THE RADIO AMATEUR'S V.H.F. MANUAL

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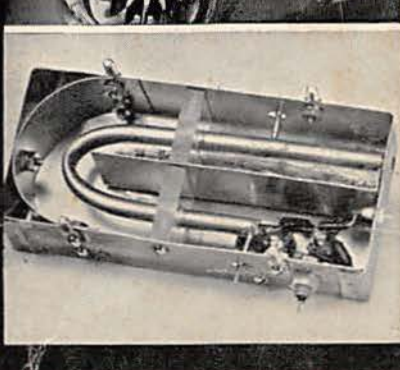
PRINCIPLES AND PRACTICE FOR THE WORLD ABOVE 50 MC.



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The Radio Amateur's V.H.F. Manual

A Manual of Amateur Radio Communication on the Frequencies Above 50 Megacycles

BY
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Foreword

Few essentially technical works enjoy the enthusiastic response that greeted *The Radio Amateur's V.H.F. Manual* upon its introduction some three years ago. V.h.f. amateurs of all shades of skills and interests found it what the author intended it to be: ". . . a book largely about things that work, and the principles behind them—a distillation of a generation of practical experience in the v.h.f. realm." The Manual quickly became a best seller, and it has remained so to this day.

Perhaps this success is explained by the fact that this is more than a reference book for the v.h.f. enthusiast. To illuminate the technical chapters to follow, it includes the first detailed history of "the world above 50 Mc." ever written. It explores the vastness and potential of the upper reaches of the radio spectrum, bringing to many readers a new appreciation of the true worth of this great radio resource. In terms understandable to the newcomer, yet useful to the experienced worker, it discusses the problems every v.h.f. enthusiast encounters, and spells out practical solutions.

It abounds in equipment projects for the fellow who likes to build his own gear. This second edition has a number of solid-state items not available elsewhere, yet there is much that will be helpful to the man who likes his equipment ready-made. There are some 70 pages of information on antennas, the worth of which is well substantiated by the outstanding records made by "W1HDQ antennas" in Antenna Measuring Parties held all over the country. V.h.f. mobile work, techniques and practice for the microwaves, solutions for the v.h.f. man's TVI problems, simple test gear you can make and use effectively—these are more of the ways by which this Manual has established itself as a major item on the list of ARRL publications.

In launching Edition One we asked your help in making it a better book. We got it, in letters by the hundreds, largely full of praise, but including many helpful suggestions and criticisms. These were studied carefully in preparing this Second Edition. As you glance through its pages for the first time, you'll note that there are quite a few new items. Upon more careful examination, it will be seen that the entire text has been gone over, word by word, line by line, to improve and update it in countless ways. We hope that you'll find it even more valuable than its predecessor.

As before, it is the work of *QST*'s long-time v.h.f. editor, Edward P. Tilton, W1HDQ. Contributions of many other dedicated v.h.f. men are gratefully acknowledged in the text.

John Huntoon
General Manager, ARRL

Newington, Conn.

Ross A. Hull, v.h.f. pioneer, and QST Associate Editor, 1931-1938.

Ross saw the potential of the then-uncharted world above 50 Mc. perhaps more clearly than any other man of his time. The technical excellence of his equipment designs and his enthusiasm in print and in person fired the imagination of a whole new generation of radio amateurs, among them the author of this book. His discovery and eventual explanation of tropospheric bending of v.h.f. waves has been called "one of the truly outstanding examples of scientific achievement by an amateur in any field of human endeavor."

CONTENTS

How It All Started

Those of us who make the frequencies above 50 Mc. our principal stamping ground tend to think of v.h.f. as the "new frontier" of amateur radio. Actually it is as old as the art of radio communication itself. While universal use of the upper reaches of the radio spectrum is a fairly modern phenomenon, some of the earliest work with electromagnetic radiation, and perhaps the first actual communication by radio, were on wavelengths near our present 2-meter band.

The resonator of Heinrich Hertz, and the practical applications of it by Marconi, operated around 150 Mc. And if you think that the beam you are using is a recent development, consider the fact that Hertz used a rudimentary form of Yagi in 1888, and Marconi employed a parabolic reflector to extend the range of his first equipment before the turn of the century. But Marconi and a generation of radio pioneers to follow him moved to the longer waves to achieve greater coverage. The ultra-high frequencies lay dormant for 20 years thereafter.

The Drive for DX

ARRL and *QST* had been in being for nearly ten years before frequencies higher than 15 Mc. were discussed in any detail. Transmitters using vacuum tubes had replaced spark rigs in the early '20s, and the unfolding possibilities of DX on wavelengths below 200 meters caused adventurous amateurs to probe ever higher in frequency. Each move upward produced new miracles of DX, culminating in worldwide communication with low power—in *daylight*—when gear was finally made to work in the 14-Mc. region. The next band then open to amateurs was at 56 Mc. It was widely assumed that if workable 5-meter gear could be built, this band would be even better for DX than 80, 40 and 20 meters had progressively turned out to be.

Just getting there was thought to be the principal problem. Technical Editor R. S. Kruse pointed the way in the October, 1924, issue of *QST*, with "Working At 5 Meters," perhaps the first v.h.f. constructional article ever published.[1] In the next few years much *QST* space would be devoted to 5-meter gear, but trying to use it was a frustrating business. Transmitters were simple oscillators; stabilization of any kind was all but unknown, and not even considered for 5-meter rigs. Receivers were regenerative detectors—hard to get going at all, and then incredibly cranky to tune. Oscillators using debased tubes, mounted bottom-up to reduce lead inductance, and receivers with foot-long insulated tuning shafts to hold down hand capacity, were the order of the day. The wonder was that hams of the middle Twenties made gear work on 7 or 14 Mc., let alone 56! [2, 3, 4]

The footnotes refer to the bibliography at the end of the chapter.—*Editor.*

This was c.w., remember. You had to chase a wandering signal with a receiver that was touchy beyond belief, even with stable signals. But 5-meter gear was made, and it worked after a fashion. When the first pioneers heard one another across town they were ready for a shot at Australia, or Europe. This technique had brought results before on lower frequencies, why not on 56 Mc.? [3] Then came several years of largely fruitless effort. There were scattered "heard" reports, some rather dubious in the light of present knowledge of v.h.f. propagation, but rarely was there two-way communication over more than a few miles.

By 1928, interest lagged. There were rumblings of DX on our new band at 28 Mc. The DX drive tended to move lower in frequency, and for about two years the world above 50 Mc. was inhabited mainly by experimenters, rather than communicators.

Short-Range Phone Does It

Up to this time, most amateurs were code men. Phone was coming in, but it was frowned on as wasteful of frequency and was often

Working at 5 Meters

By S. Kruse, Technical Editor

LAST month I said that ordinary methods worked perfectly well down to 20 meters but special care was needed below that. Since that time hundreds of stations have been working at 40 and 80 meters, not very many at 20 meters—and very few indeed at 5 meters. Most of the 20 and 5 meter work has failed because of an unsteady wave which could not be read, altho very strong at the receiver. The moral is to make a 5-watt tube work steadily rather than to make a 250-watt tube work unsteadily.

Getting Down

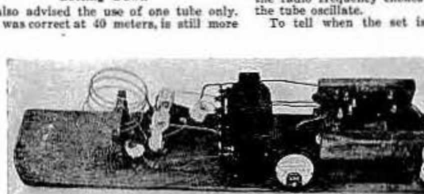
I also advised the use of one tube only. This was correct at 40 meters, is still more cuit is shown because it is simple. Series feed is used because this is to be a loose-coupled set and therefore no harm will come from series feeding. Shunt feed can be used but there will be more trouble in making the chokes work well.

Now we are down to a very small helix, and no capacity except that of the tube—can we make this thing oscillate?

The Circuit

The complete circuit is shown in Fig. 2. A little study will show this to be the same circuit as in Fig. 1, with the addition of the radio frequency chokes needed to make the tube oscillate.

To tell when the set is oscillating the



A 5 METER OSCILLATOR SET COMPLETE

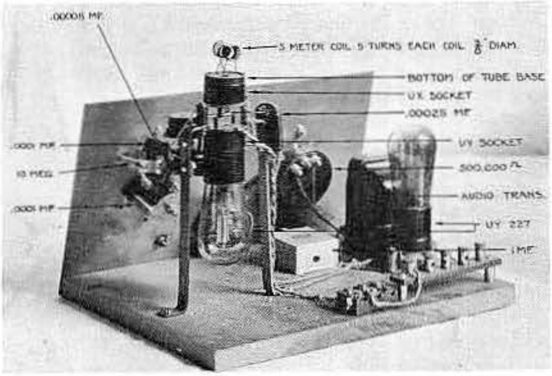
At the left is the 3" helix and next to it the baseless C-302 tube with its chokes and condensers. Beyond that is the Acme filament transformer, then a Weston 100-mil. D.C. instrument used to measure the plate current. The plate power is supplied by the home-made transformer at the right. The filament voltage is controlled by the E210 Bradleystat at the right end of hs board. This rheostat is connected in the primary circuit of the filament transformer. The circuit is shown in Fig. 2.

important at 20 meters, and it is almost out of the question to make several American tubes work in parallel at 5 meters. The reason for this is the insistence of our tube makers in bringing all terminals out in a bunch at one end of the tube. If they would only bring the plate out somewhere else we would have little trouble. The simplest test is to touch the plate coil with a wood handled screwdriver. Be careful—the burns from even a 5-watt oscillator are pretty painful.

If the screwdriver does not spark try raising the plate and filament voltage a bit, then try putting the plate and grid turns a bit closer together. If the screwdriver

In probably the first v.h.f. constructional article ever published, Technical Editor Kruse described a 5-meter oscillator in QST for October, 1924. The oscillator tube, barely visible just to the right of the tank coil, was a debased C-302 resting bottom-up on its glass envelope.

treated as an unwanted stepchild. Then a few u.h.f. experimenters began modulating their rigs, and unwittingly triggered off a boom that was to establish the 5-meter band as desirable communications territory in the minds of a whole new generation.



Autodyne receiver used by W8AZL in pioneering 5-meter work with W8PK and W8ABX. V.h.f. adaptation of the super-regenerative detector, which was to popularize v.h.f. operation in a big way, was still a year away. From September, 1930, *QST*.

The experience of early 5-meter phone experimenters John Long, W8ABX, and E. O. Seiler, W8PK, was typical. They were working on 80 one night in the summer of 1930, when a thunderstorm not far away made communication difficult. W8ABX was running his 5-meter rig simultaneously, and when W8PK listened on 5 he found, to his amazement, that signals were far clearer than on 80. Here, for the first time, the 5-meter band was seen to have real worth: it would work over short paths with voice, when noise levels, high activity, or other adverse factors were present on lower frequencies.[5]

Receivers were still a bottleneck, however. The regenerative receiver, critical enough on any frequency, was an operator's nightmare at 56 Mc. Enter here the superregenerative detector. Invented years before, the superregen had not found much favor with amateurs.[6] It was useless for c.w., and its broad frequency response and raucous audio quality gave it a bad name in voice work. But on 56 Mc., where there was band width to burn, broad tuning was almost a blessing.

In retrospect we can see several factors combining to accelerate 5-meter interest as ham radio moved into the Thirties. The wedding of the modulated oscillator and the superregenerative detector would start things rolling again. The modulated oscillator sounded awful on selective receivers now in use on lower frequencies, and it used up more than its share of high-priority kilocycles. But there was plenty of room on 5, and the unstable signal didn't sound bad at all when received on the broad-tuning "rushbox."

Soon it was found that these two castoffs had something else in their favor: If stations were not too close together in frequency, their transmitters and receivers could be operated simultaneously. This was "duplex phone," a wholly new concept. Two hams could converse as easily by radio as over the telephone, or face to face. Duplex was an overnight sensation, with obvious advantages over c.w., and the monologue voice technique then used on lower bands.

The ill wind of economic depression blowing across the land had made thousands of hams idle. With much time and little money, they were ripe for a kind of hamming that could be carried on with makeshift gear, largely made by hand or with parts robbed from discarded radio receivers. Not only from junk sets, either; as 5-meter interest boomed, more than one family's radio listening time was rationed, while the ham of the house reddened the plates of Type 45 or 71A tubes, lifted from the broadcast set for service in a 5-meter oscillator. Simple low-cost gear; duplex phone; the thrill of something new, yet within the reach of nearly everyone—these were magnets that drew countless newcomers, including the author of these lines, into amateur radio in the early Thirties.

As all through the history of the hobby, *QST* struck the spark. The July 1931 issue was fat with v.h.f. lures. Technical Editor Lamb had 11 pages on u.h.f. oscillators,[7] some working as high as 400 Mc., where some farsighted administrator had set aside a narrow band for amateur experimenters. Associate Editor Ross A. Hull, who would become one of the v.h.f. man's legendary heroes, fanned the flame with down-to-earth 5-meter receivers,[8] adding reports on *mobile* receiving tests, for good measure. Hull's "Duplex Phone on 56 Mc." in August [9] described more simple gear, and set forth the operating concept that was to build the fire to conflagration proportions.

