy-Murray Wireless Station.

They soon found that that was quite a fist-full on CW, so they took the first two letters of each name and Station HYALMU went on the air. They used this callsign for several months, but were nearly involved in an international incident when a Mexican ship named the HYALMO almost went aground off the New Jersey coast. They decided that their HYALMU callsign was too close to HYALMO for comfort, so they took the first initials of the three names, and put Station HAM on the air. The first ham station? No. But probably the first, and possibly the only amateur radio station with the HAM callsign.

Several sources have labeled this story as outright fiction, but it has been just persistent enough to arouse my curiosity. Several years ago I decided to track it down, and to determine once and for all if the story had any grain of truth.

Since at least one of the young men was supposedly a student at Harvard University and a member of the Harvard Wireless Club, I decided to start my search in the archives at Cambridge, Massachusetts. Sure enough, deep in the yellowed files there was an entry: Dr. Albert Salisbury Hyman, A.B., 1915; M.D., 1918. Could this be the same man who gave his last initial to Wireless Station HAM?

It looked promising. The Harvard Alumni Records Office revealed that Dr. Hyman was alive and well, and furnished me with his current address. I wrote to him, and his gracious reply confirmed that, yes, he was the same person who, with boyhood friends Robert Almy and Reginald Murray, had put Wireless Station HAM on the air. He also revealed that the "original ham" story got its start in a story written by wartime correspondent Percy Greenwood for a New York medical publication.

According to Dr. Hyman, Station HAM was not located at Harvard, as Greenwood's story indicates, but was actually at the Roxbury High School, which in the early 1900s was a prep school for the Ivy League.

A further search through the records disclosed that Dr. Hyman, before his graduation, was a shipboard wireless operator for the Eastern Steamship Line that ran ships from New York to Boston through the Cape Cod Canal. After graduation from medical school Dr. Hyman became a heart specialist and owned one of the first electrocardiograph machines in New York (in 1923). He was also the inventor of the artificial pacemaker used in resuscitating the dying heart.

So goes the saga of Wireless Station HAM. It was definitely not the first amateur radio station — that honor goes to some unknown wireless operator at least 20 years earlier. Nor does it have anything to do with the fact that radio amateurs are called "hams." That term goes back to the early days of wire telegraphy when unskilled, incompetent operators were called hams by their more experienced colleagues. The connotation is less than desirable.

Jim Fisk, W1DTY
editor

polarization of satellite signals

Some background for understanding polarization losses with practical antenna information for better satellite reception

Katashi Nose, KH6IJ, 4207 Huanui Street, Honolulu, Hawaii 96816

With the upsurge of interest in amateur satellite transmissions, amateurs are becoming aware of the problems and challenges of proper antenna polarization for optimum signal reception. The subject is vast and there is still a great deal of experimentation to be done. I hope that these few introductory notes and the reports on some of my experiments might be of interest and assistance to others.

Faraday rotation

Back in 1845, Michael Faraday, the experimental genius, discovered that when a block of glass is subjected to a strong magnetic field, it becomes optically active — it is able to polarize light and, conversely, is able to detect the plane of polarization of light. Light is an electro-magnetic wave and so are radio signals.

A linearly, polarized signal (from a dipole or equivalent) originating in space rotates as it travels through space, the amount of rotation being a function of the magnetic field and electron content of the intervening medium.

Fig. 1 shows a linear wave originating in some form of synchrotron radiation (radio signal) which is rotated as it progresses towards earth through space.

At the earth end of this ribbon, the receiving dipole or array must be rotated to match the polarization at that instant in order to get maximum pickup. A simplified relationship in the case of light is given by:

$$\emptyset = KHL$$

Where:

$\emptyset$ = the angle of rotation,

K = the medium through which the wave travels, sometimes called the Verdet constant,

H = the magnetic field strength and

L = the thickness of the medium.

A more practical simplified relationship in the case of radio frequencies is given by:

$$\phi = \frac{K}{f^2} NH \cos \theta \, dh$$

where:

ϕ = amount of rotation,

K = proportionality constant,

f = frequency,

H = magnetic field

θ = angle between the plane of the incident wave and the magnetic field and

dh = differential element of the path length.

It follows that you can measure the

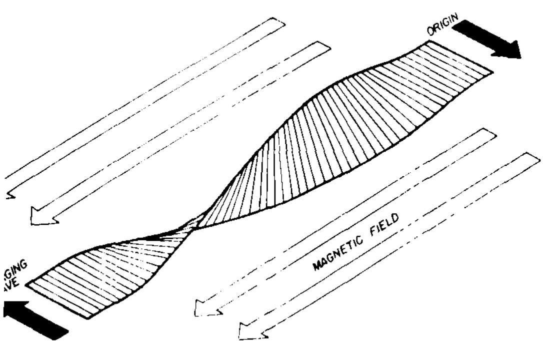


fig. 1. Rotation of wave through a magnetic field.

electron content of the intervening medium by measuring the Faraday rotation. Note that rotation is frequency sensitive.

The photograph shows the ATS-1 communications satellite which weighs 800 pounds and is spin stabilized at 100 rpm, its axis of rotation being parallel to

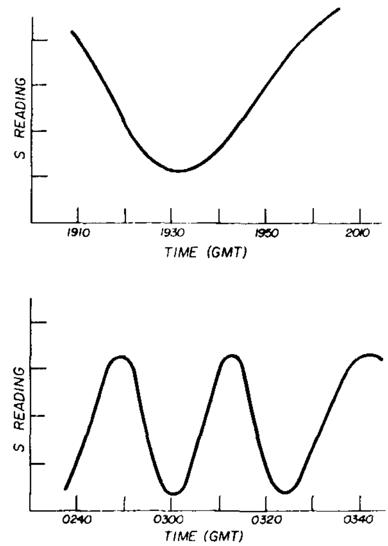


fig. 2. Signal variation due to Faraday rotation. Both readings were taken the same day — November 1, 1971.

the earth. it is geosynchronous at 149° West longitude at the equator which places it around 72° elevation from Hawaii.

The top edge of the photo shows the eight vhf antennas which are phased electronically so that the array sends a conical beam with an aperature roughly corresponding to one third the earth which it illuminates.

The signal is linearly polarized, yet when it arrives on earth, it might have undergone several rotations because of the Faraday effect.

The number of rotations varies from a few degrees per hour to as much as five turns per hour and sometimes even more. A typical rotation curve is shown in **fig. 2.** Depending on the type of receiving antenna used, signal differential between planes of polarization can be as much as 15 dB. The implication is obvious: you must correctly polarize the antenna for successful reception.

If a piece of thin mica, commonly known as a quarter-wave plate, is inserted in the path of polarized light, it will introduce a 90° phase shift and a half-wave plate can cause a 180° shift.

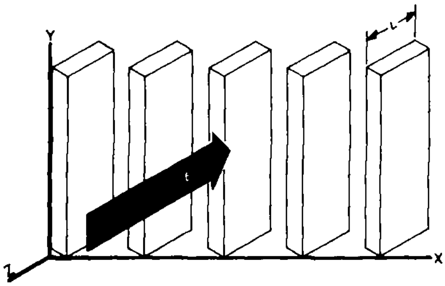


fig. 3. A wave polarizer — the E vector at 45 degrees emerges as a circular wave.

Likewise, consider **fig. 3** which shows a wave polarizer, which is merely a collection of conducting slabs. If a linearly-polarized radio wave is incident at 45° into the page (z direction), the incident electric field can be resolved into two components. The x component is unaffected since there is little interaction with the slabs in the x direction, but the y plane wave is reduced in velocity. If the L dimension of the slabs is sufficient to retard the voltage by 90°, the wave emerges from the back side (behind the page) circularly polarized. Conversely, a circularly-polarized wave originating on the back side of the page will emerge out of the page as a linearly-polarized wave.

If the slab dimension L is increased to 2L (180° phasing), then the emerging wave will again be linearly polarized since the x and y voltages will be opposite in phase, but the polarization will be displaced by 90°. Increasing the slab dimension to 3L makes the emerging wave circularly polarized again but with a counterclockwise rotation. Finally, if the slab dimension is increased to 4L, the emerging wave is linearly polarized at 45° as in the incident wave. Now, if you are confused, think what you have to contend with when fabricating a circularly polarized antenna using linear (Yagi) elements. Sometimes, at microwave frequencies, slabs of plastic are used to introduce the proper phase delay.

Incidently, the picket fence analog to explain polarization, so common in high school textbooks, is pedagogically wrong. **Fig. 4** shows a rope being vibrated in a circular mode (unpolarized light) which when passing through a slit (polarizer) emerges as a polarized wave. This polarized wave passes through a parallel slit (analyzer) but not through a perpendicular slit.

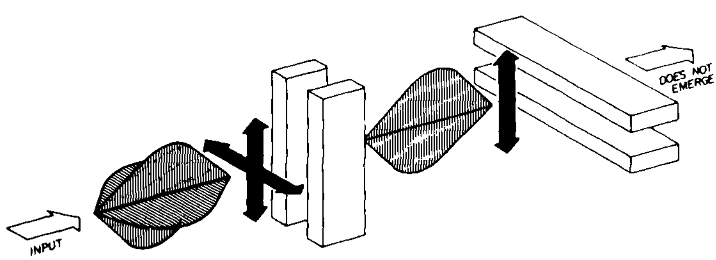


fig. 4. The erroneous picket fence analog of a polarized electromagnetic wave.

The student *erroneously thinks of the* plane of the opening as passing the polarized wave. Consider **fig. 5** which shows an unpolarized wave from a microwave generator impinging on a grid of wires held at right angles to the plane of polarization. By the picket fence analog the wave should be stopped by the second fence, but no, it goes right through without attenuation. But if the polarizer is held parallel to the plane of the wave, the wave does not emerge from the other side. If the grid elements are less than a half-wave long, there is no interaction with either orientation. Grid elements longer than a halfwave behave like a halfwave grid. The phenomenon is one of absorption and reradiation not of transmission as implied by the picket fence analog.

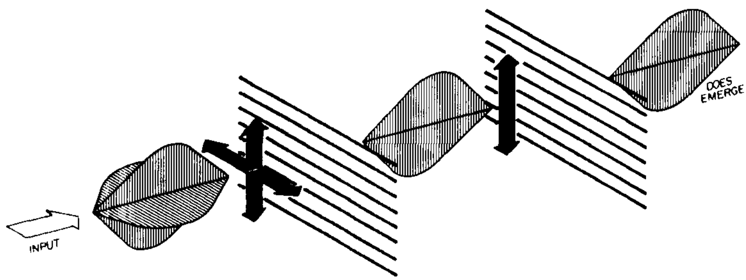


fig. 5. An electromagnetic analog of polarization.

Where does all this discussion lead to? Merely, that fabricating a circularly polarized antenna of the crossed Yagi type for picking up vertically and horizontally polarized signals is not an easy task, and that what most of us like to think of as being a circularly-polarized, crossed Yagi is wishful thinking.

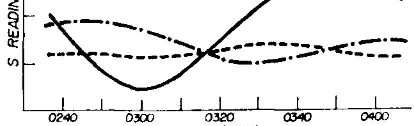


fig. 6. Comparison of different antennas during changing polarization.

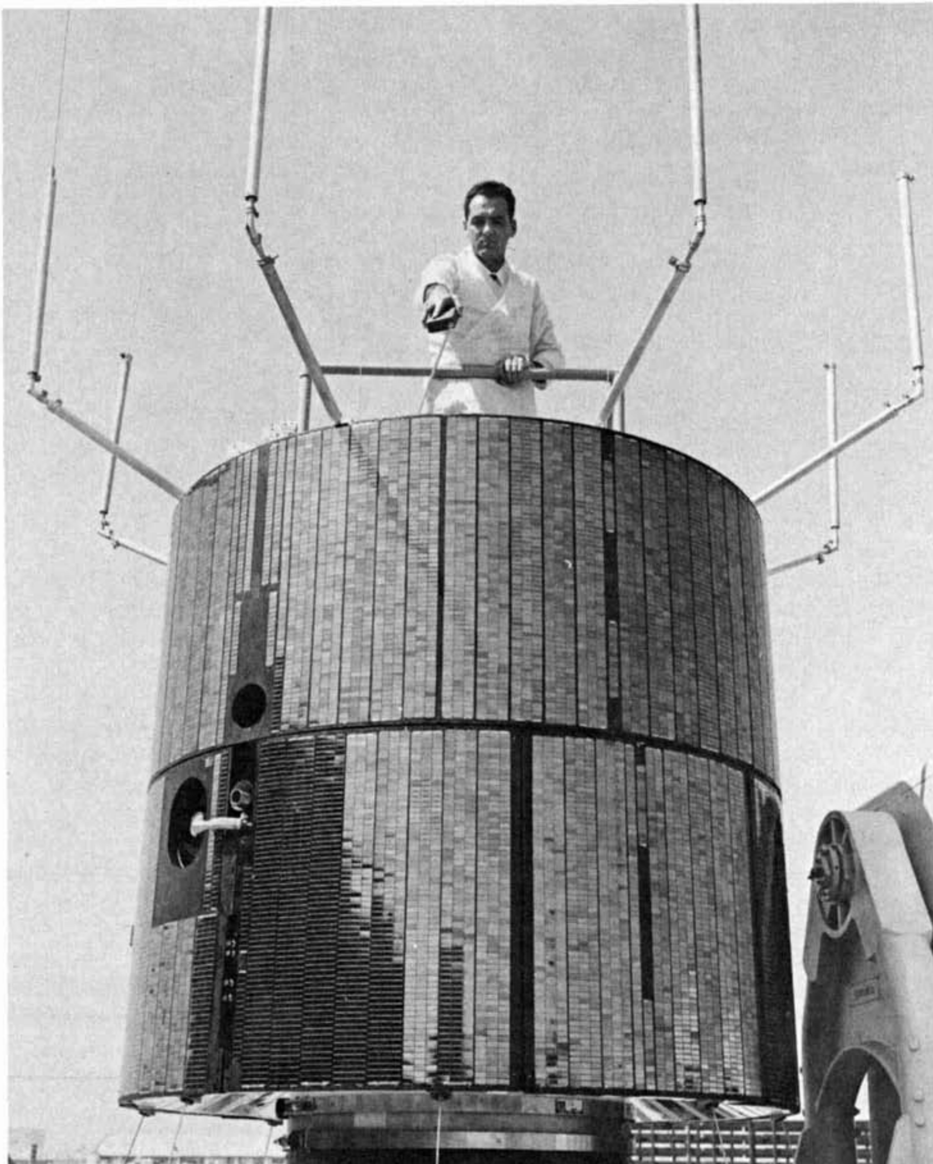
Look at what you're up against. First, each individual Yagi must produce exactly the same magnitude of voltage at its feedpoint which means that it must be electrically balanced – a difficult task at best.[1] The phasing sections must introduce the right quadrature voltage. Unless you have a line stretcher or some lumped circuit equivalent, this is difficult to do.

More basic than matching and phasing is that you lose a precious 3 dB by the mere act of phasing the two antennas. Depending on the phase and amplitude relations of the two antennas, the combination rejects some circularly polarized waves. Only one-half the energy is accepted.

practical antenna performance

Fig. 6 shows the non-circularity of a commercially made, inexpensive, 4-bay 80-element antenna designed to pick up either vertical or horizontal polarization. Also shown on the same graph is the performance of a home-made, 20-element crossed Yagi, and a four-element linear Yagi which was rotated to follow the Faraday rotation. Readings were taken in rapid sequence on the transponded signal from ATS-1.

Where does all this discussion lead to? Nothing more than what we already know – that the simplest is best. Unless you have access and the know-how to adjust a circular antenna configuration, better stick to a simple Yagi for amateur satellites. If you think big, you won't get there in time.



The Applications Technology Satellite (ATS-1) shows antennas at top. (Photo courtesy Hughes Aircraft Company.)

transmitting antenna

Unfortunately, the solution to the uplink problem is completely different. Here we have no choice but to use a circularly-polarized antenna. The entree into the satellite (assuming a linearly polarized antenna) is contingent upon correct polarity because we have no way of telling what the correct polarization is until we send up a transponding signal.

Fig. 7 shows the rotation necessary for entree into ATS-I as a function of polarization. As can be expected, with high power (one kilowatt output) there is a broad entree. With low power (ten watts), however, you have to be sure the polarity is correct.

A crossed Yagi is really a turnstile antenna shown in **fig. 8,** but arranged so that there is axial radiation by use of directors and reflectors. Note that the two radiators must be fed equal-magnitude voltage of the right quadrature – 90° lead or lag for right or left polarization. Any unbalance, either in phase or magnitude, results in an elliptical pattern which degenerates to a linear pattern in the extreme case.

Since you have no choice, make the best of a difficult situation and attempt to tune the Yagis for equal current in the driven elements. This is discussed in a previous article.[1]

The cheap and dirty method of tuning for balance is to feed considerable power into the antenna and to go around with a pencil on the end of an insulated stick and touch the various elements. By considerable juggling of element lengths as discussed in the previous article, you can get some semblance of circularity.

Meishin Electric Company of Tokyo has come up with an interesting variation. The driven elements are of turnstile configuration, but the reflectors and directors are circular discs of the proper diameter. Helix antennas are a tale in themselves.

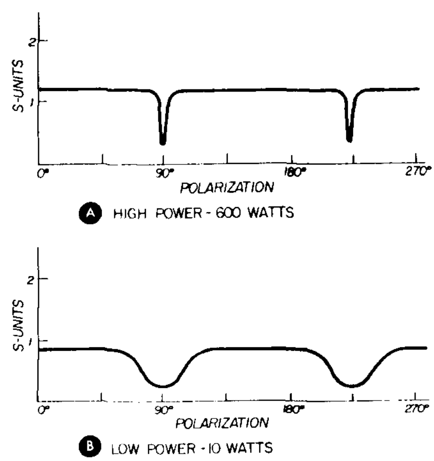


fig. 7. Results of the uplink polarization test with entree to the ATS-1 satellite. Tests were run using an 11-element linear Yagi at experimental station KB2XXK on December 13, 1971.

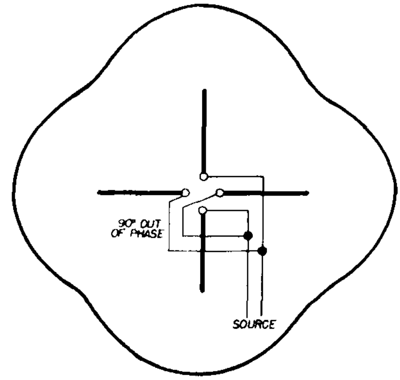


fig. 8. A turnstile antenna and the resultant pattern.

space diversity

There is little to be gained by space diversity since fading is not due to ionospheric reflection or absorption, unless you are receiving at a low elevation angle. Deep fades which occur occasionally are still unaccounted for.

acknowledgements

I wish to thank Mrs. Mary L. Burton, KH6HGO, a graduate mechanical engineer, who patiently took data for months and processed it, and to Mr. Fred Matsunaga, PhD candidate in physics, who stuck with us through the first few months when we did not know which way we were flying.

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ham radio

high-performance single sideband rf speech clipper

This solid-state rf speech clipper features the use of crystal filters to provide maximum talk power with minimum distortion

Leslie Moxon, G6XN, Hampshire, England

As I pointed out in last month's issue of *ham radio,* there is considerable confusion over the various aspects of speech processing for ssb.[1] I discussed the different processing systems that are currently being used by amateur radio operators, and tried to clear a way through some of the conflicting statements that have permeated the literature. This month I will discuss a practical ssb speech-processing system, along with some circuit details and operating advantages.

practical speech processing

The first systems I tried were based on surplus FT243 crystals built into a half-lattice filter as shown in **fig. 1.** Some crystal selection was necessary, particularly for the shunt crystals (Y1 and Y3) which produce the notch on the carrier side of the sideband. Inductor L1 was wound by cut and try to resonate at the desired i-f with about 200 pF at C1. The secondary of L1 consists of 5 or 6 turns wound symmetrically over the primary.

The value of resistor R2 is fairly critical, and if you have trouble obtaining proper passband response, this is the first component to change. If you are not able to obtain suitably-spaced commercial crystals, you may have to resort to grinding. However, the 2.2 to 2.5 kHz frequency separation required is fairly small, and crystals should be readily available.

Although I used an i-f of 5435 kHz, the same alignment procedure should be satisfactory for other frequencies. I have used these half-lattice crystal filters with 2N706 input and output circuits in several transceivers, but construction is simplified through the use of integrated circuits. I used ICs for an improved rf clipper design based on 5200-kHz Japanese crystal filters as shown in **fig. 2.** There should be no difficulty in adapting this circuit to FT243 crystal filters, although there would be some performance loss because of the lower number of crystals.

design

The first clipping stage is operated with a low value of bias to provide a large dynamic range (and relatively poor limiting). The second stage operates with a higher bias level which provides the over-all limiter characteristic plotted in **fig. 3.** The diodes I used in the circuit of **fig. 2** were salvaged from surplus computer boards, but any switching types should be suitable (I used 1N914s in earlier designs).

The value of capacitor C1 (about 3 pF) may need to be changed to insure that the characteristics of the limiter circuits coincide. The value of C2, about

2 pF, is selected so that U2 is not driven too hard. With a six-volt power supply, the output voltage was 380 mV, varying with the supply voltage as indicated in **fig. 3.**

solid-state ssb exciter

The rf clipper is part of the solid-state ssb exciter shown in block form in **fig. 4.** This transceiver is based primarily on Plessey SL600 series integrated circuits, and can be detached from the main rig for portable use. When running 400 watts PEP output from the main rig, thorough shielding of the solid-state exciter, and filtering of all power-supply, microphone and audio-output leads is essential to prevent rf feedback.

A calibrated gain control, R1, was included in the circuit for clipping level adjustments. This is quite straightforward since attenuation varies linearly with the voltage applied to pin 8 of the SL630 microphone amplifier, with zero attenuation at 0.8 volt, and 30-dB attenuation at 1.34 volts. The gain of the pre-amplifier was adjusted, through choice of components, to suit the sensitivity of the microphone. When completed, it was just possible to reach the maximum allowable input of 200 mV to the balanced modulator when speaking loudly.

Some pre-emphasis is provided by C1 and the 500-ohm input resistance of the SL201 speech preamp. The rf gain between the balanced modulator and clipper is preset by R2 so that normal voice peaks just reach limiting level with R1 set for 25 to 30 dB of attenuation.

A monitor scope on the output of the transmitter was used to observe limiting level. The monitor scope was also used in conjunction with the rf gain control, R3, to check the transmitter for linearity, an operation which is made easier by the relatively constant output level provided by the clipper.

It is not always convenient (or desirable) to place the rf gain control immediately after the second filter, and it is important to make sure that the intervening amplifier or mixer stage is linear. You might think that any non-linearity between the balanced modulator and clipper would simply leave the clipper with less work to do, but unfortunately, capacitors in bias and decoupling circuits have a habit of charging up so that stage gain varies with the peak or mean signal level. Therefore, it is advisable to check linearity and make sure there are no amplifier time constants exceeding about 1 millisecond.