The urge to work around the world was forgotten; the aim now was to work across town,

Cover picture from August, 1931, *QST* shows the modulated-oscillator transmitter that helped trigger the 5-meter boom of that eventful summer. Two 71As in a push-pull oscillator were modulated by parallel 47s.

reliably, on voice. If 5-meter waves traveled only in straight lines, then we would get up on the hills to extend our horizons. "Working portable" became a great weekend outdoor sport wherever there were hams and mountains. Duplex suggested voice relaying, and hilltop stations worked up haywire patching systems, to hook valley dwellers up with stations they could not hear direct. Message relaying also boomed, and on weekends, points as far apart as New York and Boston were connected by snappy relay circuits, thanks to three or four hilltop portables along the way.[10]

DX Again

Almost from the first widespread 5-meter activity in the summer of 1931, it was seen that signals did not always travel strictly in straight lines. Stations not within pure line-of-sight range

Another QST cover shot shows Managing Editor W1SZ hauling up the 8-element beam that was to make 5-meter history at Selden Hill in the fall of 1934.

could communicate at times, and the degree to which this was possible was seen to vary. But most of us were too busy trying to improve our equipment to pay much attention to propagation. Then, in the summer of 1934, Ross Hull began experimenting with beam antennas.[11] Hanging a stick-and-wire system of 4 half waves in phase, with reflectors, over the porch roof at Selden Hill, West Hartford, Ross fired up a 200-watt oscillator and blazed away in the direction of Boston. Selden Hill was a fine location, with a clear sweep across the Connecticut Valley, but it was less than 300 feet above sea level. The horizon, 12 to 20 miles distant, was solid with hills 1000 feet and more in elevation, for 50 miles or so, before the countryside sloped down to the coastal plain of Eastern Massachusetts and Rhode Island.

This sort of path had never been bridged on 5 meters. The best DX previously worked by Ross and his associate, W1ANA, and members of the Headquarters Staff who worked with them at Selden Hill, had been a portable on a hilltop 35 miles away. All hands were justifiably excited, therefore, when use of the new directive array produced two-way contacts with many stations in Eastern New England, 75 to 125 miles distant.

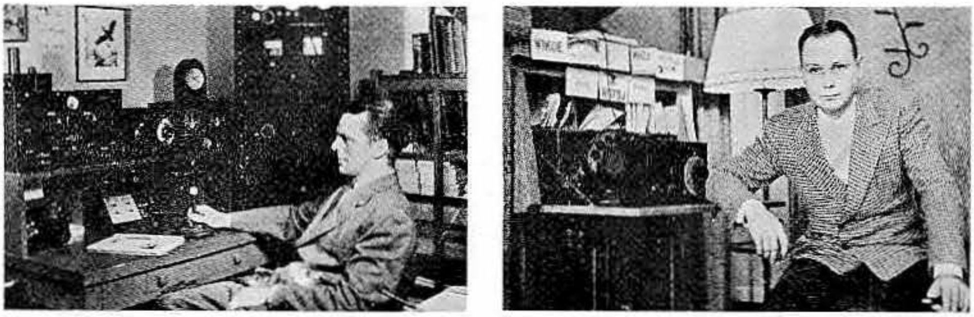
Operating on a 24-hour basis, the gang soon found that although signals of some sort were nearly always receivable over at least 100 miles, there was a tremendous variation in level. A signal would all but block the receiver at times, and then a day, an hour, or even minutes later it would drop almost into the noise.

Why? At this juncture what had been little more than a lark to the enthusiastic Hull took on the aspects of a long-term challenge. He set up schedules around the clock, and kept them religiously. He scanned reams of data on all kinds of natural phenomena. Eventually he developed a photographic recording technique, especially to compile a running record of the signal variations of experimental station W1XW, at the Blue Hill Observatory, near Boston. Records were plotted against lunar cycles, weather data, temperature and barometric pressure curves—anything that might affect 5-meter propagation.[12]

We won't attempt to retell the story here, for you can find it all in Ross' own words in *QST*. (See bibliography at the end of this chapter.) Read Hull's account for yourself, and see why his discovery of air-mass boundary bending of v.h.f waves, and eventual development of a theory to explain it,[13] has been called "one of the outstanding achievements by an amateur in any field of scientific endeavor."

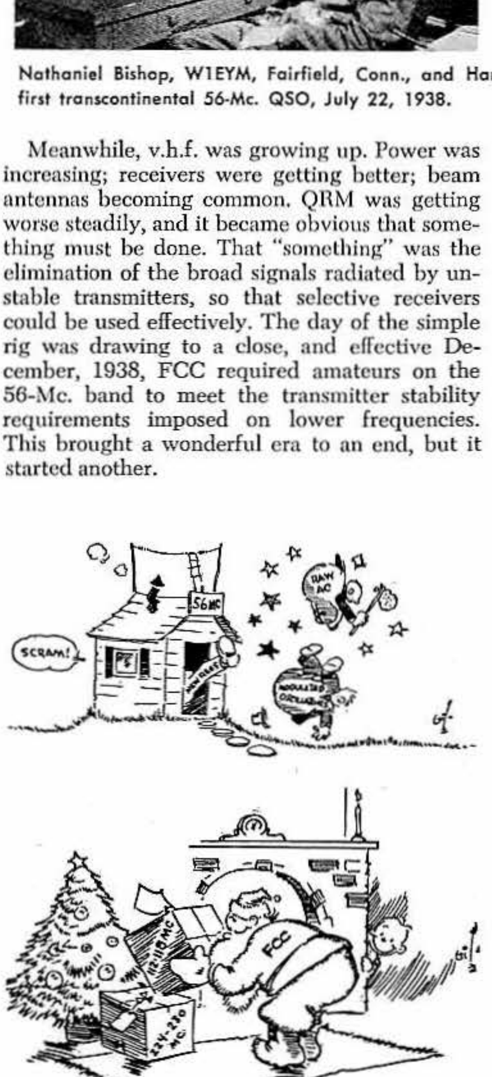
DX of a more remarkable variety was noted in the spring of 1935.[14] One day in May, 5-meter men in eastern cities were amazed to hear strangers in their midst; fellows signing W8 and W9 calls. Michigan, Ohio, Illinois—on 5 meters? Impossible—must be bootleggers trying to pull a fast one! But contacts were made and QSLs exchanged; the signals were genuine. It happened more often in 1936, and soon everyone was looking for 5-meter DX, as word spread like wildfire whenever the band opened up. Most contacts were betweeen 500 and 900 miles, with some out to 1200. A pattern for the propagation was evolved by Harvard University's J. A. Pierce, W1JFO, in a scholarly analysis of 5-meter DX in September, 1938, *QST*.[15]

In the same issue was the biggest DX news yet: details of a transcontinental QSO between W1EYM, Fairfield, Conn., and W6DNS, San Diego, Cal., a record that would stand as long as we had the 5-meter band.[16] Dr. Pierce charged this all up to sporadic ionization in the *E* region of the ionosphere, and pointed out some possible causes. His writings and those of Ross Hull are real milestones in the amateur's contribution to radio propagation knowledge. We can be proud that they came years in advance of similar work by scientific and governmental agencies.



Nathaniel Bishop, W1EYM, Fairfield, Conn., and Harold Hasenbeck, W6DNS, San Diego, Cal., made the first transcontinental 56-Mc. QSO, July 22, 1938.

Meanwhile, v.h.f. was growing up. Power was increasing; receivers were getting better; beam antennas becoming common. QRM was getting worse steadily, and it became obvious that something must be done. That "something" was the elimination of the broad signals radiated by unstable transmitters, so that selective receivers could be used effectively. The day of the simple rig was drawing to a close, and effective December, 1938, FCC required amateurs on the 56-Mc. band to meet the transmitter stability requirements imposed on lower frequencies. This brought a wonderful era to an end, but it started another.



Two of the famous "Gil" cartoons in *QST* for December, 1938, summed up the v.h.f. aspects of new FCC regulations then becoming effective.

Simple Gear Moves to 112 and 224 Mc.

Some 5-meter men saw in the stabilization regulations the death knell of their favorite band. A few gave up ham radio, rather than convert to the more complex stabilized transmitters. Others dropped to lower bands. But a sizeable number moved higher, and picked up on 112 Mc. where they had left off on 56. Using largely the same techniques that had served so well on 5, these fellows and many newcomers went through a cycle of activity and development reminiscent of early 5-meter days. Almost anything that had been done on 5 was repeated on 2½, except for the working of long distances via ionospheric propagation. There was even considerable experimentation on 1¼, and interest in both bands was still rising at the time of the World War II close-down, December, 1941. Freed of their severe QRM problem after 1938, 5-meter workers concentrated on improving receivers, transmitters, antennas and operating techniques. Reliable operating ranges stretched out to 200 miles and more, for the better stations.

One more DX propagation mode would be brought to light before the wartime close-down. Who did it first, where, or when, is not precisely known, but the first rumblings were heard in 1937. "Rumblings" is the right word, for these weird signals were characterized by terrific distortion, rendering modulation of any kind almost unreadable. Often the signal was little more than a rumble or roar. At first nobody knew what was up, but before long it was found that distortion and signal strength peaked when antennas were aimed north. Eventually this happened on a clear night, and all across the northern sky was seen the eerie glow of aurora borealis! Not much was known about the nature of the aurora then, but amateur 5-meter observations were put to good use in studying it. Until the wartime cessation of activity, and again after 1945, amateur reports gathered by ARRL contributed in a significant way to increased understanding of auroral phenomena.[17]

The Modern Era

Except for limited use of the 112-Mc. band in the War Emergency Radio Service, all amateur activity came to an abrupt end on December 7, 1941. Military communications and radar, meanwhile, expanded all through the spectrum, employing frequencies and techniques hardly dreamed of by most hams. Especially in v.h.f. and microwaves great strides were made during the war period, and many among us doubted that amateur radio could ever catch up. But hams were in the thick of it, in laboratories and in the field, and they learned their lessons well enough to be off and running when the war was over in the fall of 1945. A complete reshuffling of our allocations had been made, and after a temporary start on the old bands at 56 and 112 Mc., we moved to new assignments at 50, 144, and 220 Mc. We also now had bands at 420 Mc. and at intervals all through the assigned portion of the microwave spectrum. There was work to do!

The change from 56 to 50 Mc. was especially intriguing, in view of the rising solar activity curve. Would the new band "open up" when we reached the top of the sunspot cycle a year or two hence? By now, scientists were making predictions as to the maximum usable frequency for F_2-layer propagation, but they were not overly optimistic. The best guess was that 50 Mc. was a bit too high.

World-wide V.H.F. DX at Last!

Fortunately, most v.h.f. men did not know about these predictions. Noting that British TV signals on 45 Mc. could be heard now and then, and hearing from keen observers across the Atlantic that American signals and harmonics were filtering through in Europe on frequencies as high as 47 Mc. on occasion, amateurs set up test schedules in the fall of 1946. On mornings when conditions appeared favorable, American 50-Mc. men transmitted toward Europe, listening for replies on 28 Mc., there being no 50-Mc. band in Europe.

Just before noon on November 24, 1946, a test transmission by W1HDQ, West Hartford, Conn., brought a frantic *"I'm hearing you on 50 megacycles!"* from G6DH, Clacton-On-Sea, Essex, England, and the first v.h.f. communication across the Atlantic was on. G5BY, near Plymouth, heard the test at the same time, joining G6DH in the transatlantic cross-band QSO a few minutes later. Shortly after noon the same day, W4GJO, Orlando, Fla., worked W6QG, Santa Ana, Cal., for the first transcontinental 50-Mc. F_2-layer QSO. Pacific DX came in January, 1947, when KH6DD worked J9AAK, Okinawa, extending the 50-Mc. DX record to 4600 miles.

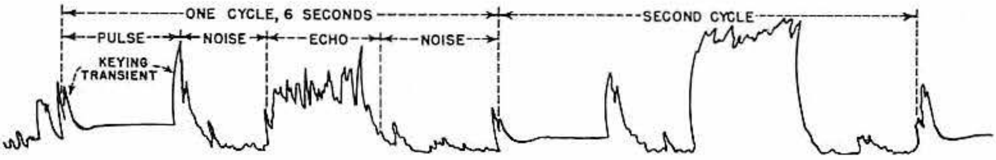
In March, 1947, W4IUJ, West Palm Beach, Fla., worked OA4AE, Lima, Peru, thereby winning the Milwaukee Radio Club trophy for the first two-way intercontinental v.h.f. QSO, the cup having been resting at ARRL Headquarters for nearly ten years. Cross-band DX of phenomenal proportions came at about this same time, resulting from checks made by PAØUN and PAØUM with ZS1P and ZS1T, 6000 miles to the south. August, 1947 brought a new two-way record, 5300 miles, between W7ACS/KH6 and VK5KL.

Though not credited as such at the time, this probably was the first DX QSO via a propagation medium that was to be exploited later on the Mexico-to-Argentina path. Around the end of August, 1947, XE1KE, Mexico City, began working LU6DO, Temperly, Argentina, and other LUs on 50 Mc. These contacts were made later in the day than F_2-layer predictions called for. Sometimes propagation lasted well into the evening hours, an unheard-of thing on frequencies this high. Eventually labeled ***transequatorial scatter,*** this mode of propagation rates as one of the outstanding discoveries in amateur v.h.f. history.[18]

The New Bands Prove Their Worth

Progress on 144 Mc. and higher frequencies was also notable in the early postwar years. Aided by the availability of the SCR-522 and other military communications units on the surplus market, 2-meter men converted largely to stabilized equipment and selective receivers, and operating ranges expanded rapidly. Tropospheric propagation was found to be more favorable on 144 than on 50 Mc., and the record for two-way work was extended gradually,

Eileen and Denis Heightman of G6DH, British end of the first transatlantic v.h.f. QSO, November 24, 1946. Receiving W1HDQ on 50 Mc., Denis replied on 28 Mc. The following year, operating with special temporary authorization, G6DH was the first British station to work two-way across the Atlantic on 50 Mc.



Visual record of an historic achievement—the first amateur signals sent to the moon and back. After three years of work, Ross Bateman, W4AO, and William L. Smith, W3GKP, shown here checking alignment of the huge stacked-rhombic array at W4AO, finally received echoes of their 144-Mc. signal reflected from the moon. The date: January 27, 1953.

reaching 1400 miles by 1951. Auroral communication was found to be possible on 144, and this mode provided much exciting 144-Mc. DX. Exploitation of the reflecting properties of ionized trails of meteors opened the way to more 2-meter DX. Two leaders in this field were W4-HHK, Collierville, Tenn., and W2UK, New Brunswick, N. J., who received the ARRL Merit Award of 1955, for their outstanding meteor-scatter work of 1953 and later.

The 220-Mc. and 420-Mc. bands had appeal for the experimentally inclined, and were soon shown to have great value for practical communications purposes as well. Development of efficient equipment and high-gain antennas showed that these bands were capable of reliable coverage nearly approximating that of 50 and 144 Mc.

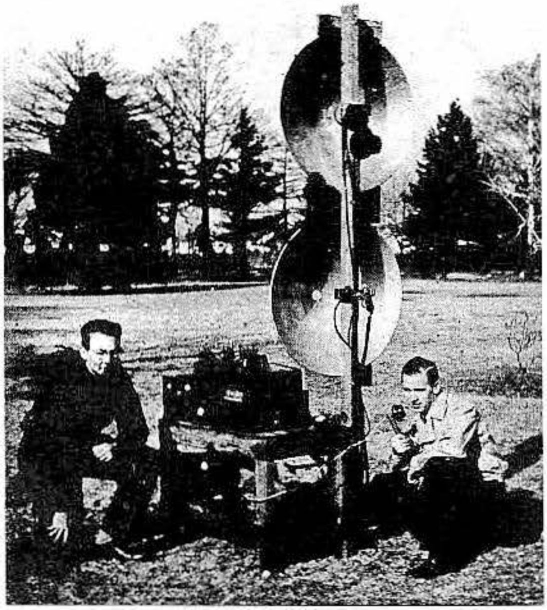
Making use of tubes and components largely salvaged from war-surplus radar and navigational equipment, amateurs developed workable communications gear for all our microwave bands before the end of 1946, and in later years were able to extend communications distances out to several hundred miles on nearly all our u.h.f. and s.h.f. bands. Development along these lines continues to this day.

Intrigued by the possibilities of weak-signal work, amateurs made notable strides in utilizing various marginal modes of propagation such as moon reflection, ionospheric and tropospheric scatter, and even satellite communication. The first successful use of the moon for the reflection of amateur signals was accomplished by W4AO and W3GKP in January, 1953.[19] These two used advanced techniques on 144 Mc. to demonstrate that lunar communication was at least a possibility for amateurs. Two-way communication via the moon was a long time coming, and was finally achieved first on 1296 Mc. The work of W1BU and W6HB in communicating over 2500 miles by way of the moon in July, 1960, is a notable milestone.[20]

Propagation know-how paid off markedly for

W6NLZ, Palos Verdes Estates, Cal., and KH6-UK (also W2UK, mentioned above), Kahuku, Hawaii, when they were able to work across 2500 miles of the Pacific on 144 Mc. in July, 1957.[21] This was the longest path ever covered by tropospheric means by any communications service, and as such it achieved world-wide acclaim. Not satisfied, Chambers and Thomas went on in subsequent years to bridge the path on 220 and 432 Mc. Their superb work won for them the Edison Award for 1960, the only instance in which this award was given for scientific accomplishment.

Thus we have touched lightly on some highlights of amateur radio's long history of pioneering the use of frequencies once thought to be useless for any practical purpose. It is well for all of us, hams of the present and future, that we have this record of achievement behind us. In the years to come, the pressure on all frequencies above 50 Mc. is certain to rise, as if it were not already high enough. Every kilocycle, even to frequencies only dreamed of a few years ago, is eyed eagerly by many users of the radio spectrum. We have shown, from the earliest times, that it is good for everyone that amateurs have access to samplings of the radio-frequency spectrum, from bottom to top, whatever that may be. To continue to merit the confidence and support of the people and agencies who will decide future allocation of frequencies should forever remain one of our highest aims.



The first amateur microwave station. A. E. Harrison, W6BMS/2, and Reuben Merchant, W8LGF, built two of these stations and had them ready for communication on November 15, 1945, the day that our microwave bands were opened to amateur use. Frequency: 5600 Mc.

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Summaries of activity appear throughout 1932 issues.

Behavior of U.H.F. Waves . . . Jones, March, 1933.

Graduating to Oscillator-Amplifier Transmitters for 56 Mc. . . . Griffin, W2AOE, May, 1933.

Firing Up on the Newly-Opened Ultra-High Frequencies . . . Hull, September and November, 1934.

[11] *Extending the Range of U.H.F. Stations* . . . Hull, October and December, 1934.

[12] *Air-Mass Conditions and the Bending of U.H.F. Waves* . . . Hull, June, 1935.

[13] *Air-Wave Bending of U.H.F. Waves* . . . Hull, May, 1937.

[14] *Five-Meter Signals Do the Impossible* . . . August, 1935, was first published report of authenticated 5-meter skip. July issues of 1936 and 1937 contain summaries of reported DX.

[15] *Interpreting 56-Mc. DX* . . . Pierce, September, 1938 (E-layer theory)

[16] *Further Reports of 50-Mc. DX* . . . September, 1938.

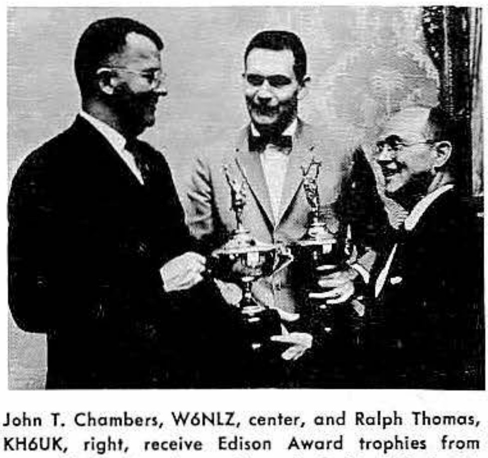
[17] Moore, "Aurora and Magnetic Storms," June, 1951.

[18] Cracknell, "Transequatorial Propagation of V.H.F. Signals," December, 1959.

[19] Tilton, "Lunar DX on 144 Mc.," March, 1953.

[20] September, 1960.

[21] September, 1957, p. 62.



John T. Chambers, W6NLZ, center, and Ralph Thomas, KH6UK, right, receive Edison Award trophies from General Electric vice-president L. Berkley Davis, in Washington ceremony, February 23, 1961. Award was in recognition of the transpacific communication by these outstanding amateurs on 144, 220 and 432 Mc.

Regular coverage of the v.h.f. and higher bands, *On the Ultra-Highs*, began in December, 1939. Later called *The World Above 50 Mc.*, it has told the month-by-month story of amateur v.h.f. progress ever since.

A Vast Resource

AMATEUR BANDS ABOVE 50 MC.

The true extent of the frequencies assigned to amateurs above 50 Mc. is rarely understood, even by those who spend most of their operating time there. Below are listed all amateur v.h.f., u.h.f., and microwave bands, together with the emissions we may use in each. Something of what these segments of the spectrum are good for will follow shortly. Let's just *look* at the table, first.

50 to 54 Mc.
- 50.0 to 50.1 Mc.—A1 (c.w. telegraphy) only.
- 50.1 to 54 Mc.—A1, A2 (tone-modulated telegraphy), A3 (amplitude modulation and narrow-band f.m.), A4 (facsimile).
- 51.0 to 54 Mc.—AØ (unmodulated carrier; duplex communication), plus above.
- 52.5 to 54 Mc.—Wide-band f.m., plus above.

144 to 148 Mc.
- 144.0 to 147.9 Mc.—AØ, A1, A2, A3, A4, f.m. (wide-band or narrow-band).
- 147.9 to 148 Mc.—A1 only.

220 to 225 Mc.
All above modes of emission. See U.S. Regs., 97.61.

420 to 450 Mc.
All above modes, plus A5 (television). See U.S. Regs., 97.61.

1215 to 1300 Mc.
All above modes.

2300 to 2450 Mc.
All above modes, plus pulse.

3300 to 3500 Mc.
All modes.

5650 to 5925 Mc.
All modes.

10,000 to 10,500 Mc.
All modes except pulse.

21,000 to 22,000 Mc.
All modes.

40,000 Mc. and all higher
All modes.

Technician Class licensees may use the entire 50-Mc. band until Nov. 22, 1968,* 145 to 147 Mc., and all higher amateur frequencies, with full amateur privileges. *Novice Class* may use 145 to 147 Mc., with crystal-controlled transmitters only, with no more than 75 watts input to the final stage. Novice use of voice rescinded after Nov. 22, 1968.* All bands above 220 Mc. are shared with the Government Radio-Location Service, which has priority. Operation in the 220-Mc. band is restricted in parts of Texas and New Mexico, and final-stage input in the 420-Mc. band is limited to 50 watts in Florida, Arizona, and parts of Alabama, Texas, New Mexico, Nevada and California, as set forth in *Part 97.61* of the *U. S. Regulations*. Permission to use 1000 watts may be obtained by amateurs in the restricted areas by individual application to FCC.

*For full information on band subdivision by class of license, see current edition of the *A.R.R.L. License Manual*.

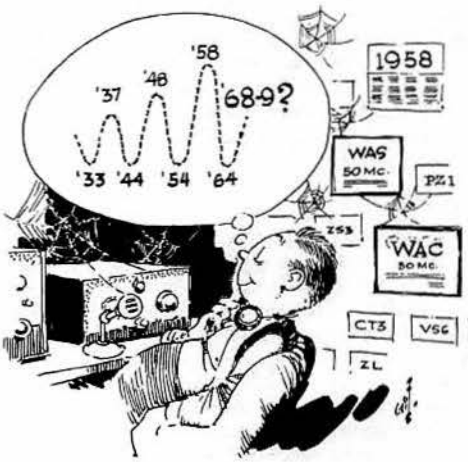
WHAT CAN WE DO HERE?

In terms of kilocycles and potential occupancy by amateurs, this is a world almost beyond comprehension. As seen in Fig. 2-1, amateur bands from 80 through 10 meters, which carry most of the occupancy load, total 3300 kilocycles—less than the width of any v.h.f. band. V.h.f. assignments include 13 Mc., or almost four times the frequency spread of all lower amateur bands combined. The 420-Mc. band is wider than the *entire spectrum* from d.c. to the top of the 10-meter band. Each band above 1000 Mc. is wider still. Our inability to show these figures in scale is worth remembering when we worry over congestion in the amateur bands between 3.5 and 30 Mc.!

Our historical review, Chapter 1, emphasized the amateur's role in uncovering the true worth of the v.h.f. bands. Our potential in this field is far from exhausted. Though great scientific strides have been made, by no means all is known of the ways by which signals in the v.h.f. and higher frequency ranges are propagated to distant points. Nature still surprises even the best-informed amateur, and admittedly this is a factor in the appeal of the world above 50 Mc. Knowing something of propagation media we can, however, take advantage of the opportunities nature affords us, and we will enjoy our work more and do it better than if we merely take what comes our way, without question or observation. Here are some propagation tips, band by band.