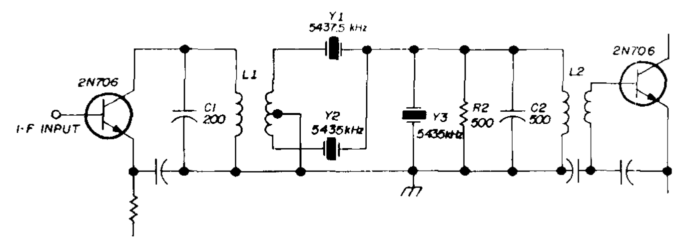

fig. 1. Half-lattice crystal filter using surplus FT243 crystals. L1-C1 and L2-C2 resonate at 5436 kHz. Crystals Y1 and Y3 are selected, as discussed in text.

performance

In use, this system provides a big increase in intelligibility when signals are weak or interference is bad. The operator's voice is always clearly recognizable by those familiar with it, and removal of all clipping produces an average drop of about two S-units. In general, other operators have no idea that speech clipping is in use, and are often quite surprised when they are told. However, the use of more than 20 dB of clipping tends to result in adverse reports.

surplus crystals

The carrier and unwanted sideband rejection of surplus crystal filters, although much inferior to commercial crystal filters, should be adequate for at

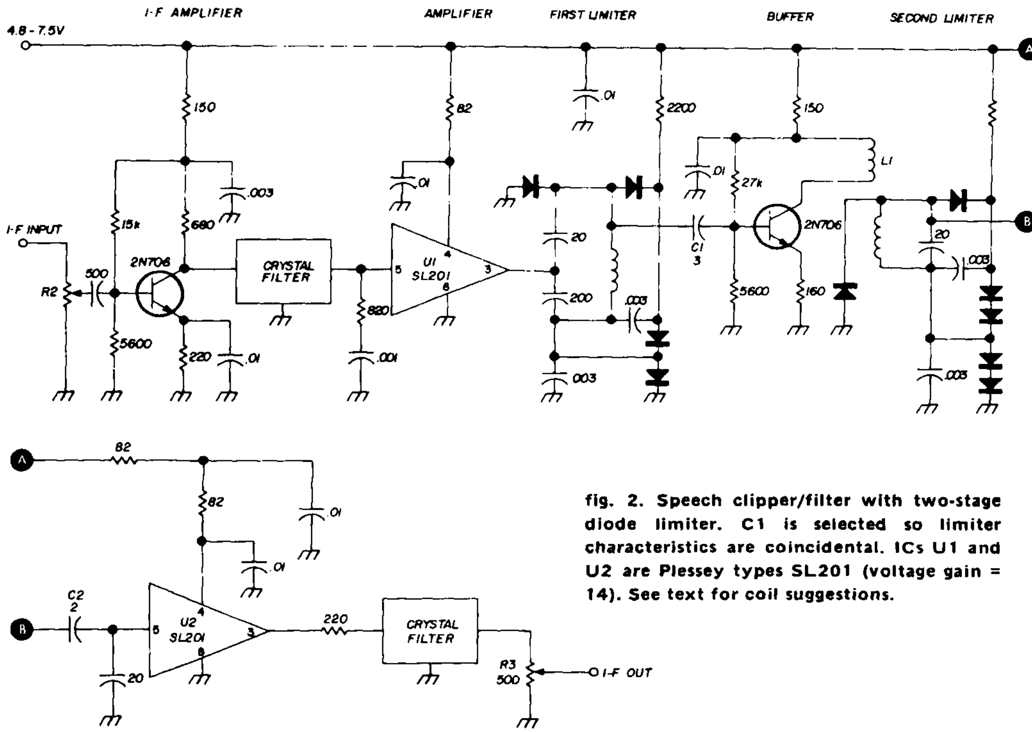

fig. 2. Speech clipper/filter with two-stage diode limiter. C1 is selected so limiter characteristics are coincidental. ICs U1 and U2 are Plessey types SL201 (voltage gain = 14). See text for coil suggestions.

least 12 dB of rf clipping. Surplus crystal filters are not normally suitable for receiving because of the high amplitude of spurious responses. However, the use of two filters for the rf speech-clipping system offered a key to the reception problem, with two separate filters. In the receiving filters the crystals were selected to stagger the spurious responses so that any unwanted frequency passing through one filter were blocked by the other.

alignment

The system shown in **fig. 2** is easily checked for correct operation. A sensitive rf indicator such as that shown in **fig. 4** is connected to the output of the balanced modulator. An audio tone is applied to the microphone input and the audio gain is increased until the rf signal output starts to limit. The audio gain is increased by about 20 dB (with R1) and the transmitter output is checked to make sure the output remains constant while the output of the balanced modulator rises linearly.

Connect the rf indicator to the output of the second filter and increase the rf at this point by increasing the size of capacitor C2. The output of the final should increase proportionately, indi-

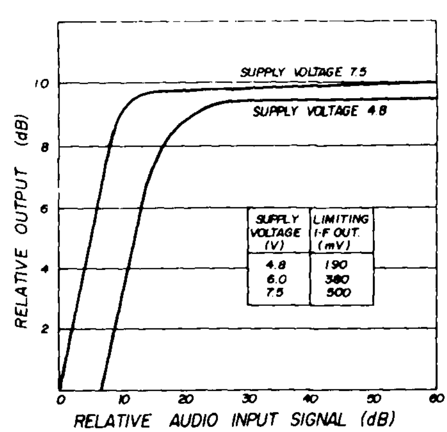

fig. 3. Limiter characteristics of the circuit of fig. 2. Curves represent extreme values of battery voltage for portable operation.

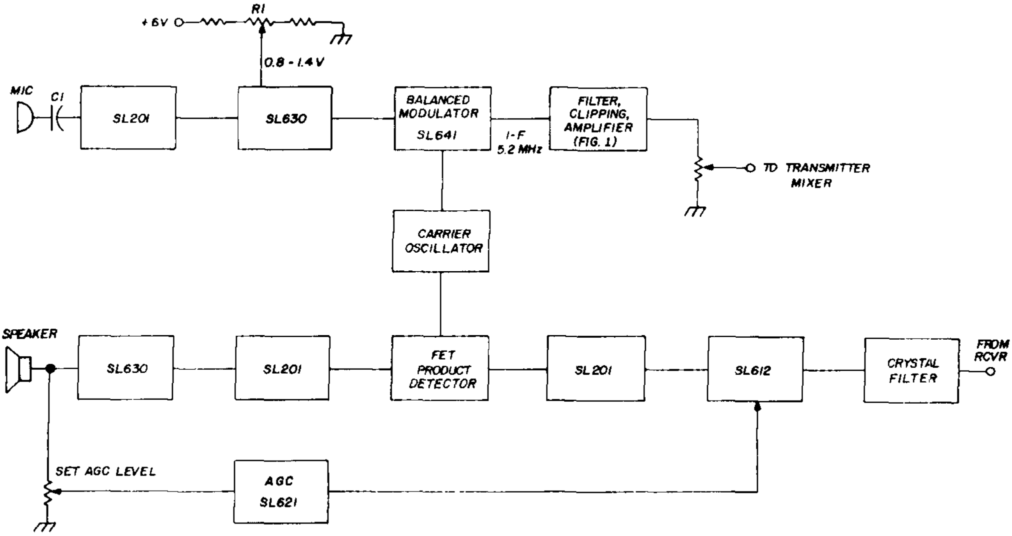

fig. 4. Solid-state ssb transceiver used by author is based on Plessey SL600 series ICs.

cating that there is no non-linearity occuring after the limiter. Alternately, if the rf gain control, R3, is not at maximum, the rf indicator may be placed after R3, and R3 varied instead of C2.

Remove modulation, place an rf short between the input and output of the second filter and check that carrier leakage remains at least 20 dB down, relative to the limiting level. Place an rf short across the input to the second filter and check that there is no rf output; repeat this test with a short across the input of the first filter.

conclusion

For the present, rf clipping is likely to appeal primarily to the serious experimenter who is looking for the ultimate in performance. There are no shortcuts to good performance, and as I noted in last month's article, careful attention to a large number of small details is essential. However, even with simple designs, you can expect some increase in talk power.

In this context, of course, I am talking about the end product, and *not* the means of achieving it. The system of **fig. 2** is not the only possible method. For example, by converting the clipped i-f signal back to audio, you have a device which can be plugged into the microphone lead (such as the Comdel speech processor).

Yet another approach is the modification of commercial gear along the lines used by K6JYO.[2] Each case presents a different problem, but some involve less difficulty than others. Although it is probably not always possible to follow all of the rules mentioned in my first article, particularly in regard to filters, the inclusion of as many points as possible is likely to prove worthwhile.

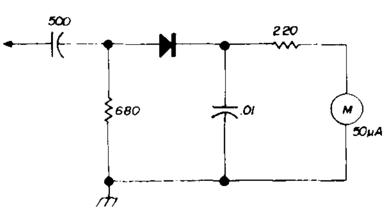

fig. 5. Sensitive rf probe for aligning rf speech clipper.

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ham radio

evaluation of simple dx antennas for 40 and 80 meters

DX is possible on 40 and 80 meters with a variety of antennas designed around practical limitations

Malcolm P. Keown, W5RUB, 5816-113 SW Archer Road, Gainesville, Florida 32601

The introduction of awards by several organizations for recognition of multiple band DX capability have considerably diversified the antenna farm of the serious DX man. Working DX on 10, 15 and 20 meters is no problem using the popular three-band beam and an average station. Things get more difficult, though, as you move inland from the east or west coast. Due to the deterioration of signals as they are propagated over land at low frequencies, amateurs in the interior require antennas more elaborate than a dipole at 25 feet to compete with their colleagues in more advantageous geographical locations.

a proposed solution

A survey of available commercial low-frequency DX antennas revealed that most were an electrical compromise and were also expensive. There was also the problem of compatibility with the present tower-rotator-tribander configuration. I looked to see what could be homebrewed. The results were somewhat discouraging — I didn't have a tuner, a 100-foot tower, 20,000 feet of wire or a big lot. The only simple low-frequency antennas which were useful as DX antennas and could be considered constructionally feasible were the vertical, the horizontal wire beam and the sloped dipole. I developed the following test plan: First, to construct simple wirebeam, ground-plane and sloped-dipole antennas for 40 meters, and to determine which antenna was the most effective for DX. Second, to extend the 40-meter ground plane to about 60 feet to act as quarter-wave radiator on 80 meters and to attempt to use the same antenna as something close to a 5/8-wave vertical on 40. Next, I wanted to determine how well the ground plane worked on 80 and if the 5/8-wave vertical is more effective on 40 than the quarter-wave ground plane. Last, I wanted to construct a sloped-dipole for 80 meters to compare with the ground plane for the same band. I did not try a wire-beam configuration on 80 because of its large size and generally unfavorable comments in the literature.[1]

two element wire beam

The *ARRL Antenna Book* contains a simple two-element wire beam in the chapter on 14-, 21- and 28-MHz antennas.[2] Extending the concept to 40 meters was easy. The original design for the two-element folded dipole beam specified that the radiating elements should be constructed of number-12 wire, spaced 3 to 6 inches. Constructing something of this nature appeared to be a lot of work, so I used regular 300-ohm twin-lead instead. As an effective flat-top configuration required four poles of at least 60 feet, I tried an inverted-vee array using a boom for proper spacing.

The boom was assembled from an old tv mast and two sections of conduit. I installed pulleys on the boom so that the folded dipoles could be pulled out to the proper spacing from the tower. I attached the boom to the tower as follows: I mounted an eyebolt through the center of gravity of the boom and attached an 18 inch threaded rod to the tower at the 55-foot level. The rod was attached with U-bolts, with about 6 inches to the threaded rod extending from the tower. The eyebolt on the boom was then slipped over the threaded rod and secured with two bolts.

Stability of the boom was improved by installation of a wooden beam as shown in **fig. 1.** The length of the wooden beam is dependent on the desired orientation of the boom. The ends of the boom drooped, of course, so I added two eyebolts five feet from either end. Cables attached to the eyebolts were connected to the tower at the 70-foot level. Axial tension was increased by turnbuckles. Once the folded dipoles were in place on the boom and the ends attached to trees, each leg of the dipoles made a 30° angle with the horizontal.

Electrically, the antenna consists of

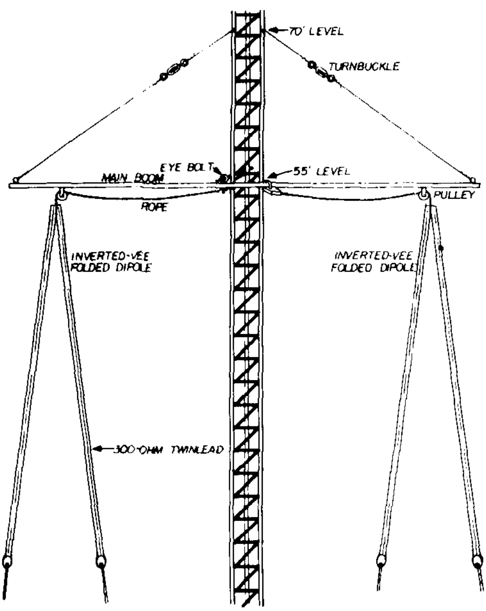
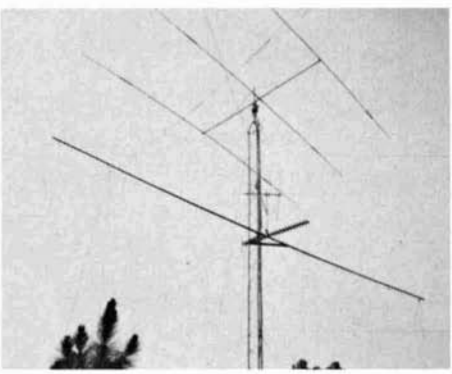


fig. 1. Wire beam for 40 meters uses two inverted-vee folded dipoles made from 300-ohm twinlead. Boom is attached to tower as shown in fig. 2.

two inverted-vee folded dipoles spaced a quarter-wave apart. The length of the feedline to the folded dipole which is the back element of the beam is a quarter-wave longer than the feedline going to the folded dipole which is the front element of the beam. The major radiation lobe is off of the front element. The feed impedance for the original flat-top configuration was 150 ohms. Since I didn't have any 150-ohm twin lead, and since I had no idea what the impedance would be for the modified array, I tried feeding the antenna with 72-ohm twinlead. The swr was very high.

I tried 300-ohm twinlead and the swr was 1.5:1 at the resonant frequency. The dipoles were originally cut for 7150 kHz but the resonant frequency was about 6950 kHz. This was anticipated because inverted vees have lower resonant frequencies than dipoles for the same antenna length. Six inches removed from each end of each dipole moved the resonant frequency to 7050 kHz. The swr was 1.5:1 at 7050 kHz and 3:1 at 7300 kHz, which was acceptable. I speculated



Installation of the folded-dipole inverted-vee antenna at W5RUB. The two folded dipoles are suspended from a boom mounted on the side of the tower at the 55-foot level (see fig. 1).

that possibly the swr measured at the end of the 300-ohm transmission line and the feed point of the antenna were different, but an swr measurement at the feedpoint indicated that they were practically the same. A dpdt relay was installed at the feedpoint so I could reverse the direction of the major radiation lobe.

40 meter ground plane

Next, I built the ground plane. The antenna was made by topping off a 50-foot push-up tv mast with a 21-foot CB whip. The radiating element for 7 MHz is only 33 feet, but the additional height was necessary so the antenna could be used as an 80-meter ground plane in future tests. The mast was guyed only at the ten-foot level during the 7-MHz tests. Unfortunately, the top section of the tv mast and the bottom section of the CB whip were the same diameter so they had to be joined with a section on one-inch inner diameter copper tubing.

Ground losses for vertical radiators can be significantly reduced by using artificial grounds like a ground plane. It is wise to erect an array of this nature as far above ground as possible. One common location for a ground-plane antenna is on the peak of the roof. Many people feel, however, that radials spread out on the roof detract from the house's appearance.

To avoid this problem the 40-meter quarter-wave radials were apprehensively installed on the ceiling of the attic. Not knowing the electromagnetic properties of my roof (asbestos shingles, tar paper and plywood), I anticipated assorted gremlins, but none have been observed. Ten 33-foot radials were attached to the ceiling of the attic spaced 30 to 40 degrees apart – some being bent to fit the available area of 30 x 70 feet. If an open area is available for erection of the ground plane, by all means use it, but if yard space or aesthetic arguments are a problem, the attic provides a good alternative.

Insulating the vertical radiator from the roof was another problem. After some experimentation, the best solution appeared to be a roof saddle with a U-flange and some method of insulating the U-flange from the vertical. The best solution was an ordinary automobile rear shock absorber with rubber inserts on the mounting eyes. Using the shock absorber, a quarter-inch bolt was inserted successively through one side of the U-flange, the lower rubber insert and the other side of the U-flange. A hole slightly larger than a quarter inch was then drilled in both sides of the bottom section of the 50-foot tv mast about six inches from the bottom end (see **fig. 3**). The bottom section was then slipped over the shock absorber and a bolt inserted through one of the previously drilled holes in the mast, through the upper rubber insert of the shock absorber and out the other side of the mast. Thus, once erected, the vertical was insulated from the roof.

If your roof is guaranteed to be a good insulator this ritual is not necessary, and the vertical can simply be attached to the U-flange. A hole was drilled, under a shingle, through the roof and a number-ten wire was inserted to connect the vertical element with the transmission line. Plenty of plastic roof cement was applied to the modified part of the roof. The ground plane was fed with 52-ohm coax (which was also in the attic). No reflected power was observed at the calculated resonant frequency.

the sloped dipole

The sloped dipole configuration consisted of a simple folded dipole attached between one point 47 feet up the tower and another point 47 feet from the base of the tower. Hence, it made an angle of 45° with the horizontal with the major radiation lobe for the antenna in the direction of the slope.

performance comparison

The relative merits of antennas are usually evaluated in terms of field strengths at various angles and distances from the point source of the radiation. However, the major object of this exercise was to work DX with a minimum amount of blood-letting, hence, whichever antenna had the best punch was obviously the best antenna.

The sloped dipole and one major lobe of the beam were oriented towards Europe which is the major source of long-haul DX from Mississippi. Averaged reports received from Europe indicate that the sloped dipole and the ground plane are equally effective. The average report for the two antennas ran about one S-unit higher than the faithful inverted vee at 65 feet. The two-element wire beam generated average reports which were two S-units higher than the inverted vee and one S-unit higher than the sloped dipole or ground plane. Included angles of other than 45° might be tried with the sloped dipole. No optimum angle has been published for 40 meters as far as I know.

The sloped dipole is economical and easy to erect if a 50-foot support is available. This antenna provides some reduction of stateside interference and a worthwhile increase of signal strength on distant propagation paths as compared with an inverted vee or dipole at a comparable height. The ground plane requires very little erection area (if the radials are in the attic) and provides a tremendous reduction in stateside interference. This antenna is a good omnidirectional performer for the DX man who does not have much space or has no

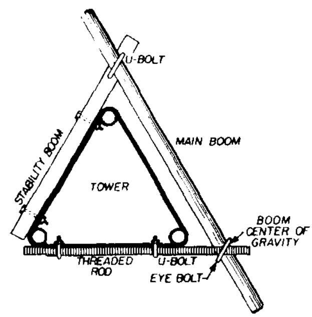


fig. 2. Boom for wire beam is attached to tower with an eyebolt and threaded rod. Short piece of wood is used for stability boom. See text for complete description.

means of raising an antenna to 50 or 60 feet.

The two-element wire beam takes up a lot of room and takes quite a bit of erection effort, but this is a good antenna to have in a pileup. Two of these antennas mounted perpendicularly to each other will provide excellent worldwide coverage on 40 meters. Rejection of signals off the side is good and fair off the back. Stateside interference, of course, increases off the front and this is sometimes a problem if the East Coast is between you and Europe. Many times I have worked Europeans who could only hear me on the wire beam, but I could hear them only on the ground plane.

On the other hand, the beam is a great contest antenna. I have often found as a stateside contest winds down on the higher frequencies and you are forced to go to the lower frequencies in search of contacts, it is sometimes difficult to work the East Coast until a couple of hours after sunset. This is because much activity is concentrated in the 1-2-3-8 and upper 4 call areas and the signals are so strong within this area that signals from the outside are buried in the interference. The beam is a very effective aid for getting the attention of the East Coast community both in the hours near sunset and later in the evening.

dual-band antenna

As previously mentioned, the 40-meter

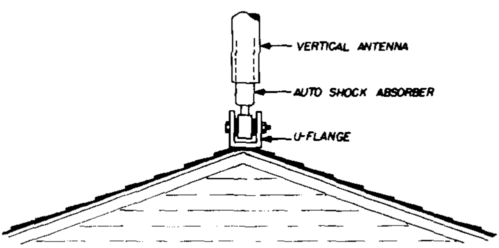


fig. 3. Vertical antenna is insulated from roof with an old automobile shock absorber with a rubber bushing.

ground plane was constructed so it could be converted to an 80-meter ground plane and something close to a 5/8-wave vertical for 40 simply by extending the push-up tv mast. The 5/8-wave on 40 is significant because at that length maximum low-angle radiation occurs.[3] The vertical was raised to 60 feet and the antenna was guyed at 10 and 28 feet. The guys were attached to eyebolts in the roof some 15 feet from the base of the antenna. The eyebolts were anchored in the rafters and plastic roof cement applied around the eyebolt on the roof surface. The vertical was insulated from the guy wires by egg insulators which were placed in each guy wire about six inches from the vertical. Six feet of the fourth section and five feet of the fifth section of the tv mast and the 21-foot CB whip were left unguyed. The 32 feet of unguyed antenna have presented no stability problems and the antenna has sustained gusts to 70 mph without damage. The lower three sections of the tv mast were extended to the maximum attainable length which came to about 28 feet total, but the fourth and fifth sections were extended to only six and five feet respectively in order to keep down the number of guy wires.

A dpdt relay was installed at the base of the ground plane (in the attic) to switch between the 40- and 80-meter matching circuits. The 40-meter configuration required a simple L-network to match the 5/8-wave vertical to the 52-ohm coax.[4] On 80 a small loading coil was necessary because the vertical was electrically short at 3.5 MHz.

Operation of the extended vertical on 40 indicated that no increased effectiveness was apparent as compared with the quarter-wave vertical. This conclusion was based on comparison with the 40-meter sloped dipole. Therefore, if you don't plan to use the vertical as a ground plane on 80, the extended vertical is not worth the effort.

ground system

Contrary to popular opinion, radials shorter than one-quarterwave can be effectively used in a ground plane antenna configuration provided the vertical radiator is *at least* 1/5 wavelength or longer. No problem was experienced in loading the 80-meter vertical against the 40-meter ground system using the L-network. I worked a lot of DX on 80 using this configuration. Some weeks later, fifteen quarter-wave radials were added to the system making it a true 80-meter ground plane. This made a total of fifteen 80-meter radials and ten 40-meter radials. Reports received from DX stations were somewhat better, but I am not sure whether the improvement was due to increasing the number of radials or adding longer radials. It is interesting to note in Lee's work[3] that increasing the length of radials from 1/8 to 1/4 wavelength increases the unattenuated field for a vertical radiator at one mile by 5 millivolts per mile while doubling the number of quarterwave radials from 15 to 30 provides a 12 millivolt per mile increase. Agreed, we are comparing apples and oranges, but it is conceivable that an amateur who is cramped for space may do nearly as well with thirty 1/8-wavelength radials as an amateur with fifteen quarterwave radials.

Quarterwave radials on 80 take up a lot of room. My attic, only 30 by 70 feet, necessitated placing the 80-meter radials on top of the roof. The 80-meter radial system was connected to the transmission line and the 40-meter ground system by an additional number ten wire through the roof. Fortunately, during DX season on 80 the nights are long and the days short; so if you are clever with the deployment of radials, evening guests will never be aware of the conglomeration of wire hanging above their heads. My gracious wife allows me to lay out my radials anytime after we go off of daylight savings time!

80-meter sloped dipole

Dalton recommends using a 100-foot tower to support a sloping dipole cut for 3650 kHz. This configuration yields an included angle of 52° between the antenna and ground, which he says is optimum for DX. Unfortunately, my tower is only 70 feet, and by using the same scheme, the included angle would decrease to 33°, obviously unacceptable. Practicality dictated a compromise. I decreased the included angle from 52° to 45° and raised the resonant frequency to 3800 kHz. The length of the antenna now was 123 feet. I stretched 100 feet of it from the ground to the top of the tower (at a 45 degree angle) and dangled the remaining 23 feet down the side of the tower (secured at the 47-foot level). This array worked fine at 3800 kHz but the swr was very high 100 kHz from the resonant frequency. It is well known that the bandwidth characteristics of a folded dipole can be improved by placing shorting straps at a distance from the center of the dipole which is equal to the velocity factor of the twin lead times half the length of the dipole.[5] This worked out to be about 7.6 feet from the ends of the antenna on 3800 kHz. Using this configuration, the swr was 2.5 to 1 at the low edge of the band.

80-meter antenna comparison

Reports received from Europe and Oceania indicated that the ground plane has an edge over the sloped dipole although I noticed no difference on the receiving end. The sloping dipole may be equal to, or more effective than, the ground plane if it were erected correctly utilizing the 100-foot tower.

recommendations

If supporting structures of 100 feet are not available, try the ground plane for an effective DX antenna. If a 100-foot structure is available, try a sloping dipole first, since the erection effort is small compared with the effort expended in putting up a ground-plane radiator with its associated radial system.

Possibly neither configuration is feasible; if this is the case, the modified sloping dipole is preferred over an inverted vee at 60 feet or less.

If 40- and 80-meter DX capability is required and only space for one antenna is available, the 80-meter ground plane/40-meter 5/8-wave vertical will deliver the DX even if there is no room for 80-meter radials. If no 80-meter radials are used, loading against earth ground may improve results. This can be accomplished by running a number 6 or 8 wire from the common junction of the 40-meter radial system and the shield of the transmission line to the nearest earth ground.

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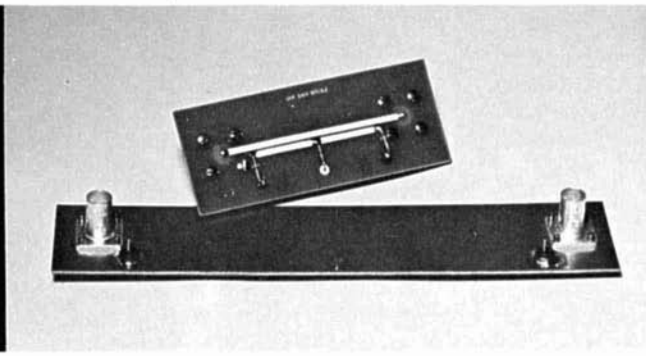
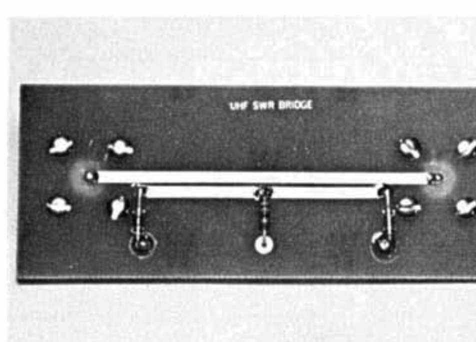


vhf and uhf micro-strip monimatch swr indicator

These two micro-strip monimatch swr indicators cover the complete frequency range from 40 to 600 MHz

Arthur R. Hall, W4CGC, Fairfax, Virginia 22030

This article describes the application of the micro-strip transmission line technique to an old favorite, in-line swr indicator, the *monimatch*[1]. Two models are described, a vhf model to cover the frequencies, 40 to 200 MHz, and a uhf model to cover the frequencies, 200 to 600 MHz. The original monimatch and



Construction details of the micro-strip uhf pickup line.

most of its decendents have been widely accepted down through the years as one of the easiest to build and most reliable of the in-line swr indicators.