50 Mc.

Perhaps no band is more interestingly placed in the radio spectrum than this, from the standpoint of propagation vagaries. Working in borderline territory between the "DX bands" and



those normally considered useful mainly for local communication, the 50-Mc. enthusiast samples both worlds. Though DX is not his daily lot, he will see, at one time or another, nearly every known form of long-distance propagation. His reliable range with moderate power and relatively simple equipment is likely to equal anyone's, for the 50-Mc. region is less susceptible than lower frequencies to adverse effects that tend to break up or impair communication. Consistent coverage over a radius of 100 miles or more is not unusual, and this can be extended considerably by use of optimum equipment and communications techniques.

Variety is frequently afforded by *tropospheric bending*, which extends local coverage by two to three times the normal. *Sporadic-*E *skip* offers DX in the range of 400 to 1200 miles or so, and multiple-hop effects may extend this up to 2500 miles or more on occasion. *Auroral propagation* to all distances up to 1000 miles is fairly common in the high latitudes. *DX via the F_2 layer* may be possible during the peak years of the sunspot cycle, providing contacts at distances of 2000 miles and more. *F_2-layer backscatter* fills in the shorter distances at these times. *Ionospheric scatter* and *reflections from meteor trails* afford the proficient operator chances for work over 600 to 1200 miles on a regular basis on 50 Mc. *Transequatorial propagation* is good for several thousand miles, in low latitudes and in periods of high solar activity. More about each of these modes later.

144 Mc.

Except that it lacks some of the long-distance ionospheric possibilities, the 144-Mc. band is not unlike 50 Mc. Tropospheric propagation tends to improve with frequency, so 144 Mc. is superior to 50 in this respect. Whereas tropospherically-propagated 50-Mc. signals are seldom heard beyond 300 miles, 144-Mc. work out to 500 miles or more by this mode is fairly common. Up to 1400 miles over land and 2500 miles over water have been covered by tropospheric bending on 144 Mc.

Sporadic-E skip is rare on 144 Mc., though lack of alert observers in the more favorable areas may have caused us to miss some 144-Mc. DX opportunities of this kind in the past. Auroral propagation is quite similar to 50 Mc., except that borderline conditions may show on the lower frequency and not on 144 Mc. Distances up to 1300 miles have been covered, but 200 to 700 miles is most common. Use of c.w. is almost a necessity because of the high degree of distortion produced by the auroral reflection.

Of the rare modes, meteor scatter and tropospheric scatter have been most exploited by 144-Mc. operators. Each requires fairly high transmitter power, skill in the use of c.w., and the best possible receivers and antennas. Communication by way of the moon is just possible on 144 Mc., and considerable progress was made in e.m.e. work in the 1960's by W6DNG, K6MYC, VK3ATN and others. Similar techniques employed with man-made reflecting satellites show some promise.

220 Mc.

This band is similar to 144 Mc. in its tropospheric propagation possibilities. The overland record is about 900 miles, and the 2500-mile path from the West Coast to Hawaii has been bridged with good signals. No ionospheric propagation has been observed. Auroral conditions are less favorable than on 144, but some DX of

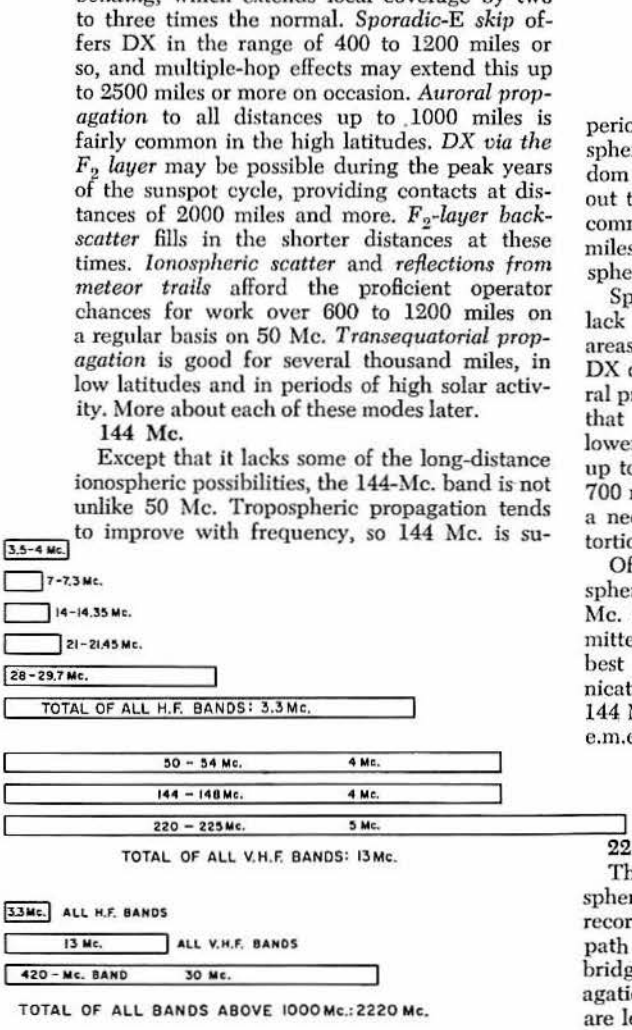


Fig. 2-1—Amateur bands at the upper end of the r.f. spectrum defy portrayal in scale. At the top are our h.f. bands, which total 3300 kc. (3.3 Mc.) in width. Next below, on the same scale, are the three amateur v.h.f. bands, each wider than all h.f. bands combined.

A new scale is needed to show these bands in relation to our 420-Mc. band, and this in turn fails to indicate the scope of amateur assignments above 1000 Mc. These would require 70 strips the size of the one shown for the 420-Mc. band, which is itself wider than the whole r.f. spectrum from d.c. through 30 Mc.!

this kind has been worked, mostly under 700 miles. At this writing, no amateur work via the moon or satellites has been done on this band, but possibilities appear at least equal to those on 144 Mc. More universal activity is needed on this and all higher amateur bands to assess their real worth for long distances.

420 Mc.

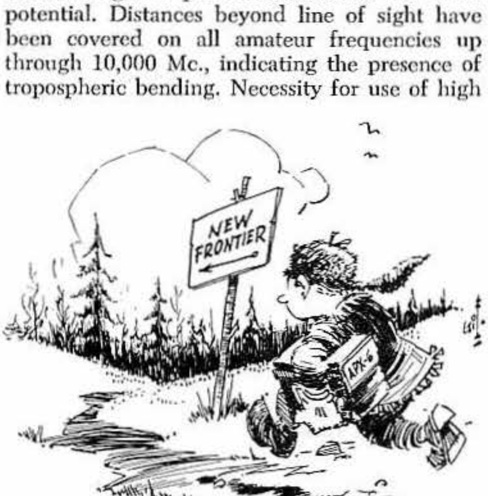
Exploitation of this band suffered because of the power restrictions imposed until recently, but it is known that tropospheric possibilities are excellent. The terrestrial two-way record, 1150 miles across the Gulf of Mexico, is an example. The West-Coast-to-Hawaii path has been covered one way on 432 Mc. with strong signals. Lunar and satellite possibilities appear better than on lower frequencies, and distances as great as New England to Hawaii and California to Europe have been covered, by way of the moon.

1215 Mc.

Though largely unexploited by amateurs thus far, the frequencies above 1000 Mc. offer vast opportunities for interesting work. Here is the true "frontier"—a world we must explore if the traditions of amateur pioneering are to be maintained. Distances up to 400 miles have been worked on 1215 Mc. with low power under conditions of tropospheric bending, and much greater distances certainly are possible. Reflection from the moon shows great promise; see records listed below. Though complex and expensive equipment is needed, results indicate that any two points on the earth's surface where the moon can be seen simultaneously are not beyond the possible range of this mode of 1215-Mc. communication.

2300 Mc. and Higher

Amateur experience in our microwave bands is too meager to permit us to assess their true potential. Distances beyond line of sight have been covered on all amateur frequencies up through 10,000 Mc., indicating the presence of tropospheric bending. Necessity for use of high



antenna gains with resultant sharp beam patterns, almost rules out the random operation that characterizes amateur radio on lower frequencies. Our microwave assignments have tremendous potential for point-to-point communication, and they might well supplement lower frequencies for scheduled work.

Use of pulse modulation is one means by which the microwaves can be put to practical use by amateurs this being usable on all our frequencies above 2300 Mc. except in the 10,000-Mc. band.

The existing DX records for each of our bands above 50 Mc. are listed below:

Terrestrial Two-Way Records

50 Mc.: LU3EX—JA6FR
12,000 Miles—March 24, 1956

144 Mc.: W6NLZ-KH6UK
2540 Miles—July 8, 1957

220 Mc.: W6NLZ—KH6UK
2540 Miles—June 22, 1959

420 Mc.: W5LUU—WA4KFW
1150 Miles—April 13, 1965

1215 Mc.: W6DQJ/6—K6AXN/6
400 Miles—June 14, 1959

2300 Mc.: W1EHF/1—W2BVU/1
170 Miles—July 13, 1963

3300 Mc.: W6IFE/6—W6VIX/6
190 Miles—June 9, 1956

5650 Mc.: WA6KKK/6—WB6JZY/6
179 Miles—October 15, 1966

10,000 Mc.: W7JIP/7—W7LHL/7
265 Miles—July 31, 1960

21,000 Mc.: W2UKL/2—WA2VWI/2
27 Miles—Oct. 24, 1964

E.M.E. Two-Way Records

144 Mc.: VK3ATN—K2MWA/2
10417 Miles—Nov. 28, 1966

420 Mc.: WA6LET—G3LTF
5370 Miles—Sept. 25, 1965

1215 Mc.: W1BU—KH6UK
5092 Miles—August 9, 1962

PROPAGATION BEYOND THE HORIZON

Radio waves travel in straight lines unless forced to do otherwise. In some respects, v.h.f. waves are less easily reflected than waves of lower frequency, so consistent v.h.f. communication with low power tends to be essentially local in character, covering only slightly more than line-of-sight distances. There are many ways by which the wave energy may be reflected, refracted or scattered, however, and the v.h.f. man will do well to become familiar with the principal ones at least. Some are shown in Fig. 2-2.

Tropospheric Bending is the most common form of v.h.f. DX.[1] Though observable on all radio frequencies, it is most pronounced in the v.h.f. range and higher. It is the result of change in refractive index of the atmosphere at the boundary between air masses of differing temperature and humidity characteristics. These boundaries occur in the first few thousand feet above the earth, so their effect is most prevalent at distances under about 150 miles, though it may extend much farther.

Air masses often move on a very large scale, retaining their original character over considerable periods of time. A large mass of cold air of polar origin may be overrun by warm air from the south. When this happens, an *inversion* is said to exist, the normal state of affairs being a 3-degree drop in temperature for each 1000 feet of altitude. Such a boundary may prevail for 1000 miles or more along a more-or-less stationary weather front, producing amazing DX in the v.h.f. and u.h.f. ranges. There is an easily-observed tie-in between visible weather conditions and v.h.f. coverage.[2] Daily weather maps published in many newspapers and often shown in rudimentary form in television weather broadcasts may help the v.h.f. enthusiast to anticipate favorable propagation.[3] Detailed weather maps may be obtained from the U. S. Weather Bureau on a subscription basis.

Tropospheric bending is most common in fair, calm weather of the warmer months, though it can occur at any season. Atmospheric convection in coastal areas, over the Great Lakes Basin, or in the valleys of major rivers, produces the required stratification of air, making these regions somewhat more desirable v.h.f. territory than irregular mountainous terrain, far inland. Though the experienced meteorologist would call the following advice an over-simplification of a complex picture, the v.h.f. man should watch for slow-moving areas of high barometric pressure, and concentrate on the trailing edges of such areas. See Fig. 2-3. Such favorable conditions are most often observed in the early fall months.

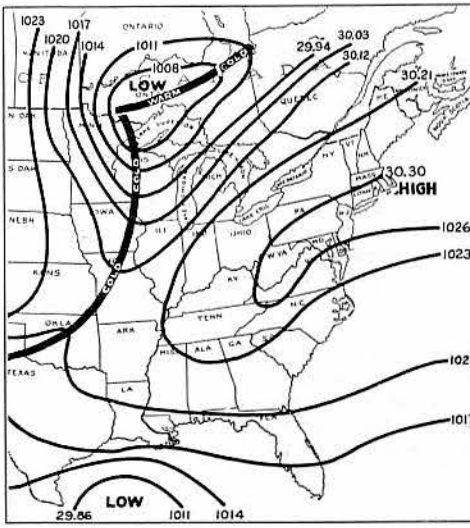


Fig. 2-3—A readily available guide to tropospheric propagation conditions is a weather map, showing pressure distribution and frontal lines. On the October day that this map appeared in eastern newspapers, the 2-meter band was open from Nova Scotia to at least North Carolina for several hours.

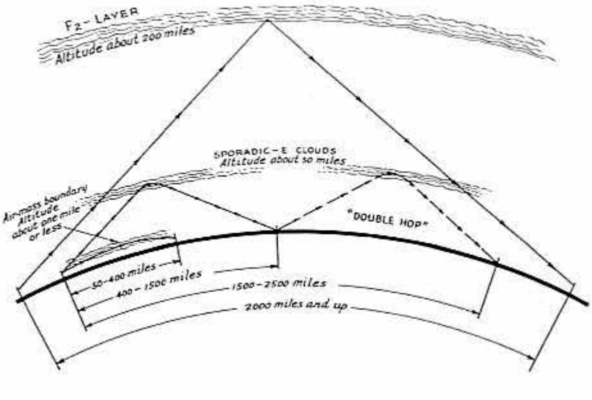


Fig. 2-2—The principal means by which v.h.f. signals may be returned to earth, showing the approximate distances over which they are effective. The F_2 layer, highest of the reflecting layers, may provide 50-Mc. DX at the peak of the 11-year sunspot cycle. Such communication may be world-wide in scope. Sporadic ionization of the *E* region produces the familiar "short skip" on 28 and 50 Mc. It is most common in early summer and late December, but may occur at any time, and regardless of the sunspot cycle. Refraction of v.h.f. waves also takes place at air-mass boundaries, making possible communication over distances of several hundred miles on all v.h.f. bands. Normally it exhibits no skip zone.

What is known as the *U. S. Standard Atmosphere* curve [4] is shown in Fig. 2-4, left. The solid line is the normal decrease in temperature with height. The broken line shows a relative humidity of 70 per cent, from ground to 12,000 feet. Figures in parenthesis are the ratio of grams of water vapor to kilograms of air, called the *mixing ratio*. No tropospheric bending would be observed under these conditions. At the right are upper-air readings of an inversion over Toledo, Ohio, on a September evening some years ago. On this occasion 2-meter signals were traversing the 750-mile path from Northern New Jersey to the Chicago area.

Particularly over water in the lower latitudes, and less often over land areas, something approximating a duct may form, in which v.h.f. waves are propagated over very long distances, following the curvature of the earth in the manner of u.h.f. energy within a waveguide. This *ducting* has accounted for much of our extreme v.h.f. and u.h.f. DX. Notable examples are the spanning of the Pacific from Southern California to Hawaii on 144, 220 and 420 Mc. by W6NLZ and KH6UK.[5] achievements which must always rank among the most significant in amateur v.h.f. annals.

Tropospheric communication on 144 Mc. along the Atlantic Seaboard over distances as great as New England to Florida, and overland contacts up to 1400 miles, are also products of tropospheric ducting. Scientific investigations over the South Atlantic have shown ducts capable of propagating signals on frequencies even below 50 Mc., but true ducting is rare in amateur experience below 144 Mc. It occurs most often in the u.h.f. region, and extensive occupancy of the amateur bands above 400 Mc. should enable us to exploit its possibilities more fully in years to come.

Sporadic-*E* Skip results from reflection of v.h.f. waves by dense patches of ionization in the *E* region of the ionosphere, roughly 50 miles above the earth. Causes are still not completely understood and its occurrence is predictable only in a general way, but its effects are well known to generations of v.h.f. enthusiasts.[6] Layer height and electron density determine the skip distance, but 50-Mc. propagation is most common over distances of 400 to 1200 miles. Often signals are very strong, though they may vary rapidly over quite wide ranges. Ionization may develop simultaneously in several areas, making multiple-hop propagation possible and extending the working range to as much as 2500 miles. Signals are usually heard from intermediate distances at such times.

E-layer v.h.f. propagation is most common in the months of May, June and July. There is a shorter season in December and January, and the effect may occur at random times throughout the year. The long and short seasons are reversed in the southern hemisphere. Duration and extent of E_s openings tend to be greater in the long season. June is the peak month ordinarily, with country-wide openings lasting for many hours at a time at this season. The early evening and before-noon hours are most productive.

The upper-frequency limit for sporadic-*E* is not known. It is observed fairly often up to about 100 Mc., and scattered instances occur in the 144-Mc. band. Increased activity on 144 Mc. has enabled amateurs to observe the effect in this band more often in recent years. Ionization develops rapidly, with effects showing first on lower frequencies. Observation of the 28-Mc. band, or commercial frequencies between 30 and 50 Mc.,[7] will usually give the 50-Mc. enthusiast some advance notice of an impending opening. Similarly, the condition of the 50-Mc. band or the v.h.f. f.m. or television frequencies may give clues as to the possibility of 144-Mc. propagation.[8]

As ionization density increases and the maximum usable frequency rises, the skip distance on a given frequency shortens. Thus, very short skip on 50 Mc. may portend a 144-Mc. opening.

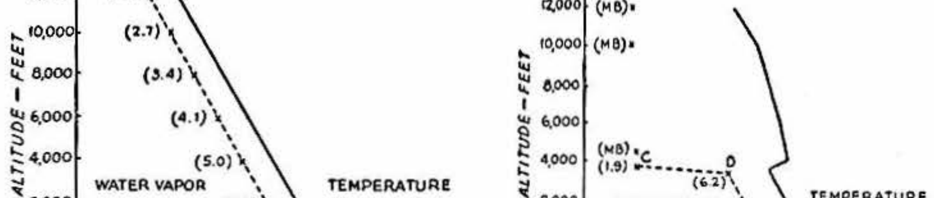


Fig. 2-4—Upper-air conditions that produce extended-range communication on the v.h.f. bands. At the left is shown the U. S. Standard Atmosphere temperature curve. The humidity curve (dotted) is that which would result if the relative humidity were 70 per cent from the ground level to 12,000 feet elevation. There is only slight refraction under this standard condition. At the right is shown a sounding that is typical of marked refraction of v.h.f. waves. Figures in parentheses are the "mixing ratio"—grams of water vapor per kilogram of dry air. Note the sharp break in both curves at about 4000 feet.

The alert observer should watch for short 50-Mc. skip near the midpoint of a potential 1200-to-1400-mile path, as the best indication of a 144-Mc. DX chance. Hearing a 50-Mc. station in Cincinnati working another in St. Louis, for example, would be a good omen for a 144-Mc. operator in Washington, D. C., indicating the possibility of 2-meter propagation to Oklahoma City or Wichita. Plotting observed skip on a map of the United States will help one to grasp the significance of what he hears.

Like most other v.h.f. DX modes, sporadic-*E* skip was discovered by amateurs (see Chapter 1) and it quickly became a popular sport among 5-meter men of the 1930s. To commercial users of the v.h.f. spectrum it is known mainly for its nuisance value, but to 50-Mc. men it is the DX mode supreme. Though the number and quality of openings vary somewhat from year to year, E_s propagation does not appear to be closely related to sunspot activity.

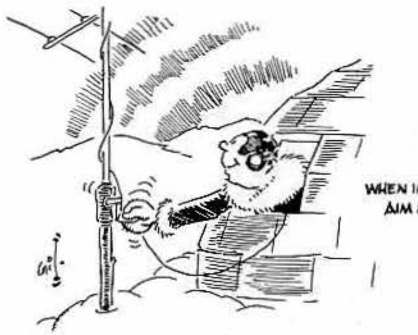
Auroral Propagation involves reflection of v.h.f. waves from the auroral curtain in the northern skies,[9] usually at acute angles. It is most common at 50 and 144 Mc., the number and duration of openings decreasing markedly at higher frequencies. Some auroral work has been done on 220 and 420 Mc. It may eventually become feasible above 1000 Mc., if very large arrays are used. Scientific investigations with very high power and large antenna arrays have shown auroral returns at frequencies of several thousand megacycles.

The reflecting properties of the aurora vary rapidly, with the result that the returned v.h.f. signal is badly distorted by multipath effects. Voice modulation is often unintelligible on 50-Mc. signals, and nearly always so at 144 Mc. Keyed c.w. is, therefore, the most effective mode of operation for auroral work. Suppressed-carrier s.s.b. is a poor second, followed by a.m., n.f.m., and wide-band f.m., in that order.

The number of auroras seen each year, and the opportunities for v.h.f. communication via the aurora, vary with *geomagnetic* latitude. Since the geomagnetic pole is near Thule, Greenland, geomagnetic latitude lines slope upward with respect to geographical latitude as we look to the west. Bangor, Maine sees many more auroras than does Seattle, though the latter city is farther north. New York, Philadelphia and Washington far outdo Reno and Northern California, which lie along the same geographical latitude. Aurora DX has been worked on 144 Mc. as far south as 30 degrees in Southeastern U.S.A., but seldom or never in El Paso, Phoenix or Los Angeles, all of which are well north of latitude 30.

Auroras follow seasonal patterns, being most common around the equinoxes (March and September). They may occur at any time, however, and summer and midwinter auroras are not uncommon in the more northerly states. Aurora effects are observed most often in the late afternoon or early evening, lasting for a few minutes to many hours. The southerly extent also varies greatly. Strong and widespread disturbances may peak in the early evening, drop off for about two hours before midnight, and then return, lasting until dawn or after.

The optimum heading for a v.h.f. antenna array varies with the position of the aurora, and may change rapidly, just as the visible display does. Usually an eastern station will work the greatest distance to the west by aiming as far west of north as possible, but this does not always follow. Constant probing with the antenna is recommended, especially if an array with a really sharp pattern is being used.



Developing auroral conditions may be observed by monitoring signals in the region from the broadcast band up to about 5 Mc. or so. If signals in the 75-meter amateur band, for example, begin to waver suddenly in the afternoon or early evening hours, taking on a dribbling sound, an auroral disturbance may be getting under way. Its effects will not be long in showing on the bands higher in frequency, if the disturbance is pronounced. Distortion of voice on 28 or 50 Mc., when the array is aimed north, is evidence that the effect has reached these bands, and it is time for the v.h.f. man to go to work on c.w.

On 50 or 144 Mc. the buzzing sound characteristic of an auroral return may be heard even on local signals, when both antennas are aimed north. The great-circle distance workable via the aurora extends from local out to more than 1000 miles, but hops of a few hundred miles are most common. Range depends to some extent on transmitter power, antenna gain and receiver sensitivity, but patience and operating skill are important.

There is much to be learned about auroral propagation. On 50 Mc., for example, an occasional aurora will produce clear voice signals from distances out to 1200 miles or more, not unlike those encountered in sporadic-*E* skip propagation. These may be accompanied by the distorted signals from shorter distances, the degree of distortion decreasing with frequency. On rare occasions, a long-haul east-west skip may be observed, permitting work over distances up to 2000 miles or more, such as between the first and seventh calls areas.

A somewhat similar type of propagation is observed more often by the few v.h.f. operators of the far north. They have found 50-Mc. communication possible occasionally with stations in the northern tier of states and adjacent Canadian areas, apparently by something approximating an ionospheric skip, using the auroral zone as a reflecting medium.[10]

The number and geographical distribution of auroras and auroral propagation on the v.h.f. bands vary with the solar activity cycle, the maximum auroral incidence apparently lagging the sunspot peak by approximately two years. The arctic effects described immediately above were still being observed on 50 Mc. at the bottom of the solar cycle, so their relation to solar conditions is by no means clear.

F_2-Layer DX may be possible in the peak years of the 11-year sunspot cycle. This ionospheric mode, responsible for most DX on lower frequencies, opened the 50-Mc. band for worldwide communication during solar peaks of 1947 to 1950 and 1956 to 1960. Particularly in the late 1950s, the 50-Mc. band was excellent for distances of 2000 miles or more, for many hours at a time, almost daily during the winter months. The first scattered F_2 DX of the current cycle came in 1967.

Frequencies near the maximum usable (*m.u.f.*) produce the strongest F_2-layer signals, and multiple-hop effects and combinations with other forms of propagation may provide 50-Mc. communication over very long paths. Africa to California, U.S.A. to Japan, Hawaii to Australia, and even South America to Japan, were covered frequently in the late 1950s, with signal strengths rivaling the best ever experienced on lower frequencies. Only lack of 50-Mc. privileges for amateurs of many countries prevented 50 Mc. from becoming the prime amateur DX band during this memorable period.

Whether like conditions will prevail during future peaks of solar activity is a matter of some conjecture. Sunspot records dating back to about 1750, Fig. 2-5, show long-term trends indicating that we may be near the end of an era of generally-high activity. Thus it is possible that peaks of the magnitude of recorded cycles 18 and 19 may not recur within the lifetimes of readers of these pages. Cycle 20, peaking as this text was written, shows a lower trend, thus far.

Meanwhile, the m.u.f. can be checked readily with a general-coverage receiver.[11] Since propagation near the m.u.f. is very good, signals will be heard from somewhere on any frequency that is alive. The 10-meter band provides good clues, for its skip distance shortens markedly as the m.u.f. rises toward 50 Mc. If 10 is open for long periods daily, and the skip shortens to 1200 miles or less during the peak hours, the m.u.f. is approaching 50 Mc. This is the time to watch the frequencies just below 50 Mc., making note of the highest frequency at which DX signals can be heard, and the time of day when they appear. A daily record of these observations will show if the m.u.f. is rising. Enough use of the v.h.f. range is made, almost everywhere in the world, so that there will be plenty of evidence of an imminent 50-Mc. opening. Usually there are many signals, both in the band and near it. European television made 50-Mc. DX work difficult for amateurs in Eastern U.S.A. during peak hours of Cycle 19, and hundreds of other European signals and harmonics were audible whenever the band was open.