As good as most of the monimatch designs are, they do have an upper frequency limit that prevents most of them from being reliable at vhf frequencies, and none seem to function reliably at uhf.

Actually, I tested the idea of adapting the micro-strip technique to the original monimatch design six years ago with a number of versions being built with good results. Since then, activity on the vhf and uhf bands has mushroomed tremendously, particularly on two-meter fm.

Most of the newer two-meter transceivers are all solid state and can be very unforgiving if they are used with an antenna that is not properly matched to the output circuit. The two micro-strip swr indicators described here will provide the means for reliably measuring swr and relative power in the region of 40 to 600 MHz.

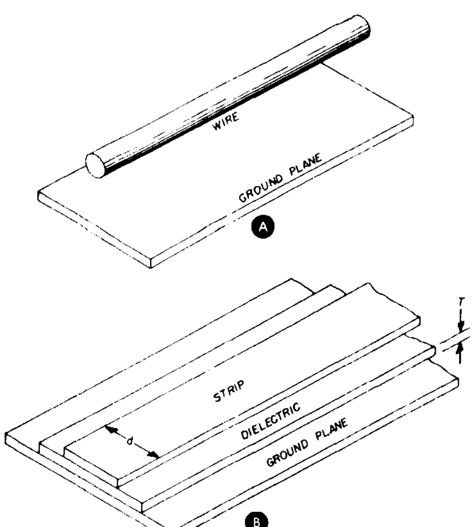


fig. 1. Cross section of the micro-strip transmission line system. Impedance is determined by the width of the strip, thickness of the dielectric and dielectric constant.

theory of operation

Basically, a micro-strip transmission line is an unbalanced, constant impedance conductor analogus to a wire above a ground plane with a dielectric in between as depicted in **fig. 1.** The characteristic impedance is dependent upon the line width, the distance of the line above the ground plane and the dielectric constant of the printed circuit material. (G-10 material was chosen because of its low loss factor at these frequencies, its wide availability and its reasonable cost.)

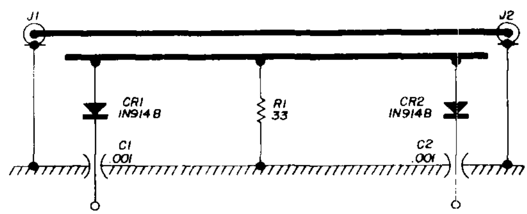
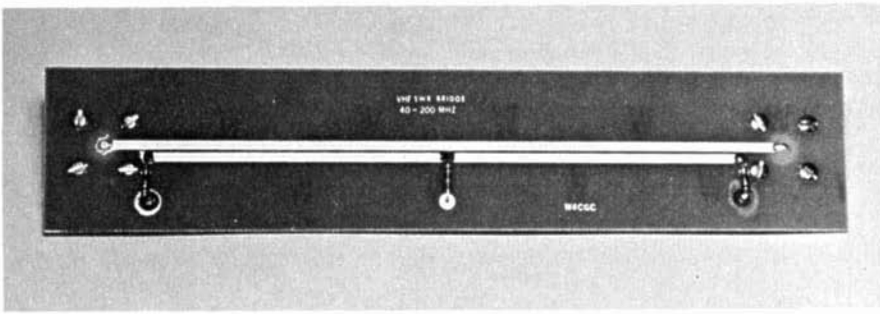


fig. 2. Circuit of the printed-circuit, micro-strip swr pickup line. C1 and C2 are Aerovox type 5601 discoidal feedthrough capacitors.

A second line running parallel to the conductor line is center tapped to ground through a 33-ohm carbon resistor. High-frequency switching diodes rectify the induced rf voltage (on the pick-up line) which is then filtered by uhf feedthrough capacitors (see **fig. 2**). The pick-up line is coupled both inductively and capacitively to the main conductor line as shown in **fig. 3.** One half of the pick-up line samples the power flowing in the center line in one direction between one diode and the terminating resistance, the other half of the line senses the rf power flowing in the opposite direction between the other diode and the terminating resistance.

The upper frequency limit of these swr indicators is dependent not upon the microstrip center line, but upon the pick-up line. *It appears that the inductance* of the pick-up line from one end, to the terminating resistance and its distributed capacitance to the ground plane, is the determining factor. One theory of limitation is that the pick-up line consists of multiple series inductances with multiple distributed capacitances to ground as illustrated in **fig. 4A.**

Another word of caution is in order at this point. Even though a micro-strip transmission line is designed very carefully to maintain a constant 50-ohm impedance, discontinuities may occur at each end of the micro-strip line where it connects to the coaxial connectors. It is for this reason that I have chosen to mount the two coaxial connectors directly to the micro-strip line. This method of attachment will maintain a near constant 50-ohm impedance between the coaxial connector and the line.



Closeup of the vhf circuit board shows the micro-strip pickup line, diodes and resistors.

Fig. 4A can be approximated as a simple pi network as in **fig. 4B.** Since a pi network is a low pass network, the cut-off frequency would then be a function of L2, C1 and C2. Laboratory measurements of these values correlate closely to the upper frequency limit of reliability of both the vhf and uhf swr indicators.

One important factor often overlooked in the design and fabrication of in-line swr indicators is that it *must* maintain a constant impedance, exactly the same as the characteristic impedance of the line it is measuring. If it does not, the in-line swr indicator can, itself, cause a mismatch even though it shows a perfect match between the generator and the load.

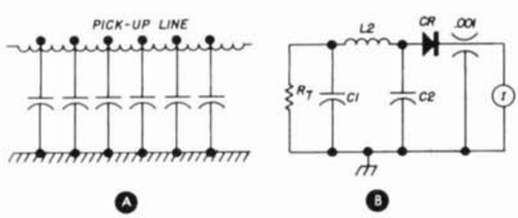


fig. 4. The distributed capacitance and inductance of the pickup line (A) limit the upper frequency limit of the swr indicator because they form the pi-network in (B).

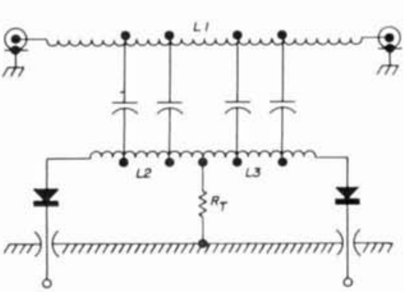


fig. 3. Upper operating frequency of swr pickup line is limited by the distributed capacitance and inductance of the line (see fig. 4).

Laboratory measurements of the inherent swr of the two micro-strips using a General Radio Model 900LB precision slotted line, is shown in **fig. 5.** The results are quite acceptable throughout the useful range of each indicator.

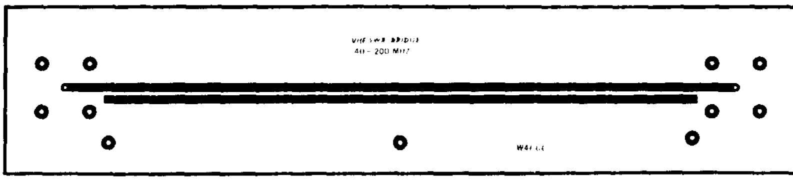
Another important factor in the design and fabrication of a reliable swr indicator is that it must be perfectly balanced between the center of both the conductor line and the pick-up line – one side must be a mirror image of the other. This is most easily achieved using the micro-strip technique.

construction details

As was pointed out earlier, accuracy of line width, thickness of dielectric and

they are actually identical except for the line lengths.*

After the boards have been photo etched and trimmed to size they should be drilled using very sharp drills of proper size at high speed. The two holes that the



Half-size layout of the vhf printed-circuit board.

dielectric constant must be maintained to close tolerance for best results. Experiments show that when using G-10 epoxy-glass material of 1/16 inch thickness, a line width of .094" ± .002" will provide the necessary 50-ohm characteristic impedance. The copper-clad should be no thicker than 1-ounce material to minimize undercutting during the etching process. Both the vhf and uhf indicators follow the same construction technique;

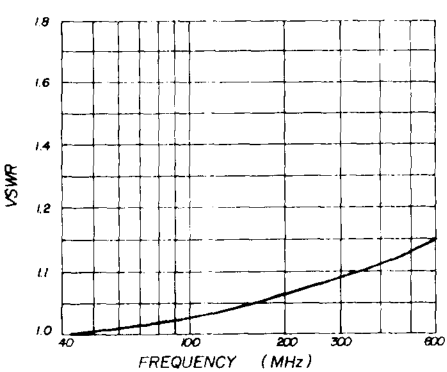


fig. 5. Insertion vswr of the two microstrip swr indicators.

*Completely cut and drilled G-10 epoxy glass boards and complete kits are available from TRI-COM, Inc., 12216 Parklawn Drive, Rockville, Maryland 20852. Either the vhf or the uhf board is $7.00 and either complete kit is $13.50. Both prices include air-mail first-class shipping charges. Maryland residents, please include the 4% sales tax. Rush orders can be placed by telephone to 301-770-5585.

center conductors of the connectors pass through should be beveled on the ground plane side so that a short does not occur. This beveling can be done with a counter-sink tool, or better still, a sheet metal 0.312-inch drill can be used with an appropriate pilot drill. The idea is to just go deep enough to remove the copper without cutting into the G-10 dielectric.

The two coaxial connectors are standard UG-290/U BNC units which have been modified for micro-strip use by being turned down on a lathe (see fig. 6). The face of the square flange should be faced off approximately .010" to remove any imperfections that might exist in the connector. When facing off the flange, continue all the way through to the center conductor, removing the shoulder boss. Then cut off the center conductor so that .080" remains. This will yield an rf connector that can be directly attached to the micro-strip printed-circuit board. Attach the modified UG-290/U connectors to the micro-strip board with ¼" long, 3-48 machine screws.

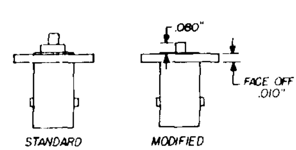
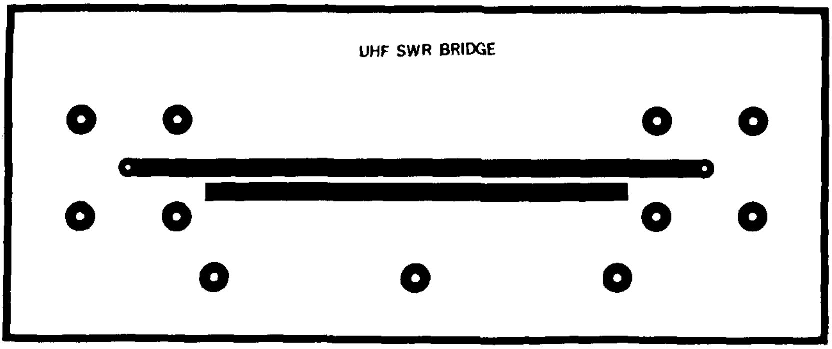


fig. 6. Modifying the UG-290/U BNC connector for use with the micro-strip line.

The next step is to sweat solder the two feedthrough capacitors in place. This operation should be done with a heavy duty soldering iron with a large tip. Make sure solder flows evenly all around the capacitor flange.

The soldering of the two center conductors of the modified UG290/U connectors should be done with the same heavy-duty iron. Use solder sparingly at these two points. This operation should be done quickly to reduce chances of peeling the foil from the board material. This is not a critical point but care should be exercised.

Lastly, install the two diodes and terminating resistor. No special care is needed here except to keep these components perpendicular to the pick-up line.



Full-size layout of the uhf micro-strip swr bridge.

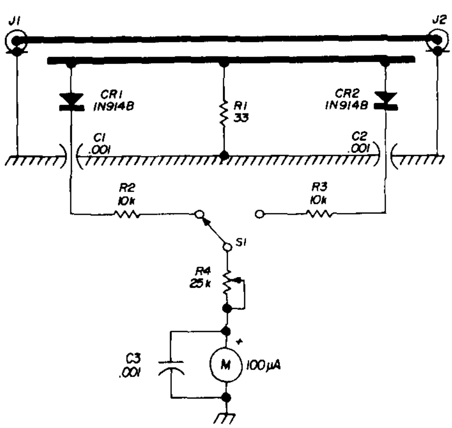


fig. 7. Complete circuit of the swr indicator, *including the micro-strip pickup line and meter.*

testing and calibration

The swr indicators can be tested for accuracy by inserting each indicator between an appropriate transmitter of 1-watt or more and a known 50-ohm *non-reactive* dummy load such as a Bird Termaline, etc. Most homebrew dummy loads, when used at vhf and uhf, will be reactive, thus producing a reflected reading on the indicators. In fact, these indicators should be accurate enough to indicate how good your dummy load is at these frequencies.

I have not made provision for enclosing the indicators in a mini-box, choosing to let the builder make his own decisions in this regard. The only word of caution here would be to keep at least a ½-inch clearance between the rear plane of the micro-strip and the enclosure. A suggested schematic including balancing resistors, forward-reflected switch, sensitivity potentiometer and meter is provided in **fig. 7.**

Acknowledgement goes to Mr. David W. Reynolds, W3QKR, for assisting with the circuit analysis, and to Mr. John Gregory, W3ATE, for his assistance in providing the photo-copy work.

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monitor receiver for RTTY autostart

A stable receiver for continuous monitoring of the RTTY autostart nets

Bert Kelley, K4EEU, Tampa, Florida 33609

This receiver was built to monitor the fixed-frequency autostart RTTY nets on the 80-, 40- and 20-meter amateur bands. If one RTTY station wants to leave a message for another, he sends it at a time when the band is expected to be open between the two stations. All of the stations in the net usually leave their equipment on 24 hours a day, and everybody copies all the messages transmitted. Of course, the addressee of a specific message doesn't have to be in attendance to receive a message. When 14.075 MHz is open, copy is excellent and results in a nationwide intercom system linking RTTY enthusiasts. Other nets on other bands are used for shorter distances.

overview

The equipment required for this type of operation is not elaborate. Many stations use ST-4 or ST-6 autostart demodu-

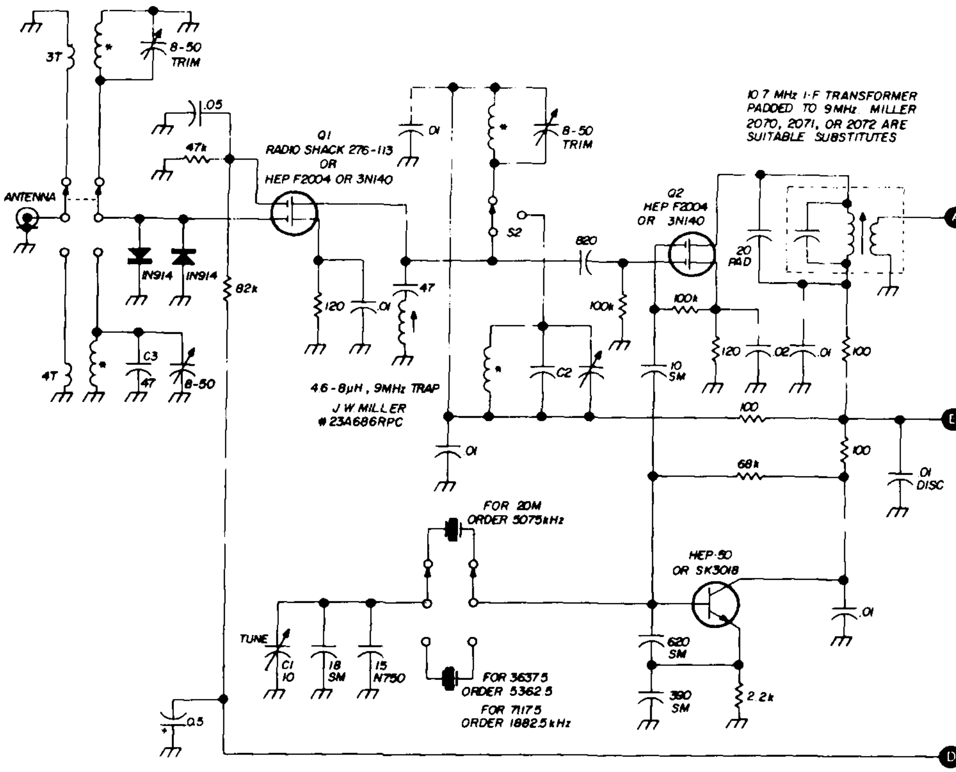


fig. 1. The fixed frequency monitoring receiver. C1 should be a panel-mounted variable. All the tuned circuits with the asterisks use the same inductance, all wound on Amidon T-50-2 forms. For 80 meters, the coils are 40 inches of no. 30 enameled wire; for 40 meters they are 29 inches of no. 26 enameled wire; and for 20 meters they are 13 inches of no. 24 enameled wire. Capacitors C2 and C3 are about 47 pF for 80 meters and are omitted on all other bands. Y1 is 5075 kHz for 20 meters, Y2 is 5362.5 kHz for 3637.5 kHz monitoring and 1882.5 kHz for monitoring 7117.5 kHz.

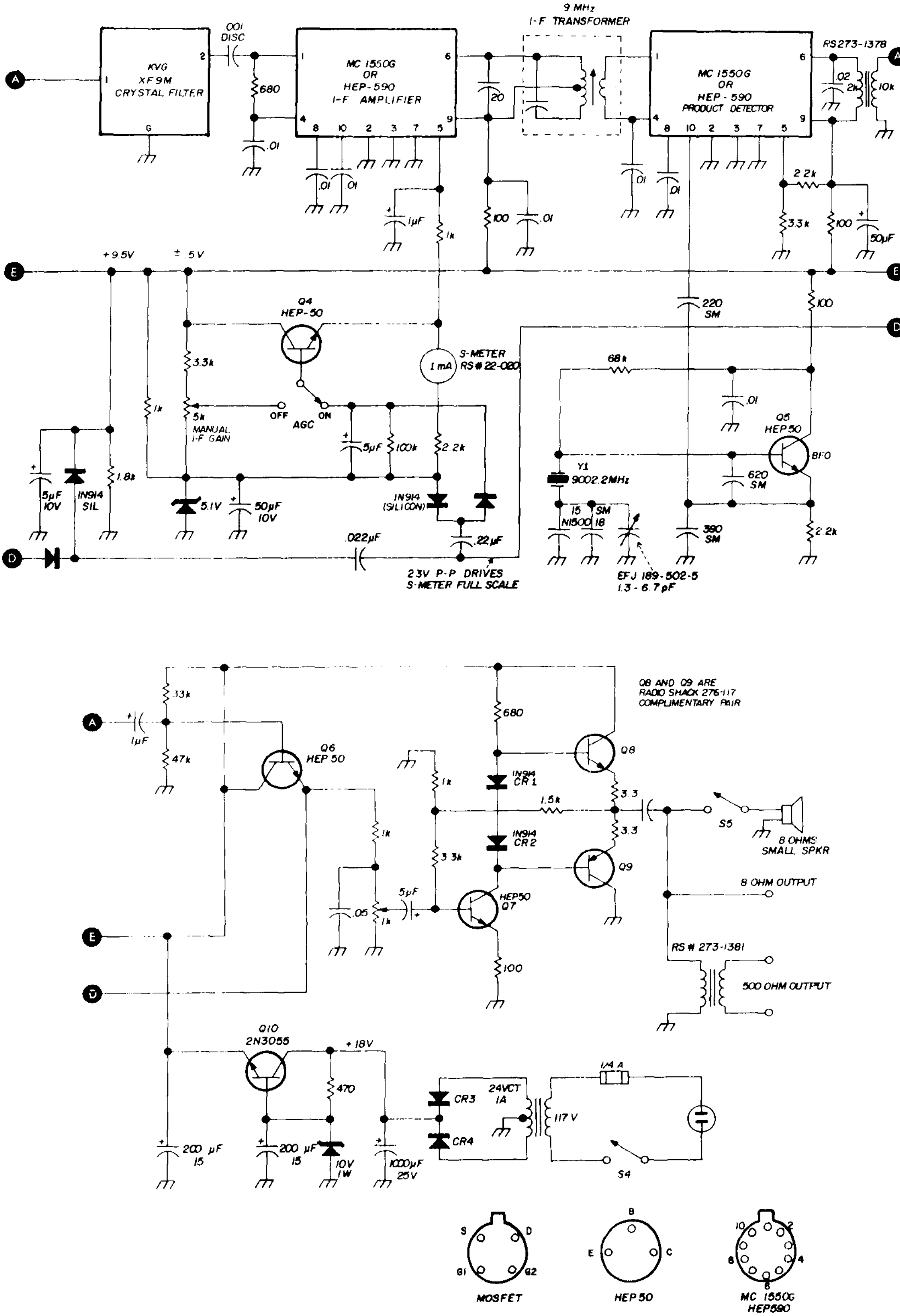
lators and Model 15 or Model 28 teleprinters.[1] Since only 170-Hz shift is used, it is necessary to operate close to the nominal frequency. Frequency precision and accuracy, therefore, are the most demanding requirements. Many stations have secondary-frequency standards and use quality amateur receivers with crystal-locked permeability-tuned oscillators.

A receiver that is to compete with such equipment must have good selectivity, sensitivity and *excellent* stability. The best way to get these features is to solid-state the old reliable superheterodyne circuit.

stability

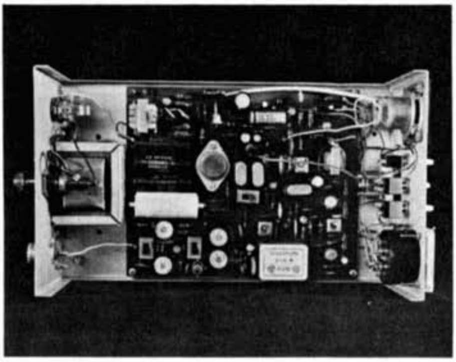
The basic receiver design may be modified to fit other applications. The KVG filter is available in bandwidths that would be suitable for ssb, slow-scan television, 850-Hz shift RTTY or WWV reception. The front panel control provides a plus or minus 1-kHz tuning range. If a greater range is needed, a transistor vfo might be considered at a trade-off in stability.[2]

Crystal control is recommended because, properly compensated, the frequency drift of this receiver is less than 10 Hertz for a 20° ambient temperature change. To get this kind of stability, it is necessary to use a reasonably-priced high-accuracy crystal, specified for a 32-pF circuit capacitance and compensated for any remaining drift with properly


KVG
XF 9M
CRYSTAL FILTER
.001
DISC
680
.01
MC 1550G
OR
HEP-590
I-F AMPLIFIER
.01
.01
20
100
.01
1μF
1k
9 MHz
I-F TRANSFORMER
MC 1550G
OR
HEP-590
PRODUCT DETECTOR
RS273-1378
.02
2k
10k
2.2k
3.3k
100
50μF
+9.5V
± .5V
Q4
HEP-50
3.3k
5k
MANUAL
I-F GAIN
OFF
AGC
ON
5μF
100k
S-METER
RS # 22-020
1 mA
2.2k
1k
1.8k
5μF
10V
IN914
SIL
5.1V
50μF
10V
IN914
(SILICON)
.022μF
22μF
2.3V P-P DRIVES
S-METER FULL SCALE
220
SM
100
68k
.01
Q5
HEP 50
BFO
Y1
9002.2MHz
620
SM
15
N1500
SM
18
390
SM
2.2k
EFJ 189-502-5
1.3 - 6.7 pF
33k
1μF
47k
Q6
HEP 50
680
Q8 AND Q9 ARE
RADIO SHACK 276-117
COMPLIMENTARY PAIR
Q8
1k
IN914
CR1
1.5k
3.3
3.3
S5
8 OHMS
SMALL SPKR
IN914
CR2
Q9
1k
3.3k
.05
5μF
1k
HEP50
Q7
8 OHM OUTPUT
100
RS # 273-1381
500 OHM OUTPUT
Q10
2N3055
+18V
470
200 μF
15
200 μF
15
10V
1W
1000μF
25V
CR3
CR4
24VCT
1A
117 V
1/4 A
S4
S
D
G1
G2
MOSFET
B
E
C
HEP 50
10
2
8
4
6
MC 1550G
HEP590

selected N750 temperature-compensating capacitors. I used an International Crystal HA type crystal in my unit.*

The oscillator is a modified Colpitts circuit, and the large silver-mica capacitors effectively disconnect the crystal from the transistor junctions. The more capacitance that is used here, the better (up to the point where oscillation ceases).



Inside of the receiver. Antenna input is at lower left, audio output is at upper left. Bfo crystal is at right center. The two crystals mounted adjacent to one another are for 80 and 20 meters.

As might be expected, the 9002.2-kHz bfo crystal is the most temperature-sensitive component. While temperature compensating various crystals, I found that the ordinary garden-variety HC-6/U crystals drifted about five times more than the International Crystal HA type recommended here.

devices

After constructing a few receiver front ends using transistors and integrated circuits, it was a pleasant surprise to find how well the mosfet performed.[3] Forget the broadcast-band rejection filter and the attenuators in the antenna leads; mosfets provide plenty of gain without feedback. Toroid coils in the rf stage help, too.

The MC1550G IC is an inexpensive three-transistor array that works well as an agc controlled i-f amplifier and makes a terrific product detector.† The HEP590 is a similar device.[4]

agc

The audio-derived automatic gain control for the mosfet front end and IC i-f requires two different voltage levels and polarities. The mosfet requires a small negative voltage for agc, but the MC1550G agc becomes effective only after voltage rises to 5.1 volts.

After the voltage gain provided by the 2k:10k transformer, and impedance matching in transistor Q6, the peak-to-peak voltage at the emitter measures about 2.3 volts maximum, and averages about 1 volt. This ac voltage is rectified and charges a 5-μF capacitor, the ground end of which is connected to a 5.1-volt reference.

Another emitter follower, Q4, is used to give high-impedance input and low-impedance output. The emitter voltage follows the charge across the small capacitor and is used to drive the s-meter and provide agc to pin 5 of the i-f amplifier. Rf agc becomes effective after about two S-units.

audio

There is a variety of audio amplifier ICs available, but they are not recommended for this receiver because their high gain is not needed. Also, they usually will not stand sustained audio overloads. The transistor audio amplifier stages used in this receiver have the right amount of gain, are very rugged, and cost less than the IC. If 600-ohm output is not needed, no transformer is required.

construction

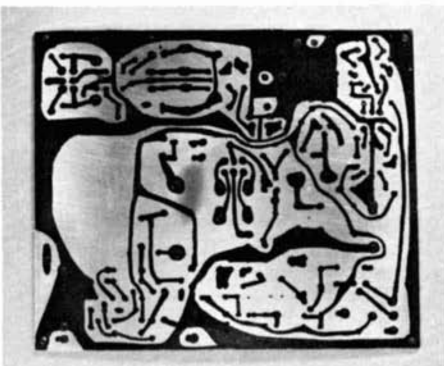
The 2N3055 used in the power supply is a rugged transistor often used in com-

*International Crystal Manufacturing Company, 10 North Lee, Oklahoma City, Oklahoma 73102. Write to them for their complete catalog with details on ordering high-accuracy crystals for your specific frequency and application.

†Both Hal Devices and Circuit Specialists stock them.

mercial equipment. You can find it surplus for less than $2. I recommend that you build the power supply and check it out first. When building the receiver, work backwards from the audio stages, checking stages as you go.

The photos show the receiver built on a 6-7/8 by 5½-inch circuit board mounted on spacers and inside a 10 by 6 by 3½-inch minibox. The board is laid out in a semicircle around the large electrolytic capacitors.



Printed-circuit board before etching. Antenna input is at top left, with mixer to right, and filter (large black area). Top to bottom on the right are the i-f, product detector and agc. Audio is at lower left.

In these days of kits and commercial equipment, it takes a special breed of amateur to build his own receiver. If you are a builder, and if you need a really good monitoring receiver, build this one. You'll like it.

references

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4. Motorola Application Note AN-247, Motorola Semiconductors, Phoenix, Arizona.

ham radio

building
Motran and Motrac
channel elements

Building simple plug-in crystal board replacements for the hard-to-find Motran and Motrac channel elements

Thomas McLaughlin, WB4NEX, St. Petersburg, Florida 33704

Finding channel elements for Motorola's popular Motrac and Motran fm transceivers can often be a problem. Getting crystals for these rigs is relatively easy, but channel elements are both scarce and expensive. You can get around the problem, however, by building your own receiver and transmit elements.

It is fairly easy to make an oscillator unit on a perf board and press it into service as a channel element by connecting it to the proper pins on the radio circuit board. The general circuit of a receiver element, fig. 2, consists of a crystal oscillator and resistor-diode com-

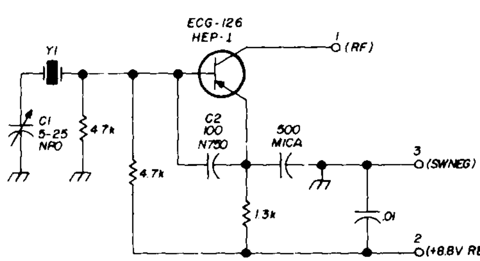

fig. 1. Simple receiver oscillator circuit. Precision grade crystals help to eliminate the need for the temperature-compensating components in fig. 1.

pensating network to correct for the effects of temperature change on the crystal. The circuitry for the transmitter element is the same except for the changes in the collector circuit as noted later. If you need a receiver element, build the circuit in **fig. 1** and order a *precision grade* crystal.

crystals

I have made several elements and ordered crystals from Sentry cut for circuit number 5 in their catalog. Each has tuned to frequency without prob-

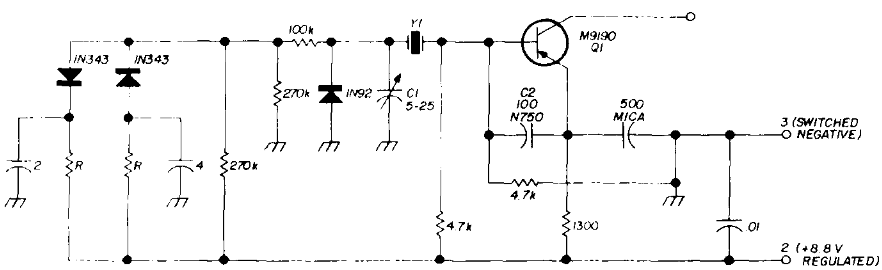

fig. 2. Basic receiver channel element circuit. The text explains about the resistors marked R.

lems. I found that a precision grade crystal will eliminate most drift problems even though the homebrew element does not have the compensating network. You will certainly stay within 0.001%, which is close enough for amateur work. This is exactly the same as buying a used element and installing the crystal in it yourself without adjusting the precision 1% resistors in the compensating network (**fig. 1**, labeled R) to eliminate oscillator drift with temperature change. These resistors may be anywhere from 9k to 15k depending on the particular characteristics of the crystal involved, and it really isn't worth the time it would take to find the correct values unless you have access to some expensive test equipment.

If you need a transmitter element, build the circuit in **fig. 3.** This is exactly the same as **fig.** 2 except the collector of Q1 is grounded to pin 3 and the rf is taken directly from the emitter. **Fig. 4** is a guide to finding the proper pins.

If you have a used element or home-brew oscillator which won't tune on frequency, pad C1 with a 10-pF 10% NPO ceramic disc. This will work if the frequency is too high. If the crystal is too low in frequency or if no amount of padding helps, change C2 from 100-pF N750 to 75-pF N750 disc ceramic. This decreases the circuit capacitance enough to raise the frequency a little. Note that the new style single oscillator elements have holes in the circuit board so you can pad C1 if necessary. The older dual oscillator elements don't have this provision, but C1 is a 5 to 25 pF NPO unit so it shouldn't need padding. The new elements have a 3.5 to 14 pF N300 unit for C1 which accounts for the padding provision. Other than this slight difference, the circuitry for both the old and new types is the same.

If you have regular elements and are ordering crystals, it is best to get precision grade crystals cut for the particular model of element you are using. **Table 1** gives the proper designation for many elements. If you are ordering crystals for home-built elements, I would advise specifying them to be cut for element model

table 1. Motorola channel element designation.

New style single-frequency element		
	0.0005%	**0.0002%**
transmitter	TLN 1083	TLN 1087
receiver	TLN 1081	TLN 1086
Old style two-frequency element (all are 0.0005%)		
	single freq	**dual freq**
transmitter	TLN 1024	TLN 1025
receiver	TLN 1020	TLN 1021

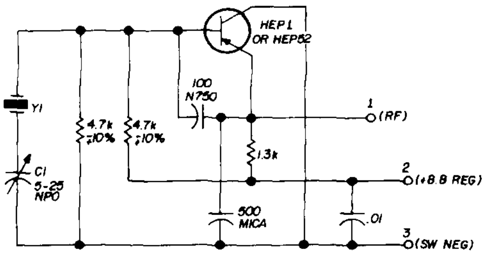

fig. 3. Transmitter oscillator circuit.

TLN1081, or if you buy from Sentry, specify circuit 5 in their catalog. The crystals should tune without any problems in the homebrew circuits.

The ersatz elements I made are all on perf boards, and I mounted them in the rigs using the pins from an old octal socket — they just fit on the male pins on the radio circuit board. If you're desperate or just don't care, you can even solder the element right into the radio.

transistors

The transistors in the standard elements are Motorola type M9190, a house number only. Since this is a PNP rf type, HEP-1 works well, but it is no longer being produced and is not available at all distributors. HEP-52 works well in transmitter elements, but the output is a little too low for use in a receiver element whose output is run right into a tripler stage, unlike the transmitter exciter. Sylvania ECG-126 is also a good bet for either receive or transmit elements because these are fairly high-output devices.

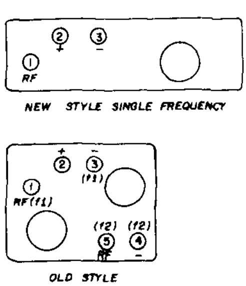

fig. 4. Bottom view of old and new Motorola channel elements shows pin designations.

There are countless other types which will work; even some audio types will have high enough output to be used.

If you have one of the older two-frequency elements which has only one oscillator board in the can, it can easily be converted into a functioning two-frequency element if you have a 5 to 25 pF NPO trimmer and oscillator board out of an unused element. Likewise, any receiver element can be changed to a transmitter element and vice versa. This only requires connecting the wire from the rf terminal to the proper transistor lead on the board and either grounding or ungrounding the collector as the situation dictates (see fig.

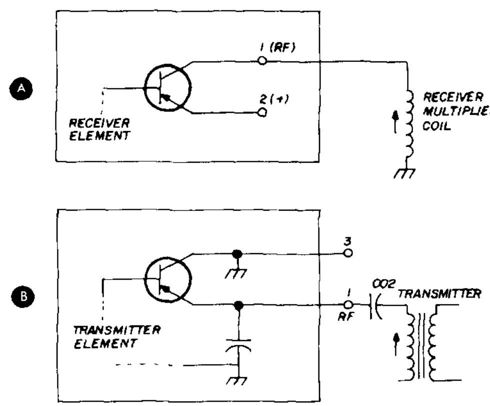

fig. 5. Difference in the hookup arrangements for a receiver element (A) and a transmitter element (B).

5.). If you haven't noticed, receiver and transmitter elements are not interchangeable.

birdies

Another characteristic of Motracs and Motrans is that the i-f crystal is usually 8.455 MHz, Motorola part G09. If the receiver is set up to receive on 146.94 MHz an intermod type birdie will appear in the receiver. This condition can easily be eliminated by changing the i-f crystal to 7.545 MHz, Motorola part G11, available either from Motorola, Sentry or International.

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monitoring oscillator

The step-by-step evolution and modernization of a very handy station accessory

R. W. Gunderson, W2JIO, 980 Waring Avenue, Bronx, New York 10469

However sophisticated some of our modern transceivers may be, sometimes they can benefit from some old tricks. The monitoring oscillator, for instance, has been around for decades but it still can serve very useful functions. I feel the modernization of this handy gadget might be of interest to old-timers who used one in the 1930's, to newcomers who probably never heard of the critter and to experimenters who might enjoy sharing my adventures in adapting modern components to an old circuit.

When you operate your transceiver on CW, you have a so-called side-tone oscillator to monitor your sending. However, you can't monitor your signal as it sounds to the distant station unless you have an additional receiver. If something goes awry while you're on the air — your final takes off, your frequency shifts suddenly or the note goes sour — you will never know about the trouble until the operator at the other end of the circuit tells you that something has happened. Further, the regulations specify that the frequency of the station must be measured by external means other than the station receiver. When you speak of frequency measurements, you mean the operating frequency, of course, and the *quality* of the emitted signal — since its frequency will be shifted if ripple appears in the output or if the stage takes off due to amplifier instability or loss of excitation.

The side-tone oscillator operates at just one frequency (usually about 1 kHz). I like to shift the frequency of the monitoring oscillator as I operate to break the monotony. Of course, this is impossible with the built-in side-tone generator. Rf-actuated monitors which derive their dc power from the radiated energy from the transmitting antenna are satisfactory. However, they do not check the quality of the emitted signal as it sounds to the distant operator.

the monitoring oscillator

In the days when amateur transmitters employed self-excited oscillators to generate the carrier frequency, a monitoring oscillator was an invaluable tuning and operating aid. The monitor is a simple oscillating detector enclosed in a shielded container. The shielding guarantees that the signals picked up on this receiver from the output of your transmitter will be weak, and will not overload the monitor. The typical instrument is made with a minimum of components and is fitted with plug-in coils to make band changing easy. My original pre-war monitor was made in a metal lunch box and it operated from 160 through 20 meters by means of four properly wound plug-in coils. The other day I dug it out of the junk box and modified it by replacing the 1G4 triode oscillating detector tube with a 3N139 field effect transistor. I've been using it ever since, and it does such a good job that I thought the monitor and its applications might make interesting reading for other amateurs, newcomer and old-timer alike.

The circuit for the original monitor as it was constructed here at W2JIO is shown in **fig. 1.** S1 was part of the "filament lighting" type phone jack. You will note that the earphones are in the negative (ground) lead of the detector. This is done to keep hand capacity effects to a minimum, improving the frequency stability. Pickup from the phone cord is also minimized, and the unit really performs as a weak-signal receiver.

The unit is used to set the amateur transmitter within an amateur band in the following manner: Set the station receiver inside the amateur band. This may be checked by means of the receiver's dial calibration and by listening to other stations in the particular band of frequencies. Tune the monitor to zero beat with the receiver frequency by turning on the receiver bfo and then setting the monitor to zero beat with the receiver. Disable the receiver by placing it in the stand-by position and listen to the monitoring oscillator. Then tune your transmitter to give zero beat with the signal in the oscillating monitor. The transmitter frequency is now set close to the original frequency tuned in on your station receiver.

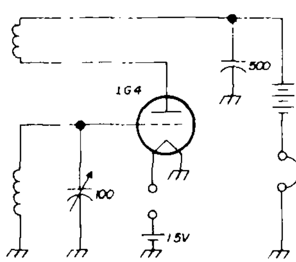


fig. 1. The original monitoring oscillator used a 1G4 triode and plug-in coils to cover 160 through 20 meters. The unit was turned on by a filament lighting type headphone jack (not shown here for clarity).

In the case of a radio-telephone transmitter operating on a-m, the quality of the carrier may be checked by listening to a beat note produced between the monitor and the carrier frequency, and then modulating the transmitter. If the pitch of the signal varies during modulation, there is frequency modulation of the a-m carrier rather than just amplitude modulation. In the case of your CW transmitter, tune the monitor to give a beat note with the transmitted signal and monitor the keying characteristics just as you might monitor them in a modern receiver. The overload characteristics of the monitor are relatively good because the entire monitor circuitry is shielded, and the character of the signal may be judged just as it appears to the distant receiving station.

When most amateurs used regenerative receivers, the monitor was a must in the operation of a station. Today, however, the stability of the average communications receiver is good enough so that the transmitter may be set directly to the operating frequency. In the case of transceivers, there is no problem because the transmitter is automatically set to the receiver's frequency.

board to accomodate the leads for the fet, and wired it in place of the tube, just as in the original circuit. The transistor oscillated right off the bat, and I could hear the beat note from my transmitter — the job was finished — or was it?

When I pulled the coil out of the socket to try the monitor on 20 meters, I found that the transistor had gone bad. I had read that you have to be careful with these insulated gate fets, and I hadn't been careful. That gate must remain at ground potential or random charges will simply burn out the microscopically thin insulation. I simply connected a one-megohm resistor from gate to ground, and found that I could change the coil without fear of destroying the fet.

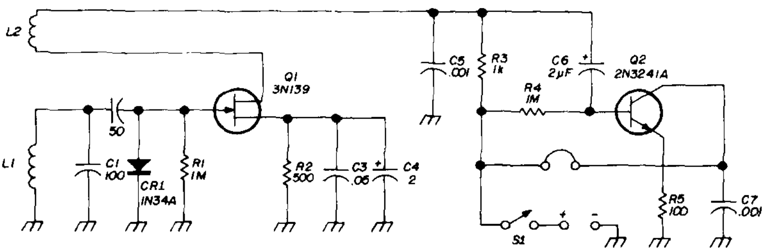


fig. 2. The modernized monitoring oscillator. Coil-winding data on the L1-L2 plug-in coils is given in table 1.

modernization

The final, updated schematic of the monitoring oscillator is shown in **fig. 2.** The steps to get to this point, however, were many. When I dug the old monitor out of the junk pile, I decided that the gadget would have to be transistorized. The tube filament was 1.4 volts at 50 mA, and the B supply was a 45 volt battery. Both batteries were expensive, and if I was to use the monitor as much as I expected, batteries would become costly. My first thought was simply to replace the 1G4 triode with a field effect transistor, a type 3N139 which I happened to have on hand.

I removed the tube socket and set a small piece of perforated board in its place — held with the same machine screws and nuts which held the original socket. I set four eyelet type terminals in a square through the perforations in the board to accomodate the leads for the fet

I tried the circuit with my surplus HS33 headphones, which have an impedance of 500 ohms, and decided that perhaps high-Z phones would give me slightly more audio to make monitoring easier. However, upon plugging the 2,000-ohm phones into the unit, I found that the transistor didn't oscillate. In fact, the total current from the battery had fallen to practically zero. Somewhere back in the distant past, I had read that a source resistance would tend to make the transistors less critical as to the load, and that transistors of the same type could be plugged into the socket and the circuit performance would be more uniform. A 500-ohm resistance shunted by the usual 0.05-mF disc ceramic capacitor from source to ground did the trick. I could now use the circuit with both headsets, and the output was even higher with the source resistor in the circuit.

audio stage

With a supply of 9 volts, I found that the output in the headphones wasn't quite what it was with the tube operating from the 45-volt B battery. Therefore, I decided that a single grounded-emitter stage as an audio amplifier following the oscillating monitor would be a worthwhile addition. This makes a very handy monitor — one which can now be used to check for harmonics as was the older circuit back in the 30's. The original

table 1. Coil-winding data for the updated monitoring oscillator. All coils are wound on standard 1¼-inch, four prong forms. Each coil has an approximate frequency ratio of three to one. Approximate frequency coverage is then: 160 meters from 1.75 to 5.10 MHz, 80 meters from 3.4 to 10.5 MHz, 40 meters from 7.0 to 21.5 MHz and 20 meters from 12 to 30 MHz.

160 meters	L1, 50 t. no. 28 enameled wire closewound; L2, 6 t. no. 28 enameled wire.
80 meters	L1, 34 t. no. 24 enameled wire, closewound; L2, 4 t. no. 24 enameled wire.
40 meters	L1, 16 t. no. 24 enameled, space wound to diameter of wire. L2, 3 t. no. 24 enameled wire.
20 meters	L1, 6 t. no. 24 enameled space wound at about twice the wire diameter. L2, 2 t. no. 24 enameled wire.

In all cases, L2 is closewound at the cold end of L1, in the same direction as L1.

phone jack with the additional filament switching leaves is still used, although it was necessary to insulate it from the chassis because the jack is now in the hot side of the circuit. As these jacks are scarce, you might prefer using a plain jack and simple SPST switch. This switch and the old bakelite vernier tuning dial are the only panel controls.

I found that the circuit starts more readily at the higher frequencies if the tank circuit is isolated from the gate of the fet by means of a small coupling capacitor of about 50 to 100 pF. The current drain of the oscillator is about 5 mA with the circuit oscillating, and this current falls to about 2.5 mA when the circuit falls out of oscillation (as produced by placing your hand on the tuning capacitor stator). This effect is just the opposite of that occurring in a vacuum tube oscillator and I decided to look into it before completing the design of the instrument.

As the gate is completely insulated from the rest of the field effect transistor, there can be absolutely no rectified dc flowing from the gate to the source through the gate leak resistor, R1. Therefore, there will be no dc voltage developed across this resistor to provide additional operating bias during oscillation of the circuitry. I decided to furnish the necessary bias by connecting a 1N34A small-signal germanium diode with its anode to the gate and its cathode grounded.

With a supply of 9 volts applied to the oscillator drain, the negative voltage is now about minus 3 volts from gate to ground. The drain current now decreases when the circuit oscillates, just as it does with the vacuum tube oscillator, and the audio output from the monitor has increased markedly with the introduction of the diode. It is quite possible that a higher value of source resistance from the fet source to ground would enhance the weak-signal performance of the instrument. However, it is quite satisfactory with the values shown in the circuit.

The monitor performs very well all the way from 160 to 10 meters, by inserting the proper plug-in-coil. It will oscillate at frequencies higher than 30 megacycles, although the oscillator stability is rather poor at these higher frequencies. This instability is produced by the poor mechanical arrangement of components and the relatively large amount of slip in the old bakelite vernier dial. However, it is reasonably satisfactory — good enough for monitoring my CW.

ham radio

helically wound mobile antenna

Improved performance over a manufactured whip is claimed for this antenna design

H. L. Booth, ZE6JP, Salisbury, Rhodesia

This article describes a helically wound whip antenna for mobile operation. The final design evolved over a period of about five years. An antenna was desired that performed better than those available on the market; tests have indicated that this objective has been achieved.*

The antenna has flat response at resonance and frequencies above resonance, with pronounced fall-off at frequencies below the design frequency.

Design data and construction details are given to enable you to duplicate the antenna, either as a single-band design (1-150 MHz) or as a 4-band amateur antenna covering 10-80 meters.

Construction procedures, dimensions, and winding instructions must be followed explicitly, otherwise the antenna may not perform as claimed. After you've built the antenna from the instructions provided here, *then* try your own variations. But it's important to "stick to the script" to start with.

*A copy of the test report is available from *ham radio for $1.00 and a self-addressed* stamped envelope. **editor.**

During the 1964/65 period, I conceived an idea to build a single-band helically wound whip into a two-band antenna while also trying to improve the coupling to space by the production of a near-sinusoidal current and voltage distribution over a short antenna. The results have been very satisfactory.

Having made many single-band helically wound whips, I noticed that a second resonance was apparent around 18-19 MHz on most antennas, using rod about 3/8 inch in diameter for the dielectric. While developing the technique to wind single-band whips for frequencies from 3.5 to over 100 MHz, trends were noted, and an antenna for the 40- and 20-meter bands was attempted. My first attempt, which was pure luck, was a helically wound antenna similar to the present multiband antenna. After rewinding and making adjustments, the first design was born. It worked on 40 and 20 meters, so I tried it on 10. It loaded and worked, but as this was done during the quiet period of 10 meters, only local results were obtained. Finally the antenna was tried on 15. It worked on that band also. Subsequent results have been most satisfactory. Tests on 20 showed 3 dB gain over a Hustler at a distance of 14,500 miles.

The form factor was a compromise, producing a near-sinusoidal distribution of voltage and current similar to that of a ¼-wave atnenna.

single-band design

Experimentation has resulted in a formula for determining the approximate length of wire for a helically wound antenna for one frequency. The formula is at best an approximation as shape, dielectric rod length, and wire gage affect the formula. To find the approximate length of wire for a helically wound whip for one frequency, use

$$L = \frac{840}{F}$$

where

L = wire length (ft)

F = frequency (MHz)

This formula will result in a little more wire than required providing the top third of the antenna length is close-wound. If less than one-third is close-wound, more wire will be required; conversely, if more than a third is close-wound less wire will be required.

The dielectric rod must be of constant diameter. Tapered rods will result in a different configuration than that specified, which may affect performance.

The rod length represents a quarter wavelength or 90 electrical degrees. Divide the rod length into nine sections, each of which represents 10 electrical degrees. To find the percentage of turns required at each 10-degree segment, use the data in **fig. 1**.

wire gage and length

Consider now one-third of the rod length. From **fig. 1** note that 71 percent of the total turns, or wire length, (using a constant-diameter rod) must occupy that space. From geometry the rod circumference is πD. Therefore, dividing 71 percent of the wire length by the rod circumference will give the approximate number of turns to be closewound. From the wire table in the Handbook, find a suitable gage of enamelled wire. The wire diameter should not be less than 0.028 inch.* If the wire table gives a size smaller than 0.028 inch, use a larger-diameter rod and recalculate.

Helically wound mobile whip for 10-80 meters used by ZE6JP.

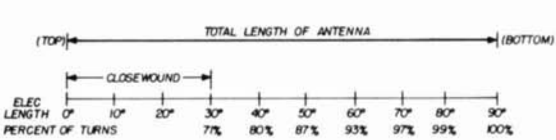


fig. 1. Diagram showing percentage of wire turns as a function of antenna length in electrical degrees. Approximately 70 percent of the total number of turns must be close-wound over the top one-third of the antenna.