The National Bureau of Standards publishes information which shows the predicted m.u.f. for any path, at any time of day. Their bulletin, *Ionospheric Predictions,* issued three months in advance, may be obtained individually for 25 cents per copy, or on a subscription basis, \$2.75 yearly (\$3.50 foreign), from the Superintendent of Documents, Washington 20402. Use of the charts is explained in *Handbook 90, Handbook for CRPL Ionospheric Predictions,* price 40 cents, from the same source.

Since the m.u.f. is related to the position of the sun, it is highest at roughly noon at the midpoint of a given path. It tends to be highest in the low altitudes, and lowest along paths traversing the auroral zones. The highest recorded F_2 m.u.f. was in the vicinity of 75 Mc.

Back-Scatter signals from amateur stations inside the skip zone indicate high m.u.f., and also show the direction in which conditions are most favorable for long-distance work. The F_2 layer

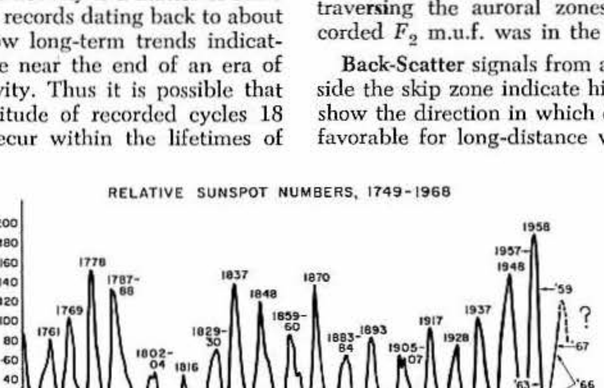


Fig. 2-5—Relative sunspot number records dating back to before 1750 show that the last two solar peaks, known as cycles 18 and 19, were the highest in all of man's observation of the sun. Looking at the long-term curve indicates that we may be near the end of a 40-year era of generally high solar activity.

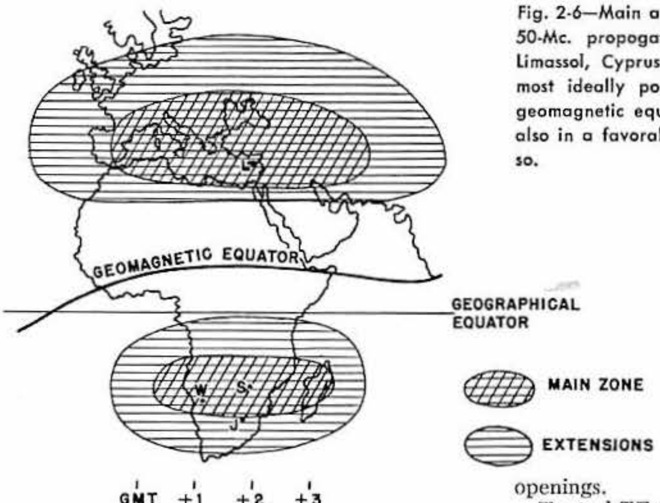


Fig. 2-6—Main and occasional zones of transequatorial 50-Mc. propogation, as described by ZE2JV, show Limassol, Cyprus, and Salisbury, Rhodesia, to be almost ideally positioned with respect to the curving geomagnetic equator. Windhoek, Southwest Africa, is also in a favorable spot. Johannesburg somewhat less so.

has almost mirror qualities near the m.u.f. Signals reflected from it come down at a distant point on the earth's surface, from which they scatter in all directions. Some of the energy comes back to the ionosphere and is reflected back to earth. Thus a station in Virginia, for example, will be heard by a station in Ohio, when both have their antennas aimed at Europe and that path is open for both of them.

Signals scattered back to or near to their point of origin are weak and have a high degree of multipath distortion, somewhat like those reflected from the aurora. Voice may be only partly readable, and c.w. is highly effective under such conditions. Back-scatter is usually strongest for stations no more than a few hundred miles apart, but back-scatter QSOs have been made between points as widely separated as New England and Mexico City. When this occurred, both stations were aiming at a common "open" point in the South Atlantic. Alaska and California, with a common opening to Japan, have had similar experiences. Often the direct path between the two stations will produce no signals, the circuit being open only via the longer back-scatter route.

A similar condition may be observed during periods of sporadic-E propagation (see page 18) though the shorter skip may make E-layer back-scatter hard to distinguish from other modes of propagation.

Transequatorial Propagation during the afternoon and evening hours is possible for 50-Mc. stations situated at optimum distances from the geomagnetic equator, roughly 1500 to 2500 miles above and below. See Fig. 2-6. This mode is associated with high sunspot periods, but because it is effective at frequencies up to at last 1.5 times the observed daytime m.u.f. for the F_2 layer, it runs for a longer portion of the solar cycle than does the normal F-layer propagation described above. It has been observed over paths that cross the equator at angles as low as 45 degrees, and at greater distances than given above, but long paths and large deviations from the north-south route show lower m.u.f. and shorter openings.

Typical TE paths of high reliability are Puerto Rico to Argentina, Japan to Australia, and the Mediterranean area to Southern Rhodesia.[12] These circuits continue to show TE propagation in the 50-Mc. band long after F_2-layer DX has expired. At the peak of Cycle 19 the TE mode was observed over most of the United States, and in Europe as far north as the British Isles on isolated occasions. The spring and fall months show this mode to the best advantage. Signals may have a high degree of flutter, but voice readability is seldom seriously impaired.

Ionospheric Scatter is usable for marginal 50-Mc. communication over distances comparable to those encountered in single-hop E_s, mainly 600 to 1200 miles. Because only a very small part of the energy scattered in the E region of the ionosphere returns to earth, such signals are extremely weak. Large antennas, fairly high transmitter power and good receivers are essential, and even with all these only c.w. emission can be expected to produce consistent results. Though ionospheric scatter is now widely used for military communication over long distances, is adaptation to amateur use is limited.[13] Operation on carefully-kept schedules, with the precise frequencies to be used known in advance, is almost a necessity, but amateur experience on many 50-Mc. scatter circuits has shown that the bare essentials of communication can be exchanged, if good equipment and operating skill are available at both ends of the path.

When there are aiding factors, such as developing sporadic-E ionization, or more than the average number of meteors near the midpoint, ionospheric scatter circuits improve markedly, even though the path may not be considered "open" in the normal sense.

Tropospheric Scatter is similar to the ionospheric form, except that it occurs nearer to the earth's surface, and consequently shows mainly as an extension of the normal working range of v.h.f. stations. It is effective at all frequencies, 50 Mc. and higher, but has been used by amateurs mainly for 144-Mc. work. Experience has shown that with optimum equipment, signals can be exchanged on 144 Mc. consistently at distances out to 450 to 500 miles,[14] regardless of propagation conditions. Such signals are very weak and difficult to copy, and as in all weak-signal work, c.w. is the only mode by which communication can be carried on effectively in tropospheric scatter over the extreme distances.

Meteor Scatter is one of the more esoteric forms of v.h.f. DX currently worked by amateurs. Meteors entering the earth's atmosphere from outer space burn up rapidly in the *E* region of the ionosphere due to friction. In the process a cylinder of dense ionization is formed as a trail behind the meteor. A 50-Mc. signal reflected from this trail may appear as a few words of readable voice, from a station up to 1200 miles away. If you make a habit of tuning the 50-Mc. band carefully for considerable periods, you may have heard these bursts from time to time. On 144 Mc., the same meteor would provide a much shorter burst, perhaps no more than a "ping" heard when the receiver beat oscillator is on.

Meteors are constantly entering the earth's atmosphere, so if a v.h.f. receiver is left tuned to a distant station its signal pings will be heard at random intervals at all times. The number and duration of bursts increase greatly during major showers listed in Table I, and they may come often enough to permit communication of sorts between cooperating stations.[15] Meteor bursts are heard frequently on any 50-Mc. ionospheric-scatter circuit, rising suddenly far above the weak residual signal that is characteristic of such communication. On 144 Mc. there is normally no residual signal beyond a few hundred miles, so only the meteor bursts are heard. Occasionally during major meteor showers, bursts of up to a minute or more of continuous 144-Mc. signal may be received.

Such fortunate breaks are by no means necessary for communication between dedicated enthusiasts. Using high keying speeds and precisely-timed transmissions on accurately-known frequencies, they often achieve exchanges of information on a series of bursts of no more than a few seconds' duration each.

The usual arrangement is for one station to transmit for exactly one minute, 30 seconds, or even 15 seconds, following which the other takes over for a like period. Detailed procedure is agreed on in advance. Typically, a signal report (S1, S2 or S3, indicating duration of bursts, rather than signal strength) is sent immediately upon identification of the other station. When the signal report is received, an *R* is transmitted repeatedly, signifying receipt.

Daily schedules kept through a major shower peak will usually yield enough bursts on at least one day to complete a contact within an hour or less of such cooperative effort. A long loud burst is usually the signal to abandon timed transmissions abruptly and try for a complete exchange before signals disappear. Very short sequences, such as 15 seconds each way, have the advantage of making it likely that any appreciable burst will be usable in this way.

Information in the meteor shower table should be used as a guide, and not relied on completely. Showers may not peak at exactly the same time from year to year, and those having widely-spaced periodic peaks may be deflected and not appear at all.

After putting on the biggest show in amateur radio history up to that time, October, 1946,[16] the Giacobinids were deflected, and were hardly discernible 13 years later. On the other hand, the surprizing resurgence of the November Leonids in 1965 and 1966 provided 144-Mc. meteor DX of unprecedented proportions.[17] The August Perseids and December Geminids are dependable year after year, and are highly favored for v.h.f. scheduling on that account.

Minor showers listed in the table may offer little more opportunity for communication than do the random meteors always entering the earth's atmosphere. Scheduled work over the right distances, especially between 2200 and 0900 local time, should yield some results practically every day.

Voice contacts are made occasionally during meteor showers, and success in this has increased as voice-controlled s.s.b. has become more widely used.

Lunar Communication has been the dream of v.h.f. men for a generation, but few have succeeded in using the e.m.e. path to date. Pioneering 2-meter men received their own signals reflected from the moon in 1953,[18] and long distances have been covered two-way in recent years.[19] Marginal communication via the moon has been carried out over long distances on the 420-and 1215-Mc. bands,[20] but only with very sophisticated equipment. Communication by reflection from the moon remains a challenge to the more advanced amateur, and it is a fine project for clubs having the facilities, ability and determination to achieve a measure of success.

The requirements are fairly well established. They include the maximum legal power, the ultimate in receiver performance, very large antennas capable of being aimed and controlled accurately,[21] and a willingess to work with very weak c.w. signals. Because very high receiver selectivity is often used, to realize the best possible signal-to-noise ratio, stability requirements in both transmitter and receiver are critical.

Antenna problems are similarly staggering. At 144 Mc., an antenna gain well in excess of 20 db. is required. Polarization rotation along the path, out and back, makes it appear that circular polarization offers the best chance of success. Because of the reflection of the signal, the polarization sense should be reversed between transmission and reception, in listening for one's own echoes.