Using the formula above, a 3/16-inch-diameter rod, 18 inches long, was used to build an antenna for 10 meters. The antenna was mounted on the car and tuned. An input of 22 watts was used. Good reports were received across town, but after one minute of operation the antenna was too hot to touch because the wire gage was too small.

winding procedure

Mark off the rod into 9 sections. It will be easy to determine the number of turns in each section as the rod circumference is known; also the total length of wire. Divide the circumference into the length to obtain the total number of turns. Divide each section into inches. Note that a change of turns per inch

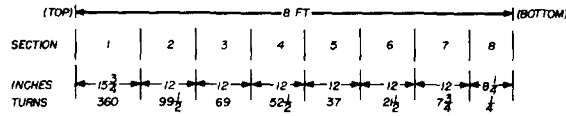
*The ARRL Handbook shows this wire diameter as No. 21 B&S gage. The current-carrying capacity of No. 21 B&S gage, at 1500 circular mils/ampere, is 0.54 amp. **editor.**

exists, section-to-section. Mark each number of turns in each progressive inch to accommodate the change of turns per inch. The winding then will have a constant change in pitch, and no sudden change of pitch will be obvious. Indelibly mark the position of each turn. Anchor one end of the wire, then secure the other end to the "loose-wound" end of the fiberglass rod and wind on. After completing the winding lock the top end with tape, then adjust the turns to smooth out any uneveness in the winding. Secure the entire winding with epoxy.

Mount the antenna in its operational position. Make certain the car has an open space of at least 20 feet around it. Use a two-turn loop to ground the bottom end of the antenna. Couple a gdo to the two-turn loop. Check the gdo frequency with an accurately calibrated receiver. The frequency should initially be lower than that required. Remove turns from the close-wound (top) end, turn-by-turn, until the gdo dips at the low end of the band. The antenna will load over the band by adjusting the transmitter tank circuit.

multiband design

This is an extension of the single-band design, but by its size and shape it will operate satisfactorily on the 40-10 meter bands. The multiband version behaves like an hf choke. As frequency is increased, resonances occur at different frequencies. These resonances are governed by the antenna shape, wire inductance, and distributed capacity. The low-impedance feed conditions at resonance are not normally harmonically related; however, this antenna does have this property. The resonances occur in the ham bands, and the feed impedance allows the antenna to be loaded by the conventional mobile pi-section tank circuit. By adding approximately 60 μH in series with the antenna base most of the 80-meter band may be covered. Similarly, the antenna will tune 160 meters with suitable inductance added at the base.



SECTION	WINDING DATA	
1	360 TURNS CLOSE WOUND	
2	/9½/9/9/9/8½/8½/8½/8/7½/7½/7½/7/	(TURNS/IN.)
3	/7/6½/6/6/6/5½/5½/5½/5/5/5/	(TURNS/IN.)
4	/5/5/5/4½/4½/4½/4½/4/4/4/4/3½/	(TURNS/IN.)
5	/7/7/6½/6/5½/5/	(TURNS/2 IN.)
6	/4½/4½/4/3½/3/2/	(TURNS/2 IN.)
7	/2/2/1½/1½/½/	(TURNS/2 IN.)
8	¼ TURN OVER LAST 8¼ IN	

fig. 2. Winding details for a 4-band amateur antenna. No. 18 AWG enamelled wire is recommended.

From band-to-band, the feed-point impedance at resonance varies but is generally between 15-50 ohms. No difficulty has been experienced when feeding with RG8/U about 15 feet long. Forget swr so long as the antenna can be loaded – improving swr adds very little to the radiation. In general, the usual pi section is adequate, unless in certain manufactured transceivers the 50-ohm termination is restricted.

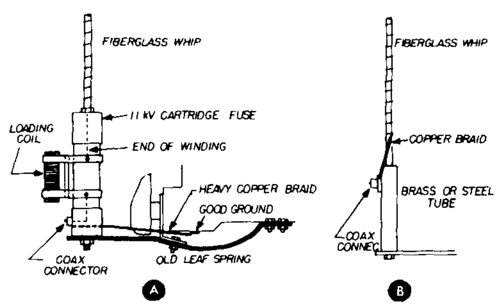


fig. 3. Suggested mounting arrangements. A shows author's mount, which includes loading inductance. A simplified version is shown in B.

The rod is of fiberglass with a diameter of ¾ inch, from 8 to 8 ft. 3 in. long. Lay the rod on a bench or table. Mark off the turn positions along the rod (**fig. 2**). Scratch the marks so they won't rub off when winding. Mark off from the top end as in **fig. 2**. Wind as previously instructed and terminate in the same way. Use only 0.040-inch enamelled wire (18 AWG). A suggested antenna mounting is given in **fig. 3**.

tuning

All previous instructions apply except as follows. Remove turns from the top of the antenna until it resonates in the low end of the 40-meter band. Check resonances on the other bands with the gdo. An increase or decrease in rod diameter will change the resonant frequency.

Note that after each adjustment of the antenna a check over the band, on each band, should be made. A compromise may be necessary in some cases, but this was not found to be so in my experiments. While testing, the antenna must be in its normal operating position. Changing from mobile to mobile may require some readjustments. Removing turns from the top has a profound effect on 40 and a lesser effect on 15; less still on 20 and 10.

80- and 160-meter operation

The antenna may be used for the two lower amateur bands by adding a suitable loading coil. The antenna should be resonated, as before, at the lowest frequency of the band you intend to use. Again, check loading across the entire band. The antenna should take power if your transmitter output circuit is not too restricted.

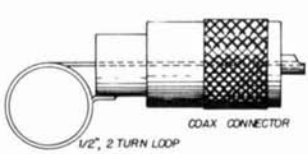


fig. 4. Loop with coax plug can be coupled to grid-dip oscillator to check resonant frequency of the antenna.

conclusion

I am informed by ZS6U that he has designed a 40, 20 and 15-meter single section, which screws onto a Hustler in place of the loading coil. Changes of wire gage are used for this antenna, but details are not available. Performance is at least equal to the single-band arrangements.

Close-up of ZE6JP's mount. Details are shown in fig. 3.

Additional resonances have been noted but no attempts have been made to use them. Typical resonances are (in MHz): 3.62, 7.05, 14.2, 21.1, 28.28, 31.8, 37.42, 44.5, 56, 67, etc. I don't know what the polarization really is, except it is mainly vertical by response on vhf. There is less decrease in signal strength when the antenna is moved from vertical than that measured from a base-loaded vertical antenna under the same conditions.

It's nice to change bands inside the mobile simply by reloading or by switching a relay to remove the short across the 80-meter coil.

ham radio

simplified

autostart and antispace

for your ST-5

This simple circuit can be easily added to your ST-5 RTTY demodulator to provide both autostart and antispace operation

Joseph M. Hood, K2YAH, 67 Mountain Ash Drive, Rochester, New York

It wasn't long after getting my teletype station operating that I became annoyed at having to constantly switch my teleprinter off, on and into *mark* when tuning or following a transmission. Before getting the teletype on the air the concept of autostart and antispace circuitry seemed to be a luxury that I could do without. However, after a few days of operating, it fast became a necessity. So began the search for suitable circuitry that would put an end to my switch-throwing frenzy.

I use an ST-5 terminal unit, fathered by W6FFC, which has been described sufficiently in previous articles. Autostart and antispace circuitry for the ST-5 has been also described previously, but every circuit which I happened upon seemed much too complex for the task at hand. So I decided to come up with something on my own – a simplified autostart, antispace (SAA) circuit.

Both autostart and antispace functions require a level-sensing device with a fairly high input impedance and good current handling capability in its output stage. After a short perusal through the integrated circuits catalogs I came across the Amperex TAA560 level detector Schmidt trigger. Its characteristics include high input impedance, low power requirements and high output current control. It looked like just the thing for a start on a simple autostart and antispace circuit.

autostart operation

The TAA560 is a four-terminal device and its operation is straightforward. A TAA560 is used to detect the presence of a signal for the front-end portion of the autostart circuitry. However, to complete the circuit it is necessary to build a time-delay network for use in the teleprinter on-off control portion of the autostart circuit and in the antispace circuitry. Again, the TAA560 is simple to

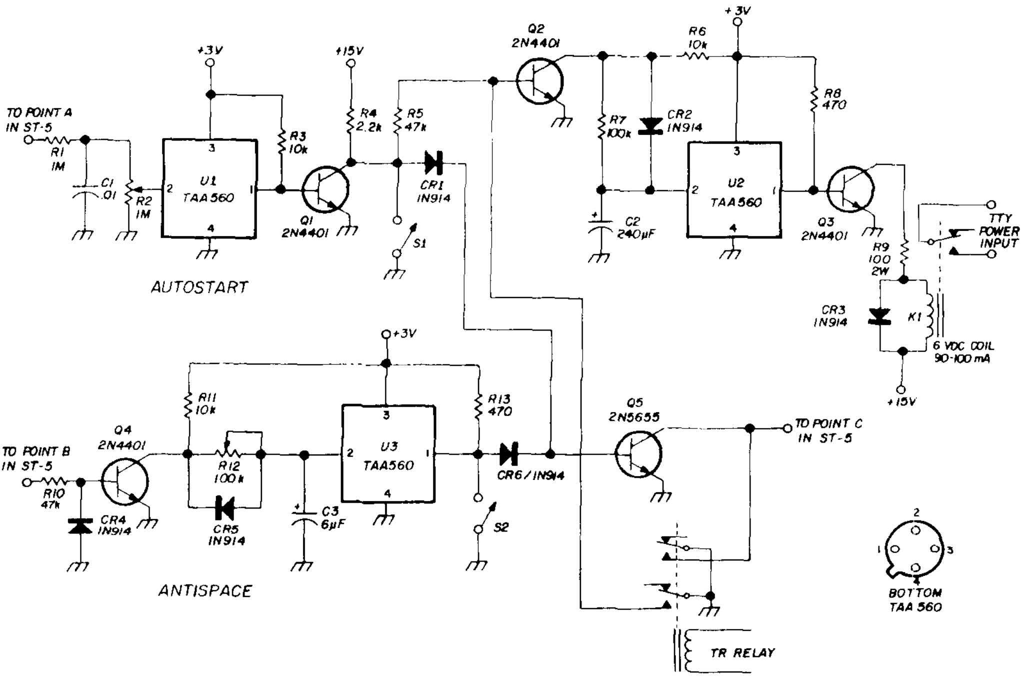


fig. 1. Complete diagram of the simple autostart, antispace circuit for the ST-5 RTTY demodulator. K1 is a 6-Vdc relay, spst.

supply voltage of 2.5 to 4.5 volts is connected to pin 3, and pin 4 is connected to return. If the input signal applied between pins 2 and 4 is above the circuit trip point (between 1 and 1.5 volts) the output transistor will be off, and the pin 1 to pin 4 output path is open circuited. However, once the input level drops below the circuit trip point, the output transistor in the TAA560 is biased on, and the pin 1 to pin 4 path will pass up to 50 mA of current.

This level-detecting feature of the use. A time delay may be had by using an RC network in association with the TAA560 input circuit. The capacitor is tied directly across the input of the TAA560 and is charged through a series resistor.

When a dc level is applied to the free end of the charging resistor the circuit waits until the capacitor is charged to the trip point of the TAA560 before changing state. The resistor-capacitor time constant determines the resultant delay. If a different turn-on versus turn-off delay is required, a diode can be used to switch a resistor into or out of the charge or discharge path, as required.

With the level detector and time delay functions in mind, let's look at the total circuit operation by referring to **fig. 1**. Start with the autostart portion of the circuit.

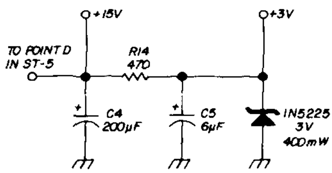


fig. 2. Power supply for the autostart, antispace circuit is connected to the ST-5 power supply (see fig. 3).

The obvious place to detect signal presence or absence is the output of the ST-5 meter amplifier at point A. The autostart input (using high impedances to minimize circuit loading) is connected here. With R2 properly adjusted, U1 will be off when a signal (+5 V or so) is present at point A and on when no signal (+2.5 V or so) is present at point A. If signal is present and U1 is off, Q1 is biased on, causing Q5 and Q2 to be biased off. When Q5 is off it allows the loop transistor in the ST-5 to control the printer magnets. When Q5 is on, the machine is locked in *mark* regardless of the ST-5 output state.

Now, back to Q2. When Q2 gets biased off by Q1 going on, capacitor C2 begins to charge through R6 and CR2. When the voltage across C2 reaches the trip point of U2 (about 3 seconds delay), U2 goes off, causing Q3 and then K1 to switch on, which turns the power on to the printer.

After loss of signal U1 goes on, causing Q1 to go off, which biases Q5 and Q2 on. With Q5 on the printer is locked in *mark.* When Q2 goes on capacitor C2 discharges through R7 and Q2. After a delay of about 30 seconds C2 is discharged below the U2 trip point, causing it to go on, which biases Q3 off, subsequently turning K1 and the teleprinter off. Diode CR3 merely protects Q3 from the voltage spike created in switching the inductive relay coil.

antispace operation

Now, how about antispace? Here you must sense the presence of a *space* condition, and after a short time delay, cause the machine to be set to *mark* even though the *space* signal persists. If you look at the output of the slicer in the ST-5 (point B in **fig. 1**) you will note that in *space* it is in negative saturation or at about − 12 volts. This causes transistor Q4 to go off which starts C3 charging through R11 and R12.

When the voltage across C2 reaches the trip point of U3, its output opens, causing Q5 to turn on, placing the machine in *mark*. The time delay set by the C3 charging circuit time constant (adjusted by R12) must be long enough to allow normal RTTY copy, but short enough to prevent annoying signals from causing the machine to run open.

Diode CR5 provides a low-impedance, fast discharge path for C3 to reset U3 immediately when a *mark* signal (+12V) reappears at point B. Diode CR4 merely protects the base-emitter junction of Q4 from breakdown due to the presence of the −12 volts at point B in the *space* signal condition. Diodes CR1 and CR6 allow the outputs of the autostart and antispace circuitry to be *ORed* into Q5 so that either can control Q5 without affecting the other circuits' output state. Again, since R10 is high, loading of the ST-5 circuitry is negligible.

construction

As far as construction goes, just about anything will do. I used a 3- by 5-inch perforated board with stake terminals and had room to spare. An etched board or any other construction technique is satisfactory. Circuit layout is not critical. Beware of mistakes in connections to Q5 since its base connections are different than you might expect.

After you have constructed your *SAA,*

some circuit setup will be required. The autostart input potentiometer should be set to produce about 1.5 to 1.6 volts at pin 2 of U1 with a *mark* or *space* signal peaked in the ST-5. Check to make sure that the circuit is operating by observing the voltage at the collector of Q1, with the antispace circuit disabled (S2 closed).

should be set so that no spikes appear at the output when receiving a normal teletype signal. However, if you don't own an oscilloscope, merely set the delay long enough to get good printout. The adjustment is not critical once you've allowed enough time in the delay for normal teletype.

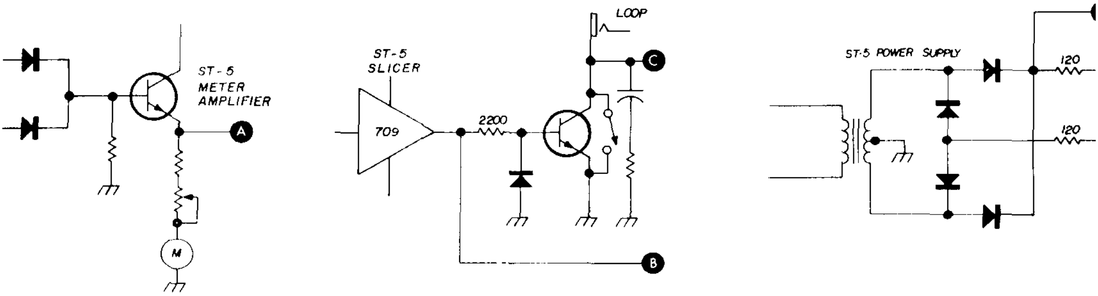


fig. 3. Connection points for the autostart, antispace circuit in the original ST-5 demodulator circuit.

As a *space* is tuned into the ST-5 space filter the machine loop should remain in *mark* until the signal goes above the trip point of U1. When this occurs, Q1 should switch off (about 1.5 to 2.0 V at its collector), K1 should then close after about 3 seconds of delay, and the machine should start and run open.

You will note that there is a slight hysteresis in the autostart circuit – the trip point to turn on is higher than the drop-out trip point. This, however, is no disadvantage since it prevents noise from initially triggering the autostart, but, once a signal trips it, the hysteresis acts to keep the printer functioning during signal fading.

setup

Setting up the antispace requires that the 100k potentiometer R12 be adjusted to provide a delay that will allow good copy, but is short enough to prevent an extended *space* condition from causing the machine to run open. This control is best adjusted by using an oscilloscope to look at the output of U3. The time delay

Switches S1 and S2 were added to allow the operator to disable the autostart and antispace functions if desired. The base of Q2 is also switched to ground during transmit to keep the autostart from turning the teleprinter off while transmitting. A most embarassing situation!

Don't try to operate the circuit from the zener-regulated, ST-5 supply. The additional loading will cause its output voltage to drop to an unacceptable level. Use the power supply connections shown in the diagram and you'll have no trouble. The ST-5 power transformer will easily handle the additional load.

That's the SAA. It has certainly made my RTTY operation less frantic and more enjoyable. Your new ST-5 and SAA may not equal an ST-6 but it comes close . . . say, an ST-5.8?

reference

1. Irv Hoff, W6FFC, "Mainline ST-5 RTTY Demodulator," *ham radio*, September, 1970, page 14.

ham radio

a single-element DX antenna

Almost unknown, the half-wave vertical can out-perform the popular ground-plane and quarter-wave verticals

B.H. Brunemeier, VQ9N, c/o 944 East Cardinal Drive, Sunnyvale, California

I have spent many years in Asia. In the Asian context, DX usually means 20- to 100-watts input on CW. A rotary beam is almost a curiosity on the CW bands as the overwhelming majority of CW DX chasers here are using a single-element wire dipole or a simple ground-plane vertical. One element, properly erected and matched, however, can produce some astonishing results.

The quarter-wave vertical, or ground plane, is too well known to require an exhaustive description. It is traditionally accepted as a very simple, and yet effective, DX antenna. However, it does have some disadvantages that are worth considering. The greatest disadvantage is its characteristic inefficiency. It is fed at a low-impedance point with a relatively high rf current. For every ampere flowing in the vertical portion producing useful radiation, there is also an ampere flowing in the ground screen. This ground-current ampere produces no useful radiation, but does account for some very significant power losses.

Most amateurs using this antenna content themselves with a ground screen of four wires, little realizing how much of their rf power is simply warming the wires and contributing nothing to the outgoing signal. The same disadvantage applies to receiving as inefficiency in the ground system saps the incoming signal to the same degree. Yet another disadvantage is that the quarter-wave antenna just isn't very tall and doesn't have nearly the receiving capture area of a full dipole which is twice as long.

half-wave advantages

From my personal observation on the air, I've noted that the full half-wave vertical is unknown around the world. I have never yet contacted another station using one. This is indeed a mystery. The half-wave vertical has several distinct advantages which make it much more attractive than the quarter-wave. Because it is a full resonant half wave, and twice as tall, it is that much better for receiving. Its base impedance is much higher than the quarter wave, and this contributes to high efficiency. A simple example will clarify this point.

Feeding 100 watts of rf into a quarter-wave vertical with a nominal base impedance of 50 ohms would produce a current of 1.4 rf amperes. A full half-wave vertical made of typical tubing would have a nominal base impedance of 900 ohms.

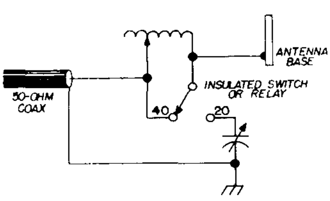


fig. 1. Switching method to use a 34-foot vertical as a half wave on 20 meters and a quarter wave on 40 meters.

Feeding 100 watts of rf into this impedance would produce a current of 0.33 rf amperes. Because the current flowing into the ground screen is the same as that which flows into the antenna, the quarter wave system would have 4.25 times more ground current than the half-wave system. The losses in the ground screen are the product of I^2R (where I is rf current and R is ground losses), and assuming the same ground screen for both antennas, the power

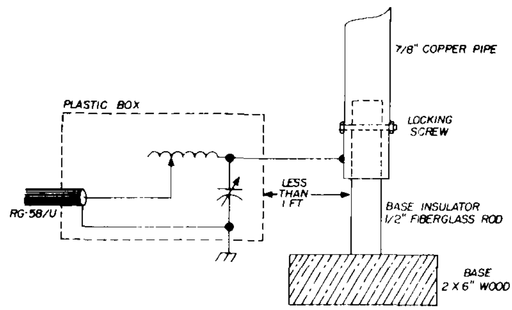


fig. 2. The antenna matching unit. The coil is 10 turns, no. 10 AWG wire, 1½-in. diameter and 1½-in. long. The capacitor is 100-pF maximum.

losses in the quarter-wave system would be eighteen times higher than in the half-wave system!

Another advantage to the half-wave system is that it has a theoretical gain of about 2 dB over the quarter wave, and concentrates that gain at a slightly lower angle above the horizon. With all these advantages to recommend the half-wave vertical, I can't help wondering why DXers around the world aren't using it. Is the 900-ohm base impedance the problem? It need not be. A simple coil and capacitor matching network takes care of that quite easily.

construction

Fig. 3 shows a half-wave vertical now in use at VQ9N. The material used is copper tubing, 7/8-inch outside diameter. It is a standard plumber's stock item on this island. Aluminum tubing is unavailable here. Note that the length is only 31 feet, rather than 34 feet, which would be a resonant half wave for 20-meter operation.

The reason for this shortage was purely economic. I bought one new 20-foot length. It was so expensive I didn't feel like buying another whole length to cut up. A scrap 11-foot length of 5/8-inch diameter was on hand, so I spliced the two to create a 31-foot vertical. The logic was that 0.45 wavelength is so close to full resonance, that it would give essentially the same performance. This logic has proven valid in practice. Also, the supporting insulators contribute some capacitive loading, which would tend to make the antenna a little taller electrically.

The most difficult part of the project was erecting the vertical. Copper is a very soft metal and cannot support its own weight in such a length, let alone the weight of guys and insulators. During my first three attempts at erecting it, my copper column suddenly became a folded dipole in the middle. This wasn't quite what I had in mind!

On the fourth attempt I enlisted a few extra helpers. Two pulled on the upper guys, one walked up under it and the fourth pushed at the top with a long wooden pushing prop. The fourth attempt was successful, though the copper column sustained some permanent standing waves along its length, created by the earlier collapses. A vertical made of 1-inch galvanized water pipe would be much easier to set up than the copper tubing I used.

The base matching coil is made of number 10 AWG copper wire (see **fig. 2**). It is wound on a form and then slipped off to make an airwound coil with a 1½-inch diameter. The original coil was made with 15 turns, close spaced. The finished coil should be spread just enough so that adjacent turns don't short together. The matching capacitor is an APC air padder, 100-pF maximum. This plate spacing is adequate for rf powers up to 200 watts, which would put about 600 peak volts across the capacitor.

tuning

The matching process is a simple matter of trial and error that can be accomplished in minutes. Insert a reflected power meter or swr bridge in the line at the transmitter end and apply enough power to give some meter deflection. Begin with the full coil in the circuit, and turn the capacitor through its range. If no dipping trend is noted on the meter, remove one coil turn and repeat the process.

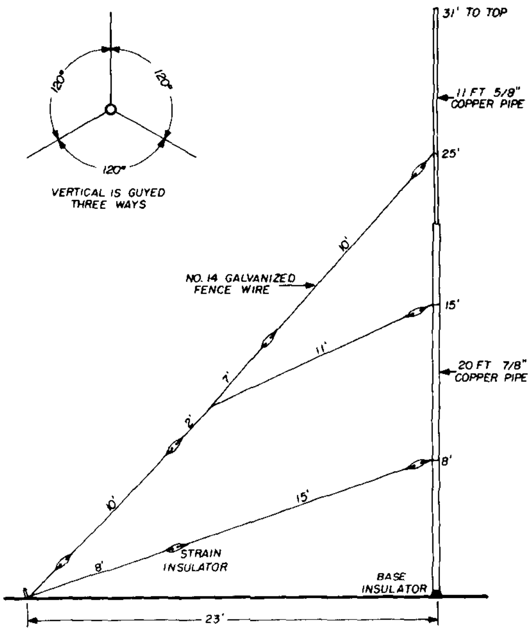


fig. 3. Overview of the 20-meter half-wave vertical. Although copper is used here, many different types of tubing could be used for the radiator.

Because the matching is quite critical, you won't see much of a meter null until you reach a point about two turns from the optimum one. Then the meter starts going down fast, and on the proper turn it can be nulled right down to zero with the capacitor. That's all there is to it.

I did my matching at 14.175 MHz, and got an swr of 1:1. The antenna response is so broad that at 14.000 and 14.350 MHz it rose to only 1.05! When the matching was finished, I had ten active turns in the circuit, which gave a coil length of 1½ inches. The unused turns were then snipped off and discarded. The capacitor was meshed to about 60-pF.

The proof of the pudding is in the signal reports. Corrugated metal roofs are almost the standard in Asia, but I went one better. My roof is corrugated aluminum, and almost level at that. A more ideal rf ground can hardly be imagined, although galvanized iron roofing does very well too. A number of tests were run on DX paths in excess of 4,000 miles to evaluate this half-wave vertical antenna in relation to other more familiar types.

I compared the half wave with the two-element quad at VQ9R and the standard quarter-wave vertical at VQ9DM (also using an almost-level aluminum roof for a ground plane). Allowing for the difficulty of taking accurate signal readings over a long path with fading, seasoned operators at the other end of the circuit gave the quad about a 6-dB advantage over the half-wave vertical.

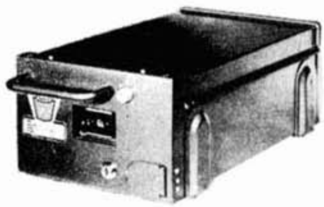
Some of you may find it hard to believe that a single vertical element could deliver a signal only one S-unit below the popular two-element quad. Comparing the half-wave vertical to the quarter wave vertical, it was found that the half wave was considerably better. In the case at hand, the aluminum roofing rf ground plane was practically lossless for both vertical antennas.

Finally, the half-wave vertical was compared to a regular half-wave horizontal wire dipole at about the same elevation. The vertical beat the horizontal dipole by a considerable margin in any direction. So then, low-budget DXers of the world, take heart! Now's the time to pull down those wire dipoles and start standing half waves on end. At VQ9N, I run only 35 watts input on CW, and I work the world with this antenna. Where a level metal roof is not available, a ground plane of wires can come close to the same performance. An increase of signal performance over a wire dipole is very effective. Can any one imagine a simpler way to achieve so much DX gain for so little investment?

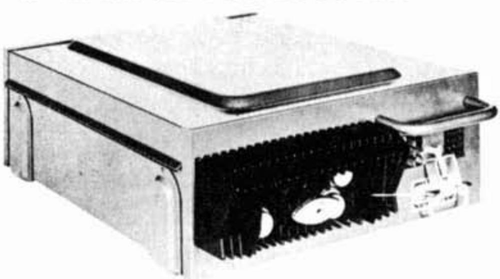
ham radio

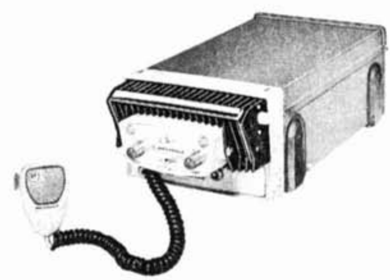
especially during their 2 for 1 Sale

plus other specials also to numerous to mention



If you've wanted to try 2 meter FM, but didn't want to invest a lot of money doing it, now is the time to get your feet wet. From now until **December 30** you can do just that. During this period you can buy 2 Motorola T43 series radios for our normal single unit price of $85.00 plus shipping. These radios are 25-30 watts output, 6/12v vibrator powered and easily convertable to 2 meters. Both units come complete with head, cable, mic, speaker and housing. You can use one unit for your mobile and convert the other to AC or find a friend who would like to get on FM with you and share the cost.





For those of you who are already on 2 meter FM our 2 for 1 sale has something in it for you also. If you've wanted to try the other popular FM bands, 6 & 10 and UHF, but were hesitant to spend big money doing it now is the time to give it a try. During this sale we will also offer 2 for 1 on all lo band and UHF completely tube type mobile radios in our 1972 catalog. You can even pick one lo band and one UHF unit and pay for the higher priced of the two plus shipping.

As usual sale limited to licensed amateurs for amateur use only. Due to fantastic discount offered, no club discounts will apply. Many units in short supply so don't wait. We will be glad to send catalog on request to help you select your radio and take advantage of such savings.

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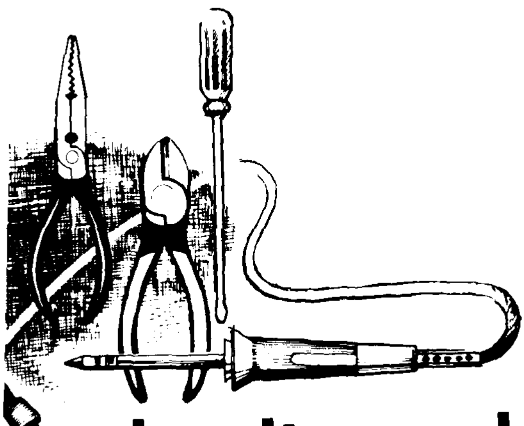


circuits and techniques

ed noll, W3FQJ

antenna tuners

Radio amateurs in the United States and, in other parts of the world too, have developed a 50 to 70 ohm transmitter and 50 to 70 ohm antenna syndrome; manufacturers of amateur equipment have contracted the same ailment. Antenna experimenters, who probably dominate ham experimentation today, have complained for years about the lack of versatility in the output system of modern amateur transmitters. Their contention is, especially at exciter power levels, that a variety of output impedances could be made available economically and with little additional space at least up to 600 or 800 ohms. In fact, the transmitter with a little more versatility and a built-in swr meter and tuner might well become a very popular model.

The usual antenna experimenter prefers to work at low-power level because it is easier to obtain more conclusive results. This is not a factor that should preclude installation of more versatility in the output systems of high-powered transmitters as well, although it is true that cost and space factors are more significant for high-powered output systems.

enter the tuner

The antenna tuner unit (atu) provides the matching capability that the transmitter lacks, **fig. 1.** Its principal duty is to see that the transmitter output is matched regardless of the impedance conditions at the transmission line input. If the transmitter is made to see a proper load, it operates in an efficient and normal manner. A second fine advantage of most tuners is that they block harmonics and other spurious signals from the antenna system. This advantage holds up even when using an antenna system that can be matched directly to the transmitter.

It also is important to know what an antenna tuner does not do. It does not alter the standing wave ratio (swr), reduce attenuation or otherwise improve operating conditions on the transmission line connected between the tuner and the antenna proper. It does not improve the operating performance of the antenna. What it does do is permit you to match the transmitter to an antenna system regardless of the impedance conditions reflected to the input side of the transmission line and the other variables, serious or not so serious, that may be inherent in the antenna arrangement. An example demonstrates the above statements.

Let us assume a peak performance, narrow-band antenna is designed to operate over 100 kHz of the 40-meter band. This antenna system has been designed to permit a direct match to the transmitter output. Except for the reduction of spurious frequencies, an atu would be of little benefit in operating over this 100-kHz span.

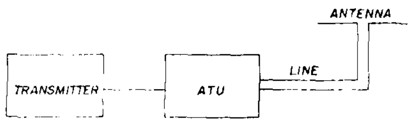


fig. 1. The antenna tuning unit on the transmitter side of the transmission line.

Off of the antenna system bandpass, impedance conditions would become unfavorable and a safe direct match would no longer be possible. Now the insertion of an atu permits the matching of the antenna system to the transmitter. Although antenna performance and line conditions are not improved by the presence of the atu, at least the antenna system can be loaded by the transmitter. In many situations the performance of the antenna system would not be noticeably different in practical communications than if the antenna were recut to this new operating resonant frequency.

Usually the impedance conditions of the above antenna resonated on 40 meters become intolerable for direct matching on 20 meters. Here again an atu of suitable design would permit you to match the transmitter to the antenna system. A least you would be able to load the antenna on 20 even though the presence of the atu does not improve the line or antenna performance on that band.

An atu is a marvel in an emergency situation and when multi-band operation is desired in a location where only a single antenna can be strung. You can at least load up the hunk of wire to obtain mediocre to good performance on a number of bands.

In summary, the atu:

Provides proper transmitter loading.

Provides harmonic and spurious signal rejection.

Permits you to accommodate an antenna that has a resonant impedance other than 50-70 ohms.

Permits you to accommodate the impedance of a broad-band, non-resonant antenna when its impedance is other than 50-70 ohms.

Permits you to load an antenna off of its resonant frequency on a given band.

Permits you to load an antenna on a band for which the antenna was not designed.

Does not change line conditions and swr.

Does not change antenna performance.

line considerations

Line factors are a consideration when using an atu at high power level. The atu does not change the line attenuation, and line attenuation does increase with the standing-wave ratio. If the line is especially long, the swr high and the attenuation per foot high, you may lose considerable power on the line even though the transmitter is matched properly.

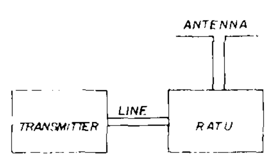


fig. 2. The remote antenna tuning unit between the antenna and the transmission line.

The power handling capability of the line is important. A high swr means voltage loops become very high on modulation crests. The rating of the line must be such that it will not break down on peaks. The higher the operating frequency the more important becomes the loss consideration because of the increase in line attenuation with frequency.

ratu

An antenna tuning unit can be made to accommodate the transmission line as a part of a matched system by locating the tuning unit between the antenna and

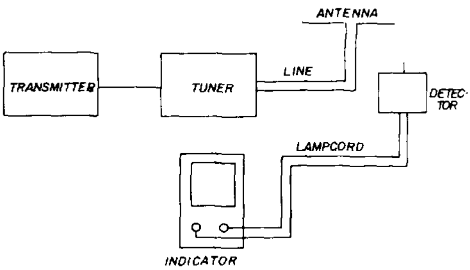


fig. 3. A remote indicator helps in adjusting the antenna tuner.

the far end of the transmission line, **fig. 2.** Minimum line attenuation and the lowest possible standing-wave ratio are now obtainable, provided the impedance of the transmission line matches the impedance of the transmitter. The remote antenna tuning unit (ratu) now matches the far end of the transmission line (same impedance as transmitter) to the antenna. This is the technique used by broadcast and other commercial transmitters that operate at a high power level.

Seldom necessary in amateur radio applications, it is employed to best advantage only when the transmission line is exceptionally long and high power is to be handled. However, it is sometimes a convenient way of matching the very low resistance and high reactance of a short 160-meter antenna. When you do wish to reduce line loss to a minimum and your transmission line is not the best in terms of minimum attenuation this arrangement is worthy of consideration.

how to tune a tuner

A critical transmitter can be damaged by reflecting an improper load from the tuner. Initial adjustments must be made at low power.

Tuners come equipped with various means of band setting — plug-in coils, switched coils or switched capacitors. Regardless of the method, set the atu to the proper operating band. In adjusting a tuner try to maintain as low an swr reading as possible with the transmitter operating at low power. Usually you will have to jocky back and forth between the tuning and matching controls of the tuner to find the very least swr. In almost all practical applications this is all that is necessary in finding a true setting and minimum swr. As the power level is increased touch-up adjustments are usually necessary. Keep records of proper settings so you can return to them after changing frequencies. In most situations it is as simple as that.

false loading

False match points are found occasionally especially when using home-built tuners or trying to accommodate wide impedance differentials between input and output. Under a false condition the component values within the tuner plus the impedance conditions presented by

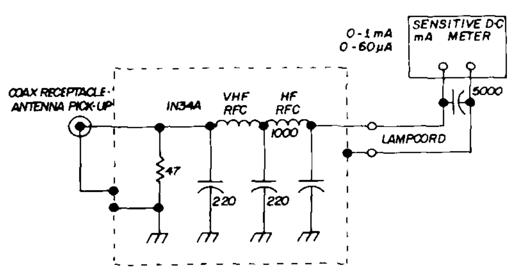


fig. 4. Simple circuit for a remote reading antenna performance indicator.

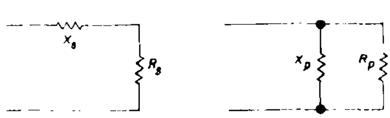
the output load are such that most of the power is absorbed by the tuner itself. What looks to be a favorable match is reflected to the transmitter. False matches can be avoided with the use of some sort of simple field-strength indicator. The pickup should be placed as near to the antenna as is possible, **fig. 3.** An occasional check of the meter reading, using binoculars or by a second person on the job, would be appropriate if you suspect a false match point.

I have found the simple arrangement of **fig. 4.** helpful. A simple diode detector and output filter are used and the antenna can be a loaded 6-meter, 10-meter or CB quarterwave vertical. Suitable readings can be obtained over the entire hf and vhf-uhf bands.

A sensitive dc meter can be used as the indicator but it need not be a part of the detector proper. A long length of ordinary lamp cord can be run between the detector output and the meter. This permits you to place the meter at a point where it can be seen as you adjust the tuner. In fact, if you keep the line well filtered and isolated as far as possible from the transmission line, you can bring the meter right into the shack or at least to a point where you can see it as you look out the shack window. Proper tuner adjustment is indicated by minimum swr and maximum field reading. A false match will result in a very weak field reading.

how does an atu function?

The atu performs two major tasks. It cancels out the reactance of the antenna system and provides the resistive step-up or step-down needed to match the resistive components. It accomplishes this by utilizing a basic characteristic of a simple or complex CL network. A series network with a specific resistance and reactance also has an equivalent parallel value of shunt resistance and shunt reactance, **fig. 5.** Conversely, a parallel combination also has an equivalent series value of resistance and reactance.



$$R_s \pm jX_s = \pm \frac{jX_pR_p}{X_pR_p}$$

fig. 5. Equivalent series and parallel combinations.

That such a relationship exists can be proven by setting down the expression for an equivalent parallel network of resistance and reactance as follows:

$$Zp = \frac{-jXpRp}{Rp-jXp}$$

This equation can be reworked to obtain the expressions for its real and reactive components (resistance and reactance) as follows:

$$Zp = \frac{Xp^2 Rp}{Rp^2 + Xp^2} - j\frac{Rp^2 Xp}{Rp^2 + Xp^2}$$

Note that the above is a simple series expression (R - jX). This is a fundamental series equivalent with the following values:

$$Rs = \frac{Xp^2 Rp}{Rp^2 + Xp^2}$$

$$Xs = \frac{Rp^2 Xp}{Rp^2 + Xp^2}$$

Further mathematical procedures can be used to set up the parallel reactance and parallel resistance equivalents of a series circuit. These are:

$$Rp = \frac{Rs^2 + Xs^2}{Rs}$$

$$Xp = \frac{Xs^2 + Rs^2}{Xs}$$

In matching an antenna system to a transmitter or line, an appropriate network (atu) is inserted between the series resistance and reactance presented by the antenna to reflect an equivalent parallel impedance that matches the strictly resistive impedance of the line or transmitter. (Sometimes the load too has a reactive component that must be considered.) *The parallel-connected network* of **fig. 6** consisting of a series inductor and parallel capacitor can serve as a

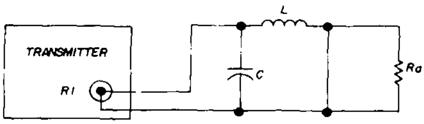


fig. 6. Simple LC matching arrangement.

simple matcher. The values of these components are determined and become of such value that an appropriate impedance match is made between the two resistive components R1/Ra. Let us assign a symbol of "n" to the latter ratio.

Further mathematical procedures can now be used to reduce the equivalent series and *equivalent parallel reactance* equations to the following simple expressions:

$$Xs = Ra\sqrt{(n\text{-}1)}$$

$$Xp = \sqrt{\frac{nRa}{(n\text{-}1)}}$$

Let us assume we are to match a 72-ohm transmitter to an antenna with an impedance of 36 ohms resistive and -160 ohms reactive (capacitive), **fig. 7.** To *balance out* the *reactive* component of the antenna it will be necessary to use a series inductor with at least an inductive reactance of +160 ohms. Additional inductive reactance will be necessary to handle the impedance match. Likewise the reactance of the shunt capacitor must *be selected* for appropriate impedance match. The two equations are now employed. Additional series inductive reactance needed is:

$$+Xs = 36\sqrt{2 - 1} = 36 \text{ ohms}$$

$$-Xp = \sqrt{\frac{2 \cdot 36}{2 - 1}} = 72 \text{ ohms}$$

The former is added to the previous 160 ohm value to obtain a required L value of:

L1 reactance = 160 + 36 - 196 ohms

The parallel reactance value becomes:

C1 reactance = 72 ohms

The above reactances can be converted to inductance and capacitance at the operating frequency by using the basic reactance equations:

$$L = \frac{Xl}{2\pi f} \qquad C = \frac{1}{2\pi f Xc}$$

If operation is centered about 1.82 MHz, actual values are as follows:

$$L = \frac{(196)\ 10^{-6}}{(6.28)\ (1.82)} = 17.1\ \mu H$$

$$C = \frac{10^{-6}}{(6.28)\ (1.82)\ (72)} = 1216\ pF$$

When antenna characteristics are not known exactly it is no great problem. You can assume very approximate values or draw from your practical knowledge of coil and capacitor sizes for a specific frequency. It is then only necessary to make one or both of the reactances adjustable. This is why antenna tuning units are indeed tunable. They permit you to adjust the matching network for an idealized match on any frequency by making adjustments and watching an swr bridge for the very best match and forward output.

counters

Thank you, Roy (R.W. Lewallen, WØETU), for sending the helpful counter data that follows:

In reference to your article in the July issue of *ham radio,* I would like to contribute a systematic method of wiring a divide-by-n counter, which besides not

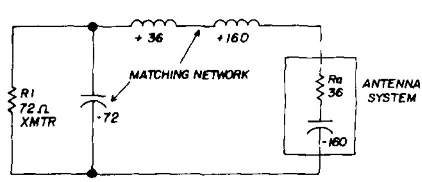


fig. 7. A typical antenna matching situation as explained in the text.

All of the above is quite understandable. However, to the uninitiated, the wiring of an integrated circuit appears to be a very complicated thing. Actually it is simple and the major complication is usually the printed-circuit board. However, this can be avoided by using straight wiring techniques as suggested in the first experimental procedures in the June column. If you use binding posts and jumpers it is also possible to change the count sequences between the combinations shown in **figs. 2** and **3.**

The pin-out wiring diagrams for the 7490 are given in **figs. 4** and **5.** Note how very simple it is. There are a number of terminals to which no connection is made and another group which are all tied to common. Of course, there are supply voltage as well as input and output connections to be made. The diagrams of **fig. 4** are for using the 2-to-1 and 5-to-1 counters separately. Both the connections of **fig. 5** provide the 10-to-1 count.

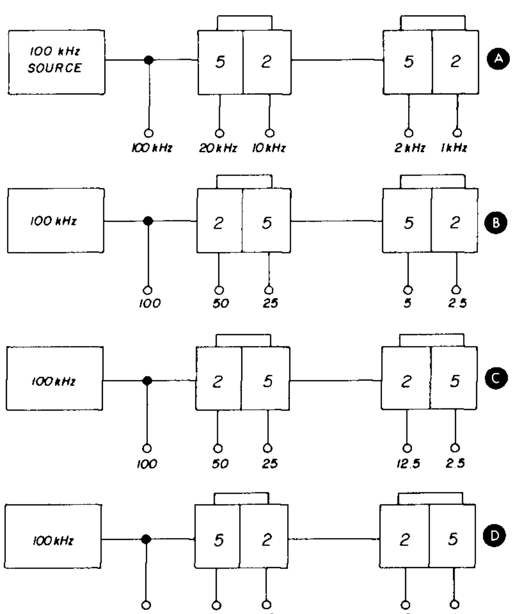


fig. 2. Some of the count possibilities using two decade dividers.

However, in the first example, the first count is 5-to-1; the second, 2-to-1. The second example is the converse, using the initial 2-to-1 count and then the 5-to-1.

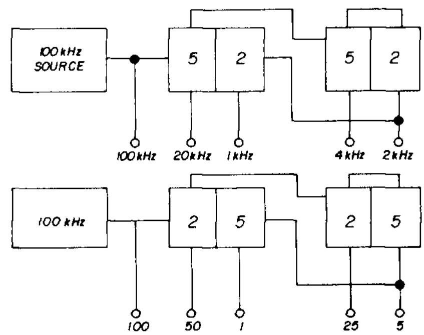


fig. 3. Two additional count possibilities using two decade dividers.

digital IC oscillators

Digital ICs of suitable design can also be used as high-frequency crystal-controlled square-wave generators. The 7400 NAND gate used initially in this series can be operated as a high-frequency oscillator. Two of the four gates are wired as a multivibrator while a third one is used as a buffer output. Doug Blakeslee, W1KLK, has used this common IC successfully with the circuit of **fig. 6A.**[1] Its output is followed by two 7490 decade dividers.

Ted Bensinger, W5PCX, uses the 7400 in the 3-MHz IC oscillator arrangement of **fig. 6B.**[2] Two of the NAND gates again serve as the multivibrator while the two other sections are pressed into service as buffer and calibrate outputs. Two decade dividers provide the countdown to 30 kHz. W1KLK operates his circuit at 3 MHz to get the same 30-kHz output. However, he employs a high-frequency 74H00 NAND gate. Theoretically this IC should provide steeper sides and higher harmonic output levels.


FUNCTION
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the ham notebook

oil-filled capacitors

When is comes to dependability, you can't beat oil-filled filter capacitors. While electrolytic capacitors have the advantages of compactness and low cost, they just don't have long lives. Whenever long-term stability and dependability are important, design engineers invariably specify oil type filter capacitors. More amateurs would be inclined to use them but for the fact that they are *reputed to be* far beyond the amateur's price range. This is not so in all cases, for surplus oil-filled capacitors appear on the market at quite reasonable prices. Some of these first-rate capacitors are rated in "working ac volts." Lacking a direct translation for this industrial rating, the average amateur will often write off these bargains.

For oil-filled capacitors, commonly used with ac, there is a unilateral conversion table for their utilization with dc voltages (**table 1**). For various reasons, there is no equivalent conversion setup for dc-to-ac, but this is of no interest for the amateur constructor.

An alternate arrangement, which has served me very well over the years, is as follows: Multiply the ac voltage listed by 2.828. What this equation shows is that the maximum steady dc voltage is equal to the *peak-to-peak ac voltage rating.* This calculation will give you the *maximum* dc voltage rating, and for the sake of conservative engineering and trouble-free operation on rectified 60-Hz ac, it is wise to de-rate the dc maximum voltages given by roughly one third.

The calculations listed above seem to work best for the higher ac voltages, and it roughly parallels the equivalent voltages in the higher ranges of **table 1**. It is worth noting that you may end up with some odd voltages, such as 2121 volts or 2750 volts. Do not allow this to confuse you, since that seemingly odd value is very close to the true rating.

In some cases, if the actual size of the oil-filled condenser in question is known, it may be possible to identify its equivalent maximum dc voltage rating by comparing its size to a dc capacitor which is catalogued and rated by the manufacturer. Armed with the foregoing knowledge, it is possible to match up the various offerings which appear from time to time. Still in doubt? Recently I picked up a 13-μF oil-filled capacitor, rated 950 Vac, equivalent dc rating approximately 2700 volts. This is the maximum rating, and when used at roughly 2000 volts, it should last a lifetime. The cost, utilizing the above information, was only five dollars.

Neil Johnson, W2OLU

table 1. Dc working voltages of ac rated oil-filled capacitors. All dc voltages listed are the nearest standard voltage.

ac working voltage	dc working voltage
110	200
220	400
330	600
440	1000
550	1500
660	2000

NE561 as an ssb detector

I presently have on the drawing board a receiver which will use a 561B as a multimode detector (fm, a-m and ssb). Needless to say, I was somewhat dismayed to read in the March 1972 "Circuits and Techniques" column that the circuit will not work as an ssb detector. I decided it was time to do some breadboarding.

In the September 1971 issue of *ham radio* WA2IKL was close. A block diagram of his detector in the sideband mode is shown in **fig. 1**. The crystal oscillator locks up the PLL and the output from the vco is fed to the balanced modulator.

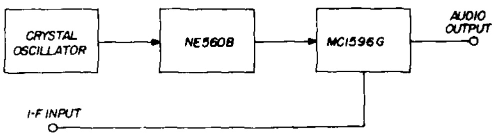


fig. 1. WA2IKL's detector circuit in ssb mode.

The NE561 is in essence an NE560 and an MC1596 combined into one package. I had assumed that the 561 would work perfectly as a multimode detector. I was given further encouragement by the Signetics applications memo on the PLL which stated, " . . . Its design is similar to the Signetics 560 Phase Locked Loop but it contains an additional product detector to perform the a-m detection function."

The block diagram of a 561 operated in the a-m mode is shown in **fig. 2**. Phase detector number one serves only to lock up the vco to the a-m carrier. This detector makes no use whatever of the

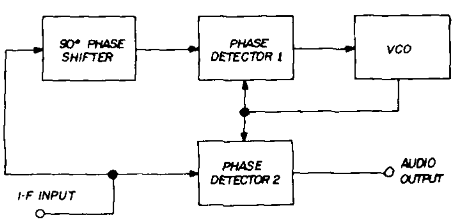


fig. 2. Phase lock loop used as a-m detector.

modulation sidebands. The vco is locked ninety degrees out of phase with the a-m carrier; therefore, when an external ninety-degree phase shift network is used the vco will be in phase with the carrier. The a-m detection occurs in phase detector number two. In the fifties, sidebanders used to call this exalted carrier detection, except for the fact that they

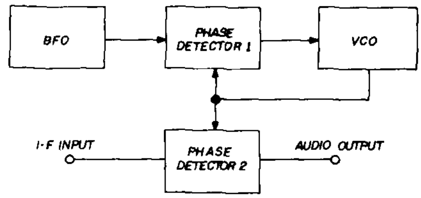


fig. 3. Phase lock loop used as ssb detector.

did not sync their bfos to the incoming signal.