The polarization rotation problem is largely eliminated above 1000 Mc., but the difficulties

Table I—Meteor Shower Data for V.H.F. Use

Shower and Date	Time Visible Rise	Time Visible Set	Optimum Paths and Times N-S	Optimum Paths and Times NW-SE	Optimum Paths and Times E-W	Optimum Paths and Times SW-NE	Hourly Rate Visual	Hourly Rate Radio	Velocity, km/sec.	Period, Years	Next Maximum
* January 3–5 Quadrantids	2300	1800	—	0300–0800 SW	0800–0900 S	0900–1400 SE	35	45	45	7	Note 1
January 17 Cygnids	0230	2130	—	0600–1100 SW	1100–1300 S	1300–1800 SE	—	—	—	—	—
February 5–10 Aurigids	1200	0330	—	1400–1730 SW	—	2130–0100 SE	—	—	—	—	—
March 10–12 Boötids	2200	0830	2330–0030 W 0530–0630 E	0330–0530 NE	0230–0330 N	0030–0230 NW	—	—	—	—	—
March 20 Coma Berenices	1800	0630	2130–2300 W 0100–0300 E	2000–2130 SW	—	0300–0430 SE	—	—	—	—	—
* April 19–23 Lyrids	2100	1100	0230 W 0530 E	2330–0100 SW	—	0700–0830 SE	8	12	51	415	Note 1
* May 1–6 Aquarids	0300	1200	—	0830–1000 NE	0630–0830 N	0500–0630 NW	12	12	66	76	Note 1
May 11–24 Herculids	1800	0630	2130–2300 W 0100–0300 E	2000–2130 SW	—	0300–0430 SE	—	—	—	—	—
May 30 Pegasids	2300	1200	0300–0430 W 0630–0800 E	0130–0300 SW	—	0800–0930 SE	—	—	—	—	—
June 2–17 Scorpiids	2000	0300	—	0100 NE	2300–2400 N	2200 NW	—	—	—	—	—
June 27–30 Pons Winnecke	Does not set; min. at 0900		—	1500–1830 SW	1830–2330 S	2330–0300 SE	—	—	—	—	—
July 14 Cygnids	1800	1000	—	2100–2330 SW	0130 S	0330–0600 SE	—	—	—	—	—
July 18–30 Capricornids	2030	0400	—	0100–0200 NE	2300–0100 N	2200–2300 NW	—	—	—	—	—
* July 26–31 Aquarids	2200	0600	—	0300–0500 NE	0100–0300 N	0000–0100 NW	10	22	50	3.6	Note 1
* July 27–August 14 Perseids	Does not set; min. at 1730		—	2330–0300 SW	0300–0800 S	0800–1130 SE	50	50	61	120	Note 1
August 10–20 Cygnids	1200	0700	—	1700–1930 SW	2130 S	2330–0200 SE	—	—	—	—	—
August 21–23 Draconids	Does not set; min. at 0900		—	1500–1830 SW	1830–2330 S	2330–0300 SE	—	—	—	—	—
August 21–31 Draconids	Does not set; min. at 0700		—	1300–1630 SW	1630–2130 S	2130–0100 SE	—	—	—	—	—
September 7–15 Perseids	2130	1200	—	0030–0200 SW	—	0700–0830 SE	—	—	—	—	—
September 22 Aurigids	2100	1230	—	0030–0200 SW	—	0700–0830 SE	—	—	—	—	—
October 2 Quadrantids	0500	0000	—	0900–1400 SW	1400–1500 S	1500–2000 SE	—	—	—	—	—
October 9 Giacobinids	0600	0300	—	1100–1600 SW	1600–1700 S	1700–2200 SE	Note 2		20	6.6	1972
October 12–23 Arietids	1900	0700	2130–2330 W 0230–0430 E	—	—	—	—	—	—	—	—
* October 18–23 Orionids	2230	0930	0000–0200 W 0600–0800 E	0430–0600 NE	0330–0430 N	0200–0330 NW	15	30	68	76	Note 1
* Oct. 26–Nov. 16 Taurids	1900	0630	2100–2300 W 0300–0500 E	0130–0300 NE	0030–0130 N	2300–0030 NW	10	16	27	3.3	Note 1
* November 14–18 Leonids	0000	1230	0300–0500 W 0800–1000 E	—	—	—	12	Note 3	72	33.2	1999
November 22–30 Andromedids	1300	0600	—	1600–2000 SW	—	2300–0300 SE	Note 4		22	6.7	1971
* December 10–14 Geminids	1900	0900	0030 W 0330 E	2130–2300 SW	—	0500–0630 SE	60	70	35	1.6	Note 1
* December 22 Ursids	Does not set; min. at 2030		—	—	0130–1530 S	—	13 Note 5	13	38	13.5	1971, 1972
* May 19–21 Cetids	0530	1430	—	1100–1230 NE	0900–1100 N	0730–0900 NW	—	—	20	37	—
* June 4–6 Perseids	0500	1730	0800–1000 W 1300–1500 E	—	—	—	—	—	40	29	—
* June 8 Arietids	0330	1530	0600–0800 W 1100–1300 E	—	—	—	Note 6		70	38	—
* June 30–July 2 Taurids	0500	1700	0700–0900 W 1300–1500 E	1130–1300 NE	1030–1130 N	0900–1030 NW	—	—	30	31	—

* Major showers—Last four are daylight showers.

NOTES

1. These streams are evenly distributed and little year to year variation is to be expected.
2. Very concentrated stream. Peak years give up to 400 meteors per minute, but with duration of only 6 hours. 1946 peak was most concentrated shower in amateur radio experience up to that time (see December, 1946, *QST*, page 43) but 1959 recurrence was deflected and was hardly observable.
3. Peak years give 60/hour visual. In the peak years of the 1800s, prior to being deflected by Jupiter and Saturn, this shower gave 1200 per minute. Spectacular results in 1965 and 1966 are reported in Jan. 1966 QST, page 80, and Jan. 1967, page 83.
4. Before being deflected by Jupiter this stream gave peak year rates of 100/minute. No notable rates have been observed since, though the stream could return.
5. Short duration shower. Peak years the radio rate is 165/hour.
6. This intense daylight shower begins June 2 and runs to June 14 with radio rates from 25 to 70/hour.

of generating sufficient power, and building a good enough receiver are multiplied. Detailed discussion of moonbounce problems is beyond the scope of this book. Study of the references at the end of this chapter is recommended to the serious would-be worker in this field.

Progress in amateur communication via the moon has been steady, though largely unspectacular, in recent years. Because the ultimate in equipment of every kind is required, it is probably unrealistic to expect that any great "breakthrough" in this field will suddenly make reliable communication by moon *reflection* practical, within the framework of the amateur regulations. The prospect of a lunar repeater is, however, quite another matter. Man's progress in the mastery of space travel being what it is, repeaters on the moon could revolutionize our whole approach to amateur radio communication within a decade.

Passive and Active Satellites offer possibilities for worldwide communication in the v.h.f. range and higher frequencies. The former consists of a reflector of some sort, as in the Echo series, presenting problems even more severe than those involved in lunar communication. Very good equipment is required, and antennas designed for tracking must be used, if this mode of communication is to be used by amateurs.

Active satellites pick up the transmitted signal and relay it, usually on another frequency. While such a device is relatively easy for the amateur to use, the design and operational problems are formidable. The number of signals it can accommodate at any one time is severely limited. The active satellite is expected to play a large role in amateur radio of the future, as are other communications ramifications of the space age. Problems and possibilities of active satellites have been discussed extensively in *QST*.[22]

RELIABLE V.H.F. COVERAGE

In preceding pages we discussed means by which our bands above 50 Mc. may be used intermittently for communication far beyond the visual horizon. In emphasizing DX we should not neglect a prime asset of the v.h.f. bands: reliable communication over relatively short distances. The v.h.f. region is far less subject to disruption of local communication than are frequencies below about 30 Mc. Since much amateur communication is essentially local in nature, our v.h.f. assignments could carry a much greater load than they presently do, and this would help solve interference problems on lower frequencies.

Possibly some amateur unwillingness to migrate to the v.h.f. bands is due to misconceptions about the coverage obtainable. This reflects the age-old idea that v.h.f. waves travel only in straight lines, except when the DX modes enumerated above happen to be present. Let us

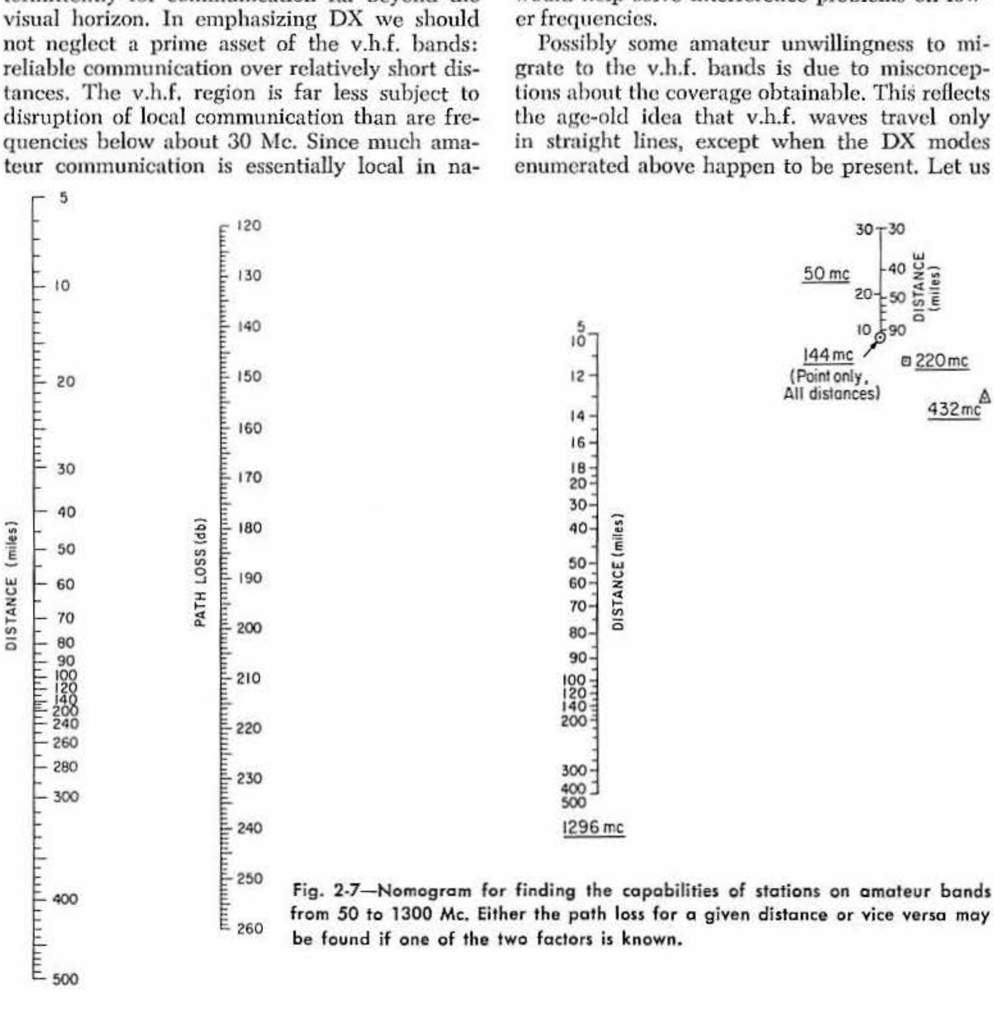


Fig. 2-7—Nomogram for finding the capabilities of stations on amateur bands from 50 to 1300 Mc. Either the path loss for a given distance or vice versa may be found if one of the two factors is known.

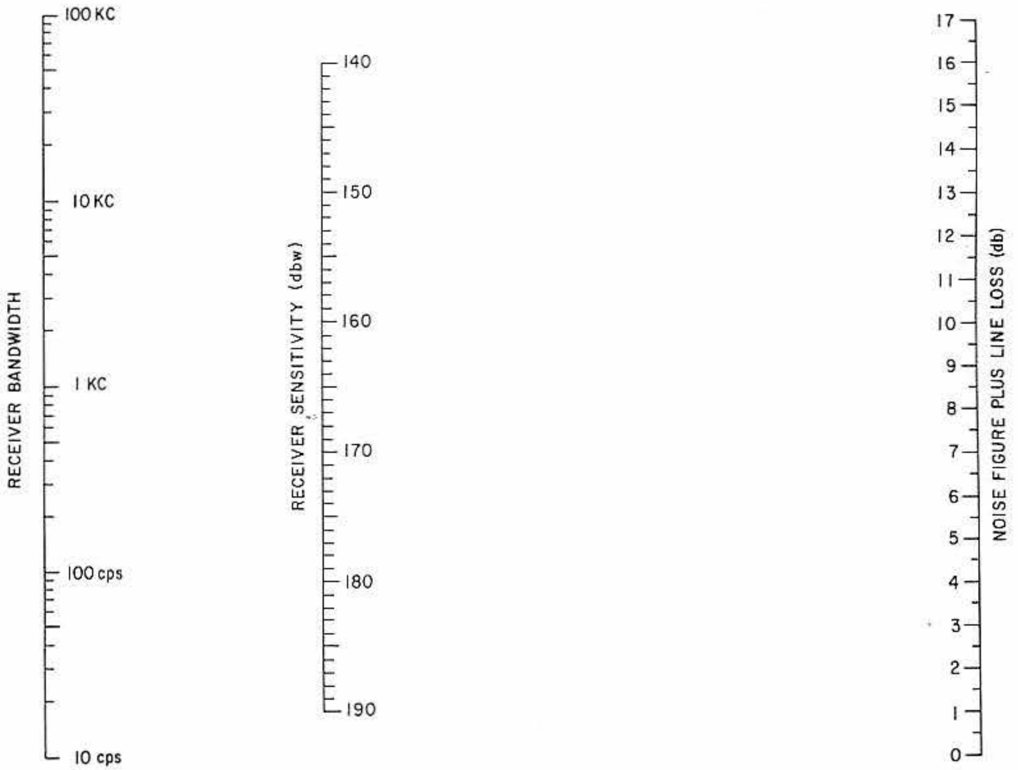


Fig. 2-8—Nomogram for finding effective receiver sensitivity.

survey the picture in the light of modern wave-propagation knowledge, and see what the bands above 50 Mc. are good for on a day-to-day basis, ignoring the anomalies that may result in extensions of normal coverage.