Because a sideband signal is transmitted without its carrier, the missing carrier must be reinserted at the receiver. There is no point in combining the bfo with the i-f signal directly. As shown in **fig. 3** the bfo signal is fed to the fm input to lock up the loop. The i-f signal is fed to the a-m input to be detected in the second phase detector. The ninety-degree phase shift network would be meaningless in this case. The vco frequency is identical to that of the bfo, and the vco becomes the reinserted carrier. Detection occurs in phase detector number two just as in any other product detector.

The circuit which I breadboarded is shown in **fig. 4**. The 455 kHz i-f signal was stolen from the Drake 2B through the Q-multiplier socket. The audio was fed into the receiver through an audio input I had added to the 2B earlier. The third converter tube was removed from its socket to disable the unused parts of the receiver.

While there are quite a few omissions in the design, sophisticated circuitry was skipped in the name of speed. This is not intended as a construction project. The hope is that it will give some good ideas.

The performance of the detector is

excellent. Noticeable distortion both by ear and by oscilloscope sets in at about 2.5 volts peak-to-peak output. Some high-frequency bfo hiss is heard, but a good low pass filter would probably take out most of it.

My ultimate receiver will have an NE561B as a multimode detector.

Max Robinson, K4ODS

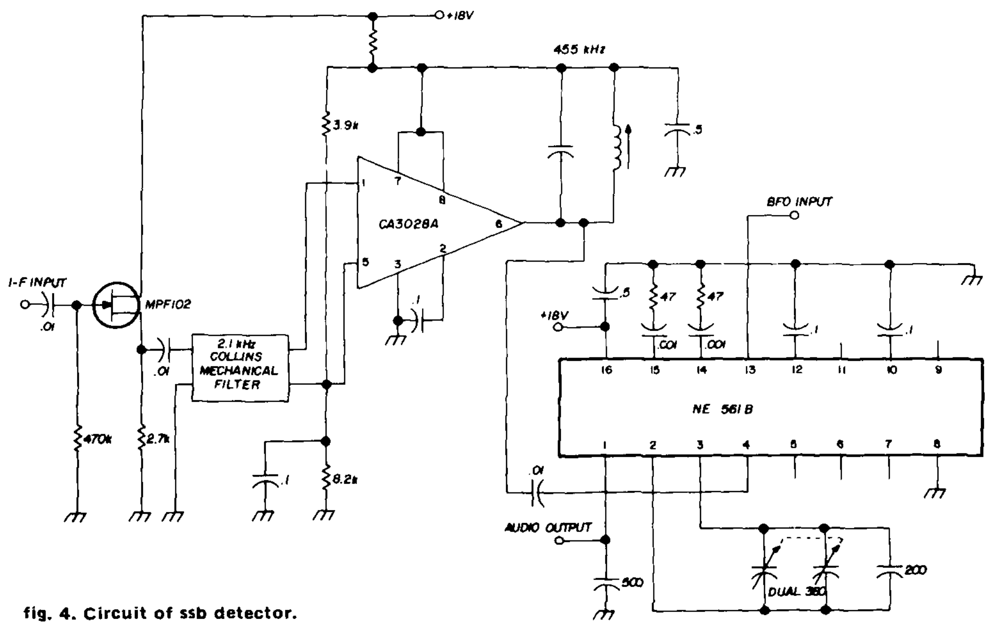


fig. 4. Circuit of ssb detector.

receiver image suppression

At a local ham swap meet I found what appeared to be a good bargain: a refurbished receiver about five years old with several interesting features, including a built-in Q multiplier. I wanted to give the set a smoke test before making the purchase, but the only source of power was being used to operate a PA system over which frustrated wives were trying to locate lost kids and husbands. Anyone who has ever been to a ham swap meet will know what I mean.

I bought the receiver anyway; it looked to be in mint condition. The set, however, had only one rf stage ahead of the mixer, which means the image response was not the best. When tuning near the low end of 20 meters, for example, the image from a strong Loran station completely wiped out the lower 5 kHz of this band. Further up the band, the image from a foreign phone station dominated another segment. The receiver was practically useless for chasing weak DX signals.

Images are easy to recognize, since they appear on the dead-zone side of zero beat in receivers with good i-f selectivity. The signal-to-image ratio in superhets can be improved in several ways, including the addition of more front-end selectivity, multiple conversion, and special circuits in the mixer input. However, I didn't want to dig into the set, so an alternate solution was needed.

The outboard trap shown in **fig. 5** is about the simplest means of attenuating images, without opening up the receiver. The LC circuit was built into a metal box to reduce hand capacitance, which makes

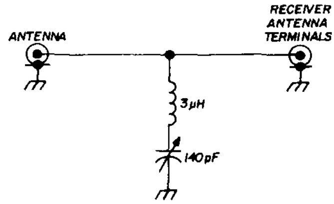


fig. 5. Series-resonant trap for improving receiver signal-to-image ratio. Circuit covers 20 and 40 meters.

tuning difficult. The trap was also effective in attenuating a strong local ssb signal that caused severe receiver overload. This device has been published in the literature many times, but I offer it again for those who may have overlooked it.

While certainly not a cure-all for receiver front-end problems, this simple circuit allowed weak cw signals to be copied that could not otherwise be heard on the 20-meter band.

Alf Wilson, W6NIF

neutralizing tip

With the tight packaging used in modern final amplifier design, it's sometimes difficult to find space for a neutralizing capacitor. The neutralizing scheme used in my mobile rig is shown in **fig. 6.** A one-inch strip of copper foil (shim stock) was formed around the final amplifier tube envelope and positioned so that it was level with the plate. The foil was

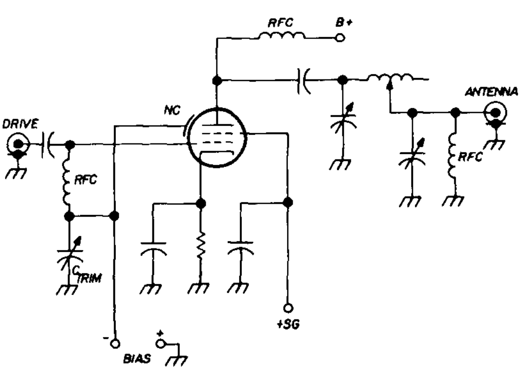


fig. 6. ZE6JP's neutralizing scheme for limited space. Neutralizing capacitor, NC, is a strip of copper foil placed around the tube envelope.

connected to a compression-type trimmer capacitor, and the amplifier was neutralized in the usual manner. A little cement on the copper prevents shifting relative to the plate.

Try the circuit first without the trimmer capacitor. It may not be needed, depending on distributed capacitance in the circuit and the interelectrode capacitance of your particular tube. Note the rf choke connected across the amplifier output. This component is sometimes omitted in pi-network amplifiers, but it's good insurance against high voltage appearing on the amtenna should the plate blocking capacitor develop a short circuit.

H. L. Booth, ZE6JP

spurious signals with the Yaesu

A number of obviously spurious ssb and CW signals have been heard from the United States and Japan recently. Investigation led to writing Yaesu for help. Yaesu's president, JA1MP, was very cooperative, and should be thanked for his assistance.

One of the signals was on lsb on 14087.37 kHz. It was caused by an usb signal on 14306. 85 (a mean frequency of 14197.11 kHz). JA1MP says that several trap coils are used in Yaesu equipment to reduce spurious radiation by at least 50 dB.

In this case the spurious signal was in an FTdx-400, and was probably caused by mistuning of the trap coils L17 and L19 which are located in the plate circuit of the transmitter first mixer. This spurious crosses at about 14,200 kHz and is strongest at that frequency. His suggestion for alignment is that the transmitter be tuned to 14,220 kHz and the receiver to about 14,180 kHz where the spurious is heard. Then adjust L17 and L19 for minimum S-meter reading on the receiver. When properly tuned, the spurious is down more than 50 dB – even at the worst point.

The CW spurious signals were heard on the 10-meter band, where they are caused in Ftdx-560 transceivers by the second harmonic of the 3180 kHz i-f that is generated by the transmitter second mixer stage when the mixer is overdriven. Especially on 28 MHz, users are apt to overdrive the rig to overcome the lower efficiency due to the higher frequency. To reduce the second harmonic, Yaesu now is modifying all rigs to install a sharp suck-out crystal filter in the i-f circuit of transceivers.

JA1MP enclosed a copy of the Spectronics "Yaesu Information Bulletin" relating to the FTdx-560/570 equipment. It shows how to place a 6358.6 kHz crystal, XT-1, across TC-3 (the middle hole in BPF-5) which tunes a tank circuit in V-203.

Bill Conklin, K6KA

current limiting

The current list price for a 2N3054 is $1.20, for a 2N3055 it is around $2.00 and for a 2N3716 it is around $6.50 (all in quantities of less than ten). For the price of a 2N1711, 75c to 85c, plus the cost of a 0.2 ohm resistor, you can save yourself many dollars in replacement costs. If you have fuses in your supply, you can save on these also. Fuses are nice, but they are just not fast enough for today's solid-state devices.

If you are using a zener-regulated supply with a series pass transistor, by merely adding a series resistor and transistor switch combination, you can have whatever range of current limiting you desire.

Fig. 7 is representative of many of today's dc supplies; however, there may or may not be a Darlington pair as I have shown here. Numerous articles have been written covering supplies for the current breed of vhf transceivers; however, some are worse than others. Some say they have current limiting, but close inspection reveals the output pass transistor is not protected as all. The November 1971 issue of *QST* had an excellent article on dc supplies. If you are building a new supply, you may prefer to follow their guidelines if you want the latest in solid-state design. The April 1972 issue of *73 Magazine* had an article on a dc supply for the HR-2. This article had "current limiting," if you want to consider a resistor in series with the bridge rectifier

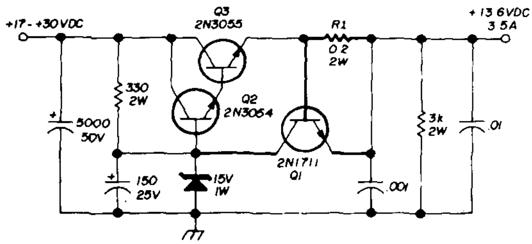


fig. 7. Adding current-limiting to an existing power supply. Q1 and R1 are new components. See text for calculating value of R1.

to collector of the pass transistor any form of limiting. This method does not prevent the pass transistor from getting extremely hot and eventually wiping itself out.

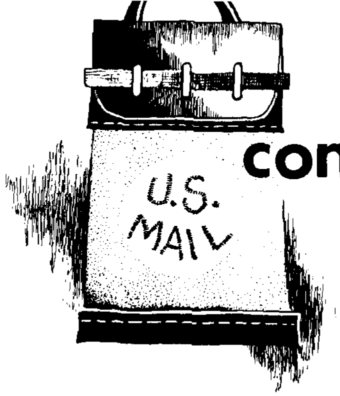
Refer back to **fig. 7**, with the added components Q1 and R1. If the current drawn by the load, exceeds a preset value (I1R1), Q1 then conducts depriving the base of Q2 of its drive voltage. The output goes to "zero" and nothing burns out. The added transistor does not need to be mounted on a heat sink, and is a TO-5 case device. The pass transistor and its Darlington driver should be mounted on heat sinks and adequately insulated with silicon grease.

If your supply does not have a Darlington driver, then connect the collector of Q1 to the pass transistor base. Just imagine that Q2 is not there.

To calculate the value for resistor R1, use the following formula: R1=(0.7)/Isc, where Isc is the short-circuit maximum current desired.

This addition to my supply has saved many power transistors from destruction. It is a very welcome addition. Thanks to Bill Durspek, WØBVR, for his help in solving my problem.

William P. Lambing, WØLPQ



comments

quartz crystals

Dear HR:

It is commendable that amateurs are becoming aware of the intricacies of quartz crystals; it is great that two authors chose to write on the subject in *ham radio.* The subject is a clear-cut science and not the black magic which some people try to make it. As a result of the black magic aura, much misinformation exists. The prime movers in the field of crystal enlightenment are the US Army Electronics Command, Fort Monmouth, New Jersey, and Bell Telephone Laboratories. Annually, interested factions of the electronics industries are brought up to date at the Frequency Control Symposium held in Atlantic City, New Jersey. IEEE-sponsored committees are continuously at work on a national and international basis to aid in standardizing terms, measurement techniques, and manufacturing procedures so that everyone can communicate on a common basis.

It would behoove the amateur radio operator to learn the terms used in the industry as he studies crystals. Author Sondgeroth interchanges terms that have specific definitions other than the meaning he intends. Such will eventually confuse any student if it has not already. While I personally have differed with the parlance of the crystal industry, I find it is necessary to use the language understood by my converser.

The term "holder" refers to the case in which the quartz is mounted. The holder is only a minor contributor to C_O. Static capacitance (C_O) is primarily a function of the electrodes sandwiching the quartz. Why must the author refer to C_O as holder capacitance? Depending on the area of the plating, and the thickness of the quartz, C_O can range from 3.5 to 7.0 pF on crystals commonly found in surplus houses. The figure of 5 pF is only a seat-of-the-pants estimate.

Elements of the motional arm are usually designated C1, L1, and R1, although C_M, L_M, and R_M are understandable. When the subscript M is used, I can't see the reasoning behind defining the terms as "equivalent" and "effective." I believe this is misleading if not incorrect. "Equivalent Series Resistance" is the term used to define the resistance of the crystal in an oscillating circuit and it includes series resistances and parallel conductances of the holder, not shown in this equivalent circuit. Hafner* uses the term equivalent a second time as equivalent reactance (X_e) of an oscillating crystal, but this cannot be broken down into the reactances of C_M and L_M,solely. The reactance of C_O also has an effect. The terms motional capacitance, motional inductance and motional resistance are preferred.

A point on which the reader should also be cautioned is Mr. Sondgeroth's statement that you can save money by tuning a lower accuracy crystal. Not necessarily so! Lower accuracy crystals are generally so designated because they have poorer temperature characteristics. If you operate under conditions of wide temperature variations, you may become a slave to the technique of crystal tuning.

It is not my intent to sharpshoot the article — I feel it is well written and pertinent to an amateur's problems.

*Standard Definitions and Methods of Measurement for Prezoelectric Vibrators, IEEE No. 177, May, 1966.

In G8ABR's contribution to the *ham notebook,* a statement is made that air-mounted crystals, e.g. 10X, will always oscillate at exact multiples of the fundamental. I believe in this case the crystal is vibrating in the fundamental mode and tripling electrically. Most of the air mounts (pressure types) will not operate in harmonic mode because the coupling of the plates to the quartz is poor. This would include FT-243 and older HC-6/U crystals. It is just about impossible to make any crystal with exact frequency multiplication by operating in harmonic modes.

Don Nelson, WB2EGZ
Voorhees, New Jersey

code practice

Dear HR:

The editorial in the May 1972 issue concerning building up code speed with *coastal station transmission* copy practice is certainly meritorious. **Table 1** is part of the schedule I compiled and distributed to the fleet of off shore tankers I am connected with, for use by the Radio Officers. Data was secured from various publications and direct observations and checks.

WSL transmits about a half hour weather copy four times a day as indicated. It is tape and about 23 or so wpm. WAX is good for two daily periods at about 16 wpm, also tape. WOE sends two daily periods, about 20 or so wpm, and most of the times is good copy. WSL on the 12-MHz frequency seems to be about the best bet for the most powerful signal.

About the only maritime telegraph stations I have found that send in the 30- to 35-wpm category are the Russians. Their merchant fleet, in recent years, has expanded phenomenally. They met the need for new frequencies without trouble. They picked out low-powered coastal stations and moved on top of them. They stayed clear of the high powered ones like WSL or WCC. One example is UQK, Riga, who is parked on WLO, Mobile, Alabama, 12704.5 kHz. They are netting in as the ships use the same frequency as the coastal stations. Weather transmissions from their ships to UQK is at about 0000, 0600, 1200 and 1800 GMT. Speed is in the 30- to 35-wpm range, very good code sending addressed to *pagoda* and is in standard international code. These messages are mostly number groups with ships names and should be good practice for our high speed boys.

I would suggest stressing the matter of *regular daily* code practice to build up speed. Once or twice a week will be of not much help in building up speed. A month or so of *daily practice should* bring a 13 worder up to 20 plus with no strain.

Paul Szabo, WB4LXJ
Tampa, Florida

table 1. Weather transmissions in International Morse Code.

station location	time GMT	frequency kilohertz
WCC	0050	436 2036 4331
Chatham,	1250	436 2036 6376 8630
Mass.	1650	436 2036 6376 8630
WSL	0500	418 8514
Amagansett,	1100	418 8514 13024.5
New York	1700	418 8514 13024.5 17021.6
	2300	418 8514 13024.5 17021.6
WSC	1418	460
Tuckerton,	2318	460
New Jersey		
WMH	0130	428 8686
Baltimore,	1600	
Maryland	1330	428 8686 12952.9
	1930	
WOE	0105	472 6411.35 8486 12970.5
Lantana,	1605	
Florida		
WAX	0135	488 4295 8526 13011
Miami,	1335	488 8526 13011 17199.2
Florida		
WPD	1800	420 13051.5
Tampa,		
Florida		
WLO	1300	438 8714 12704.5
Mobile,	1700	
Alabama	2300	
WNU	0430	478 4310 6495 8570
New Orleans,	1630	478 6495 8570 12826.5
Louisiana		
WPA	1748	416 8550 12839.5
Port Arthur,		
Texas		

power in reflected waves

Dear HR:

Various letters have been received, commenting on my paper, "Power in Reflected Waves" (*ham radio,* October, 1971). Most writers agreed, but some disagreed with my principal conclusion that there is no power in reflected waves on a transmission line. I wish to thank all who wrote me, for their interest in my paper.

Much of the disagreement appears to be based on incredulity, rather than on reasoned technical analysis. This is understandable, because most pertinent writings which have appeared in amateur periodicals, and indeed in important handbooks written for amateurs and even for professionals, have discussed power in reflected waves as if it were a reality. Rare exceptions are 'Losses in Feedlines" by Byron Goodman, *QST*, December, 1956, and "The Mismatched RF Transmission Line" by Carl C. Drumeller, *73 Magazine,* November, 1969. These writers correctly said that so-called *reflected power* is not really power at all.

The technical criticisms received can be combined and summarized as follows:

1. The voltage and current standing waves on an unmatched transmission line have a phase difference of 90 degrees only when reflection is complete, and therefore, under other circumstances they represent power.

2. The input impedance of the transmission line is not matched, and cannot be matched, to the output impedance of the transmitter unless the tube's load resistance is equal to the tube's plate resistance, which is not the case in practice; therefore, virtually complete reflection of the *power* in the reflected wave does occur at the transmitter output.

3. In the case of my **fig. 2**, there is a reflected wave on the coaxial line shown, for the reason that there is no other possible destination for the *power* in the reflected wave on the open transmission line.

I will discuss briefly these three points in order.

Point 1. The fact is, the voltage and current standing waves on a mismatched transmission line have a phase angle of 90 degrees even though reflection is not complete. To avoid going into a detailed proof, the references are cited in evidence.

Point 2. This criticism is incorrect, but in any event it is irrelevant; it does not prove or demonstrate that there is power in the reflected wave.

Point 3. This is a peculiarly circular argument. It claims that I am wrong in denying that there is a reflected wave on the coaxial line in my **fig. 2,** because, contrary to my principal conclusion, there *is* power in the reflected wave on the open transmission line, and this must appear on the coaxial line. In other words, it says that I am wrong because I am wrong! This criticism, like the one above, does not prove or demonstrate that there is power in the reflected wave on the open line.

Whoever originally wrote about power in reflected waves on transmission lines as being a reality, no doubt thought that this concept would serve to simplify, for the non-professional, the manner of formation and significance of standing waves. Historically, it has served only to complicate the matter endlessly, as writer after writer, following the original lead, and unwilling to break with precedent, has grappled with reflection of *power* in rf lines and its re-reflection back and forth ad infinitum, with trying to explain how it is that a directional wattmeter can indicate, under some circumstances, more power in the line than the transmitter is putting into it, and how power at the same frequency can travel both ways simultaneously. They seem to have forgotten that the basic definition of power is, "the rate of doing work," and have failed to show where and how this work, corresponding to the assumed power in the reflected wave, is being done. Of course, they cannot show this, because it is not power and therefore is not doing any work.

The underlying error in these misconceptions is the failure to distinguish between ac power, or volts ac times current ac, multiplied by a power factor other than zero, from volts ac times current ac, multiplied by zero power factor. The former is power; the latter is not, and it is a fundamental error to call it so. Power utility engineers know better than that.

The basic point to be recognized about power in a transmission line is simply that the power from the transmitter into the line is the sum of the power lost in and from the line and in any additional matching or other devices inserted into the line and the power delivered to the antenna. There is no other power moving in either direction. It is that simple.

Hubert Woods
Jalisco, Mexico

tuning toroidal inductors

Dear HR:

In his article, "Tuning Toroidal Inductors," (*ham radio,* April, 1972), author WAØJYK indicates that a grid-dip oscillator cannot be used because there is not enough flux leakage from the toroid.

The fact is that a grid-dip oscillator will give excellent readings on a tuned circuit having a toroidal inductance. Just put a loop of wire through the toroid and twist it into a link around the coil of the grid-dipper. If a precision capacitor is used, the inductance can be calculated to a degree of accuracy limited only by the care with which the resonant frequency is read.

Even rough checks by the gdo-capacitance-frequency method can give better results than those given by all but the best laboratory bridges since the measurement is usually made with the inductance excited at the frequency at which it will actually be used.

Barry Kirkwood, ZL1BN
Auckland, New Zealand

reciprocating detector

Dear HR:

I have received several letters regarding my "reciprocating detector" article which appeared in the March, 1972, issue of *ham radio*. Transistor Q5 is part of the reciprocating detector switch, but the questions are understandable due to the lack of a dot to show a connection in the schematic; resistors R4 and R5 should be joined with a dot where these two resistors form a junction point at the input to the diode and the base of Q5. The diode is a 1N252.

Several readers have also asked where the selectivity curve is 500-Hz wide and what is its slope. The filter I used was designed to have its 500-Hz passband at the 3-dB points on a slope which is not particularly steep for an inductive filter. Indeed, at 500 Hz, the L3 inductance is very loosely coupled to the other two sections of the transformer. The bandpass formula (f_r/Q_o) indicates that the bandpass of the filter is actually narrower than 500 Hz – in fact, bandpass is closer to 250 Hz. The 390-ohm resistor used in series with one of the differential inputs loads the thing down so it is broader. If the bandpass is too narrow, poor lock-in range is experienced on a-m, and there is very poor "presence" in the quality of ssb signals. If the bandpass is too wide, poor impulse rejection will result.

Stirling M. Olberg, W1SNN
Waltham, Massachusetts

RTTY speed converter

Dear HR:

I just completed construction of the RTTY speed converter described by WA6JYJ in the December, 1971, issue of *ham radio,* and it works like a charm. I built the converter on a printed-circuit board which greatly simplified its construction. I can furnish printed-circuit boards to interested readers for $6.00.

Earl E. Palmer, W7POG
17510 Military Road South
Seattle, Washington 98188

radio control

Dear HR:

I would like to remind your readers that the frequencies, 53.10, 53.20, 53.30, 53.40 and 53.50 MHz have been recognized by the FCC as *radio control frequencies* for licensed amateurs who engage in remote control of model boats or airplanes.

Interference on these radio control frequencies has been on the increase, causing loss of control. This can be disastrous to the model builder who has spent countless hours and a lot of money on his model, only to see it crash because of interference.

Considering all the frequencies available to amateurs who use six meters for communications (CW, ssb, RTTY, etc.), it seems reasonable to ask them to stay clear of the radio control frequencies noted above. In addition, since it is impossible to build fancy receivers into the very small space available in most models, it would be appreciated if a reasonable guard band, say 6 kHz, could be observed.

Pierre J. Catala
Needham, Massachusetts

pi-network inductors

Dear HR:

W6FFC's article on pi-network design is easily the best and most comprehensive treatment of the subject that has ever been published in a ham magazine. Congratulations. I believe I can add something on the matter of inductors for high-power tank circuits.

There has been a lot of theorizing on the effects of corrosion on bare copper, and the benefits of silver plating, but little actual measurement. Some time ago, out of curiousity, I resurrected an old 10-meter tank coil from my junkbox. It had been wound at least ten years previously, and consisted of several turns of 1/4-inch copper tubing, 2-inches long. It was the familiar chocolate brown color of old copper.

I measured the Q of this coil on a freshly-calibrated HP 260-A Q Meter. It measured 173. Next, I had the coil chemically cleaned and brightened. The Q increased to 176. Then the coil was silver plated .0002 inch. This raised Q to 178. All of these measurements were made during an eight-hour period.

As a final experiment, I wound a coil of number-14 tinned copper bus wire, of the same length, diameter and inductance as the copper tubing coil. The Q measured 172.

These measurements show that the benefits of silver plating are negligible at frequencies up to 30 MHz. The difference in efficiency of an amplifier using any of these coils could hardly be measured. However, I would not recommend the use of the wire coil, since, as W6FFC points out, at 30 MHz the coil dissipation may be as much as 100 watts, and the wire coil would be inadequate. Dissipation rather than Q is the real reason for using tubing or heavy strap at the high-frequency end of the range.

The experiments also demonstrated something that the textbook equations for the rf resistance of an inductor imply: the Q of a coil, over wide limits, is more dependent on the size and shape of the coil than on conductor size.

Harry R. Hyder, W7IV
Scottsdale, Arizona

laser communications

Dear HR:

In regards to W4KAE's letter in the May issue, his claim to the first two-way laser QSO is a little late. There was a two-way laser QSO on February 25th, 1971 between WA8WEJ/Ø and W4UDS/Ø on a frequency of 475 THz using A3 modulation (*QST*, July, 1971, page 93). This precedes W4KAE's contact by almost 9 months.

Although W4KAE's two-way laser contact was not the first, it is definitely the DX record to date. Keep up the good work, Ralph.

Lee Yazell, WB9AIU
Pensacola, Florida

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Eatontown (07724)
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Radio, Inc.
1000 S. Main Street

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Albany (97321)
Oregon Ham Sales
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Portland (97205)
Portland Radio Supply Co.
1234 S. W. Stark

PENNSYLVANIA
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Pittsburgh (15222)
Cameradio
2801 Liberty Avenue
Tydings Company
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new products

alpha 77 linear amplifier

The Alpha 77 Linear power amplifier has several new design features which have been added since the unit was originally released. The new Alpha 77 now uses an air-cooled Eimac 8877/3CX1500A7 ceramic-metal triode with 4000 volts on the plate. This power tube has a conservative 1500 watts of plate dissipation and requires only 65 watts drive for full legal amateur power input.

Also new in the Alpha 77 is a grid excess-current relay which will kick out if the final tube is under-loaded or over-driven. This protects the tube and the input circuit. Primary power requirements for the amplifier are now 120/240 volts at 50/60 Hz, single phase, making the unit compatible with overseas power sources.

The Alpha 77 is designed and rated at 3000 watts PEP continuous commercial service and is available to amateurs who demand the ultimate in every respect; it loafs along at 2000 watts PEP. The Alpha 77 features a rugged bandswitch with 20-amp silver contacts, vacuum-variable tuning capacitor, silent vacuum relays, quiet forced-air cooling and metering in all circuits. The massive plate coil is silver soldered and heavily silver plated for efficiency. Husky toroid coils minimize coupling between pi-L network sections.

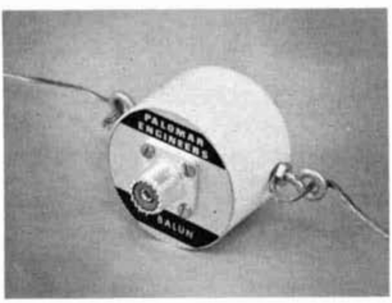
The Alpha 77 Power amplifier is built like a battleship with ¼-inch thick aluminum sides, and weighs in at 70 pounds. Modular assembly is used throughout so that the power supply, rf deck and control panel are easily removable. Second harmonic output is typically -50 dB, and the third-order intermodulation products are -35 dB below peak output. The Alpha 77 is manufactured by Ehrhorn Technological Operations, Inc., and distributed by Payne Radio, Box 525, Springfield, Tennessee 37172. For more information, use *check-off* on page 126.

two-meter portable antenna

A new industrial type, continuously-loaded vhf portable antenna has been added to Antenna Specialists' high-performance amateur line. The new Antenna Specialists HM-5, designed to withstand the rough handling that makes telescopics impractical, is completely insulated and cannot accidentally be shorted out. It features a connector fitting that attaches directly to portable equipment with SO-

239 connectors. Power rating is 25 watts with nominal input impedance of 50 ohms. A companion model, the HM-4, is identical except for a standard 5/16-32 threaded-male mounting base. Both models are available from amateur distributors at a suggested ham net price of $5.95. For additional specifications, write to Amateur Department, The Antenna Specialists Company, 12435 Euclid Avenue, Cleveland, Ohio 44106 or use *check-off* on page 126.

palomar balun



Palomar Engineers has announced a new 1:1 Balun. It matches 50- or 75-ohm coaxial cable to center-fed dipole or inverted-vee antennas. By preventing radiation from the coax, a balun improves the antenna radiation pattern, reduces noise on receive, and helps prevent tvi, bci and rf feedback within the station.

The Palomar Balun transformer is wound on a large ferrite toroid core and handles a full kilowatt from 1.7 to 30 MHz. The transformer is enclosed in a white plastic housing and is completely encapsulated to prevent moisture from entering. All hardware is stainless steel.

Eye bolts on the sides allow the balun to replace the center insulator of the antenna, and an eye bolt on top can be used to support the antenna. The balun is a compact 2¼ inch in diameter and 2 inches high. The unit is priced at $12.95 postpaid in the United States and Canada (plus 5% tax in California). A descriptive brochure is available. For more information, write to Palomar Engineers, Box 455, Escondido, California 92025 or use *check-off* on page 126.

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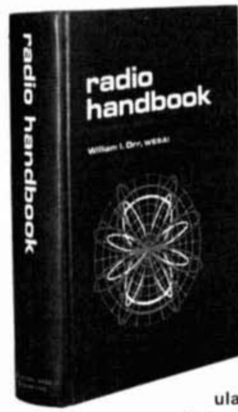

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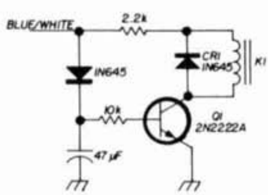
short circuits

mobile touch-tone

In the article on mobile Touch-Tone operation in the August, 1972, issue, page 60, the capacitor and resistor across relay K1 can cause immediate failure of the transistor due to excessive current flow. When the transistor is first turned on, and the capacitor is not charged, current flow is about 500 mA since the 22-ohm resistor is the only thing that limits current. WB8NAT has suggested a modification which places the time-constant capacitor in the base circuit of the transistor. This has proved quite effective and according to several fm operators who have built it, the circuit works very well. The diode has been added to the transistor base circuit to prevent the dial from shorting out the charge on the time-constant circuit.



Improved time-constant circuit.

frequency-measuring oscillator

The frequency-measuring oscillator described in the April, 1972, issue was originally designed and built by Ben Christie, K2BF. A footnote should have been included in the article to that effect. It seems that Ben designed and built an fmo and sent it to a friend, Peter Petersen, K6MFS, in Long Beach, California. It was at K6MFS's house that author W6IEL saw the original fmo, realized it would make a good construction article and obtained full information from K2BF. The photographs in the article were taken by K6MFS, and are of the original fmo built by K2BF.

multimode i-f system

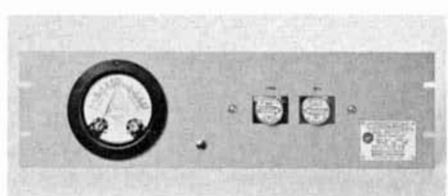
In the article on the multimode i-f system in the September, 1971, issue, several component designators were not included on the schematic. Capacitor C6 is the .001-μF capacitor from pin 14 to pin 15 of the NE560 PLL IC. C1 and C2 are the two 100-pF capacitors across L1 and L2. R1 is the 1k resistor from pin 2 to pin 3 of the MC 1596G IC. R3 is the 3.9k resistor connected to pin 9 of the MC1596G.

frequency scaler

In the circuit of the frequency scaler, **fig. 2B**, page 42 of the September, 1972 issue of *ham radio,* the .01-μF capacitor should be connected to both V_{cc} pins on the 95H90 IC. The circuit-connection dot was inadvertently left off by the draftsman.

direct-reading swr meters

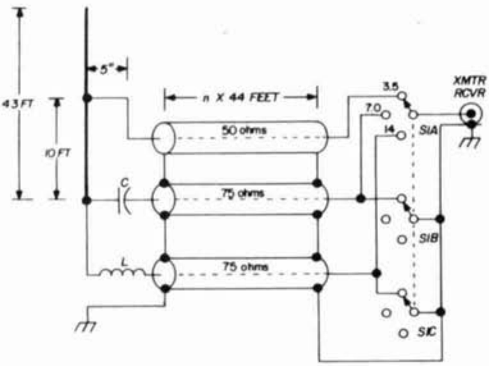
The meter scale for the direct-reading swr meter (**fig. 1**, page 29, May, 1972) is incorrect. The author states that, "An swr of 1:1 is obtained if the reflected power is equal to zero." That would place the swr line labeled "1" directly between the zero on the *reflected* scae scale and the pivot of the reflected meter pointer. In **fig. 1** that line is displaced to the left, so that a reflected reading *less* than zero would be required to obtain an swr of one. In addition, Thruline® is a registered trademark of Directional RF Wattmeters manufactured by Bird Electronic Corporation. Author WA4WDK's vswr meter was patterned after the Bird model 3122 Thruline® Wattmeter pictured here.



Bird Model 3122 Thruline wattmeter.

logic monitor

In the logic monitor circuit described in the April, 1972, issue (page 70) the output of the first section of the MC844 should be connected to the 1k pullup resistor.



LA1EI's three-band ground-plane antenna.

three-band ground plane

Fig. 6 of LA1EI's excellent article on the three-band ground plane in the May, 1972, issue is in error. The corrected drawing is shown above.

repeater control with simple timers

In **fig. 1** on page 47 of the September, 1972, issue, C1, the 400-pF capacitor, and the 33-ohm resistor should go to ground, not to Q of U1. Make the same change to **fig. 2** on page 48.

power supply ics

The RCA CA3055 voltage-regulator IC specified in the article on page 50 of the November, 1970 issue of *ham radio* has been replaced by the RCA CA3085. The CA3085 is a 12-mA device, while the CA3085A is a 100-mA device which is a plug-in replacement for the old CA3055. The CA3085B, a member of the same IC family, has the same ratings as the CA3085A, but can withstand higher input voltage surges. The price, incidentally, of the CA3085A is *less* than the CA3055, welcome news in view of the higher costs of nearly every commodity these days.

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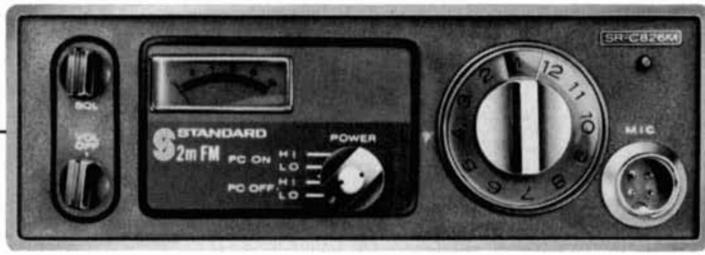
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WORLD QSL — See ad page 118.

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VK7 GOLDEN JUBILEE AWARD RULES: Radio Amateurs outside Australia and New Zealand to contact five (5) Tasmanian stations (VK7) during the period 1st January to 31st December 1973. Any amateur band may be used. Any amateur mode may be used with cross mode being acceptable. Cross band working is not acceptable. A copy of the log showing date, time, band and other relevant details signed by the operator and two (2) other licensed amateurs or by the operator and the Secretary of his Club. Isolated operators who are unable to comply with the above may request their logs be subject to check by the organizers. A suitable certificate will be issued. Sea Mail 3 IRC Air Mail 6 IRC. Address for logs: VK7 Golden Jubilee Award, Box 851J, G.P.O., Hobart, Tasmania, 7001.

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SEE INSTRUCTIONS ON PAGE 2 (REVERSE)

1. TITLE OF PUBLICATION: HAM RADIO MAGAZINE
2. DATE OF FILING: September 29, 1972
3. FREQUENCY OF ISSUE: MONTHLY
4. LOCATION OF KNOWN OFFICE OF PUBLICATION *(Street, city, county, state, ZIP code) (Not printers)*: MAIN STREET, GREENVILLE, N. H. 03048
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6. NAMES AND ADDRESSES OF PUBLISHER, EDITOR, AND MANAGING EDITOR
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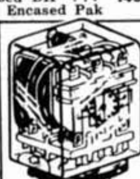

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WANTED: SWAN MARK 6B LINEAR, and TV-2C Transverter. Russell, 19680 Mountville Drive, Maple Heights, Ohio 44137. 216-662-2175.

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22nd ANNUAL DAYTON HAMVENTION *will be held* on April 28, 1973 at Wampler's Dayton Hara Arena. Technical sessions, exhibits, hidden transmitter hunt, flea market, and special program for the XYL. For info write Dayton Hamvention, Dept. H, Box 44, Dayton, Ohio 45401.

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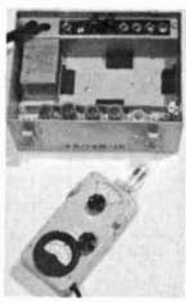

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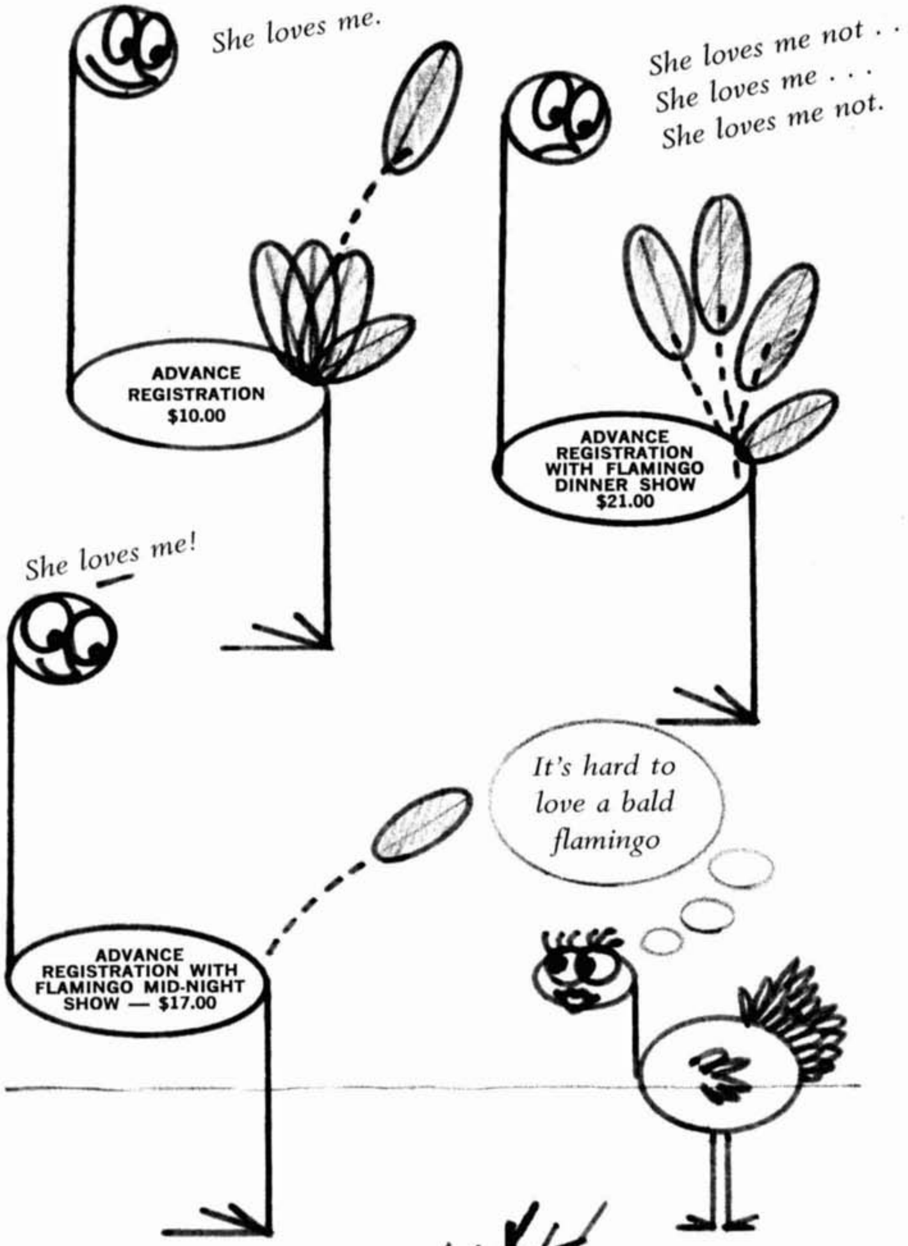
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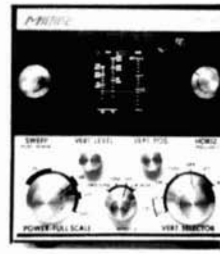

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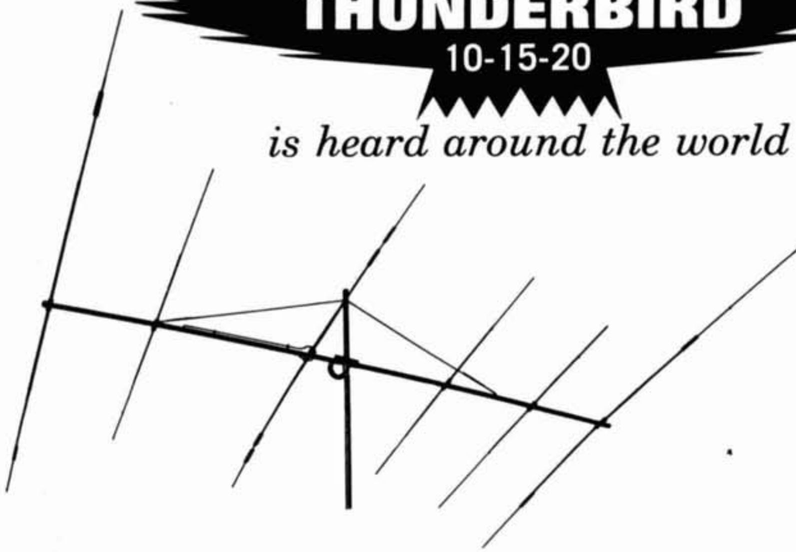

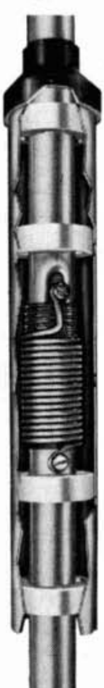

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ham radio

index to volume V – 1972

This index covers all articles published in *ham radio* during 1972. The articles are listed alphabetically under each category along with the author, page number and month. Categories are: antennas and transmission lines; commercial equipment; construction techniques; fm and repeaters; integrated circuits; keying and control; measurements and test equipment; miscellaneous technical; power supplies; receivers and converters; RTTY; semiconductors; single sideband; transmitters and power amplifiers; vhf and microwave. Articles followed by (HN) appeared in the *ham notebook*.

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fm and repeaters

integrated circuits

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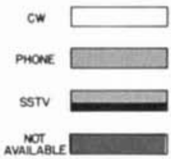

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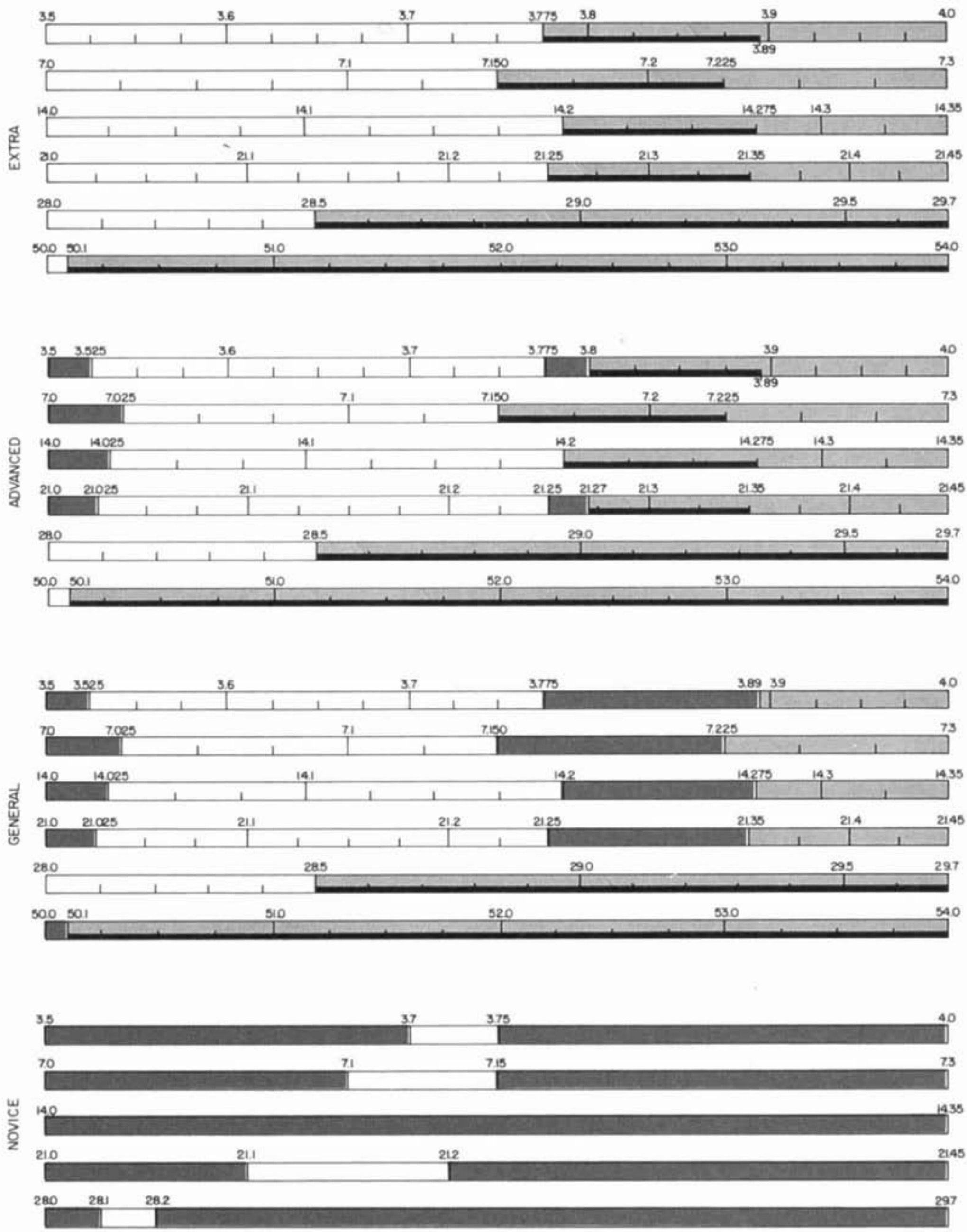

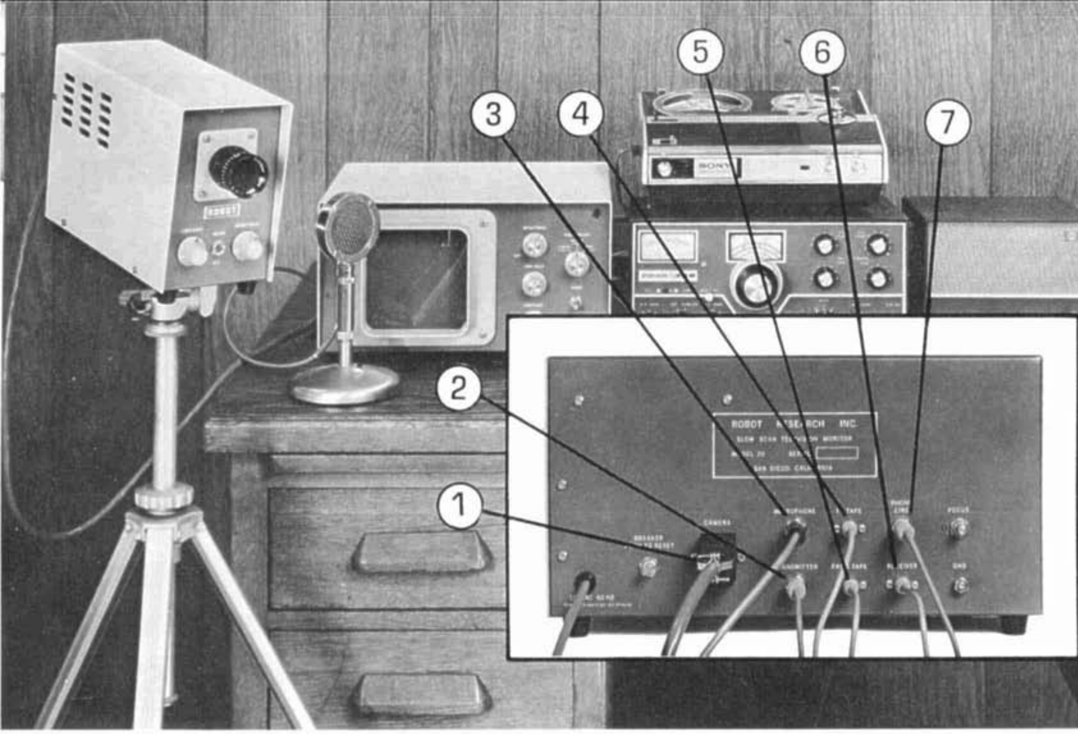
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