It is possible to predict with fair accuracy how far you should be able to work consistently on any v.h.f. or u.h.f. band, provided a few simple facts are known. The factors affecting operating range can be reduced to graph form, as described by D. W. Bray, K2LMG.[23] To estimate your station's capabilities, two basic numbers must be determined: *station gain,* and *path loss.* Station gain is made up of eight factors: receiver sensitivity, transmitted power, receiving antenna gain, receiving antenna height gain, transmitting antenna gain, transmitting antenna height gain, and required signal-to-noise ratio. This looks complicated but it really boils down to an easily-made evaluation of receiver, transmitter and antenna performance. The other number, path loss, is readily determined from the nomogram, Fig. 2-7. This gives path loss over smooth earth, for 99 per cent reliability.

For 50 Mc., lay a straightedge from the distance between stations (left side) to the appropriate distance at the right side. For 1296 Mc., use the full scale, right center. For 144, 220 and 432, use the dot in the circle, square or triangle, respectively. Example: at 300 miles the path loss for 144 Mc. is 215 db.

Station Gain

The largest of the eight factors involved in station gain is receiver sensitivity. This is obtainable from Fig. 2-8, if you know the approximate receiver noise figure and transmission-line loss. If you can't measure noise figure, assume 3 db. for 50 Mc., 5 for 144 or 220, 8 for 432, and 10 for 1296, if you know that your equipment is working moderately well. Line loss can be taken from Table 8-III for the line in use, if the antenna system is fed properly. Lay a straightedge between the appropriate points on either side of Fig. 2-8, to find effective receiver sensitivity in db. below one watt (dbw.). Use the narrowest bandwidth that is practical for the emission intended, with the receiver you will be using. For c.w., an average value for effective work is about 500 cycles. Phone bandwidth can be taken from the receiver's instruction manual.

Antenna gain is next in importance. Gains of amateur antennas are often exaggerated. For well-designed Yagis they run close to 10 times the boom length in wavelengths. (Example: a 24-foot Yagi on 144 Mc. is 3.6 wavelengths long. $3.6 \times 10 = 36$, or about 15½ db.). Add 3 db. for stacking, where used properly. Add 4 db. more for ground-reflection gain. This varies in amateur work, but averages out near this figure. We have one more plus factor: antenna height gain, obtainable from Fig. 2-9. Note that this is

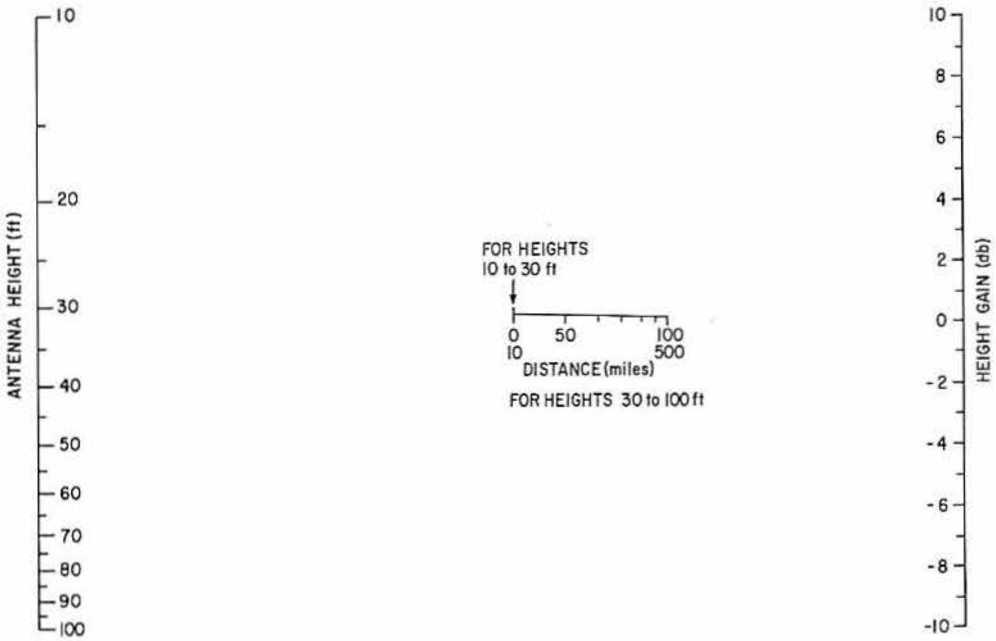


Fig. 2-9—Nomogram for determination of antenna-height gain.

greatest for short distances. The left edge of the horizontal center scale is for 0 to 10 miles, the right edge for 100 to 500 miles. Height gain for 10 to 30 feet is assumed to be zero. It will be seen that for 50 feet the height gain is 4 db. at 10 miles, 3 db. at 50 miles, and 2 db. at 100 miles. At 80 feet the height gains are roughly 8, 6 and 4 db. for these distances. Beyond 100 miles the height gain is nearly uniform for a given height, regardless of distance.

Transmitter power output must be stated in db. above 1 watt. If you have 500 watts output, add 500/1, or 27 db., to your station gain. The transmission line loss must be subtracted from the station gain. So must the required signal-to-noise ratio. The information is based on c.w. work, so the additional signal needed for other modes must be subtracted. Use 3 db. for s.s.b. and 7 db. for a.m. Loss due to fading must be accounted for. It has been shown that for distances beyond 100 miles the signal will vary plus or minus about 7 db. from the average level, so 7 db. must be subtracted from the station gain for high reliability. For distances under 100 miles, fading diminishes almost linearly with distance. For 50 miles, use minus 3.5 db. for fading.

What It All Means

After adding all the plus-and-minus factors to get the station gain, use it to find the distance over which you can expect to work reliably, from the nomogram, Fig. 2-7. Or work it the other way around: find the path loss for the distance you want to cover from the nomogram and then figure out what station changes will be needed to overcome it.

The significance of all this becomes more obvious when we see path loss plotted against frequency for the various bands, as in Fig. 2-10. At the left this is done for 50 per cent reliability. At the right is the same information for 99 per cent reliability. For near-perfect reliability, a path loss of 195 db. (easily countered at 50 or 144 Mc.) is involved in 100-mile communication. But look at the 50 per cent reliability curve: the same path loss takes us out to well over 250 miles. Few amateurs demand near-perfect reliability. By choosing our times, and accepting the necessity for some repeats or occasional loss of signal, we can maintain communication out to distances far beyond those usually worked by v.h.f. men.

Working out a few typical amateur v.h.f. station setups with these curves will show why an understanding of these factors is important to any user of the v.h.f. spectrum. Note that path loss rises very steeply in the first 100 miles or so. This is no news to v.h.f. men; locals are very strong, but stations 50 or 75 miles away are much weaker. But what happens beyond 100 miles is not so well known to some of us.

From the curves of Fig. 2-10, we see that path loss levels off markedly at what is the approximate limit of working range for average v.h.f. stations using voice. Work out the station gain for a 50-watt station with an average receiver and moderate-sized antenna, and you'll find

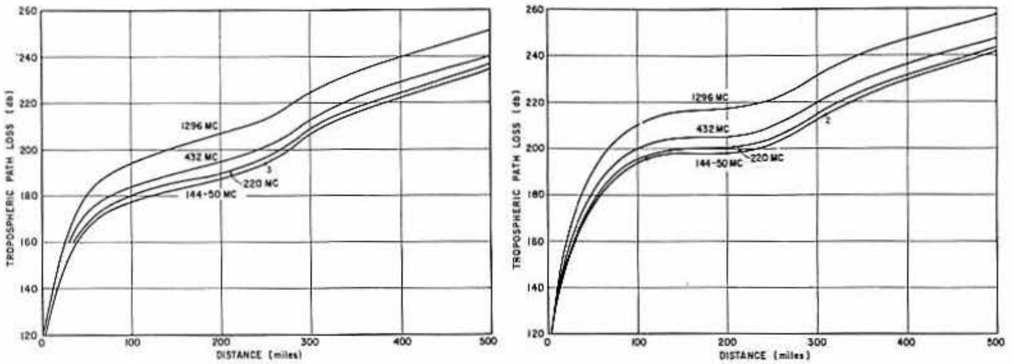


Fig. 2-10—Path loss vs. distance for amateur frequencies above 50 Mc. Curves at the left are for 50 per cent of the time; those at the right for 99 per cent. The former is the more representative of amateur radio requirements.

that it will come out around 180 db. This means about a 100-mile working radius in average terrain, for good but not perfect reliability. Another 10 db. may extend the range to as much as 250 miles. Just changing from a.m. phone to c.w. can thus do wonders for you. A bigger antenna, a higher one if your present beam is not at least 50 feet up, an increase in power to 500 watts from 50, an improvement in receiver noise figure if it is presently poor—any of these things can make a big improvement in reliable coverage. Achieve all of them, and you will have very likely tripled your sphere of influence, thanks to that hump in the path-loss curves. This goes a long way toward explaining why using a 10-watt packaged station and a small antenna, fun though it may be, does not begin to show what the v.h.f. bands are *really* good for.

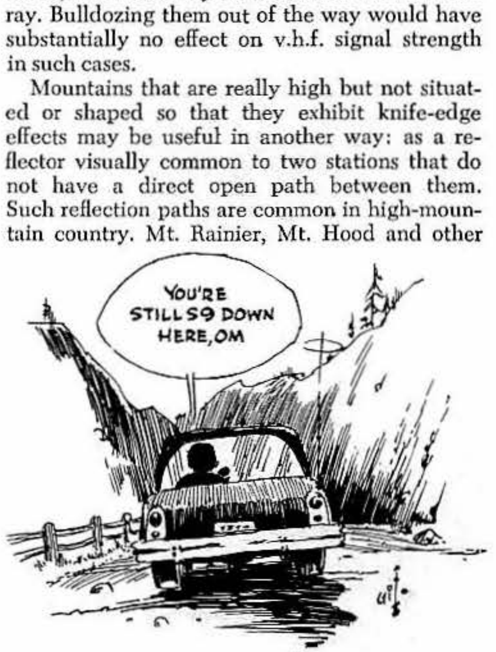
About Terrain

The coverage figures derived from the above are for average terrain. What of stations in mountainous country? Though an open horizon is generally desirable for the v.h.f. man, mountain country should not be given up as hopeless until it has been proven so. Help for the valley dweller often lies in the optical phenomenon known as *knife-edge refraction.*[24] A flashlight beam pointed at the edge of a partition does not cut off sharply at the partition edge, but is refracted around it, partially illuminating the shadow area. A similar effect is observed with v.h.f. waves passing over ridges; there is a shadow effect, but not a complete blackout. If the signal is strong where it strikes the mountain range, it will be heard well in the bottom of a valley on the far side.

This is familiar to all users of v.h.f. communications equipment who operate in hilly terrain. Where only one ridge lies in the way, signals on the far side may be almost as good as on the near. Under ideal conditions (a very high and sharp-edged obstruction near the midpoint of a path long enough so that signals would be weak over average terrain), knife-edge refraction may yield signals even stronger than would be possible with an open path.

The obstruction must project into the radiation patterns of the antennas used. Often mountains that look formidable to the viewer are not high enough to have an appreciable effect, one way or the other. Since the normal radiation from a v.h.f. array is several degrees above the horizontal, mountains that are less than three degrees above the horizon, as seen from the antenna, are missed by the radiation from the array. Bulldozing them out of the way would have substantially no effect on v.h.f. signal strength in such cases.

Mountains that are really high but not situated or shaped so that they exhibit knife-edge effects may be useful in another way: as a reflector visually common to two stations that do not have a direct open path between them. Such reflection paths are common in high-mountain country. Mt. Rainier, Mt. Hood and other



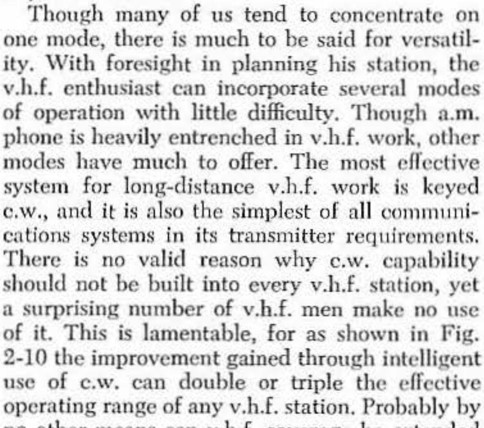
majestic peaks of the Northwest are examples, and the mountains of California provide many others. Mt. McKinley in Alaska has demonstrated remarkable capabilities for both knife-edge and reflector service.

Rolling terrain, where obstructions are not sharp enough to produce knife-edge refraction, still does not exhibit complete shadow effect. There is no complete barrier to v.h.f. propagation; only attenuation, which varies widely as the result of many factors. Thus, even valley locations are usable for v.h.f. communication. Good antenna systems, preferably as high as possible, the best available equipment, and above all, the willingness and ability to work with weak signals may make possible outstanding v.h.f. work, even in sites that show little promise by casual inspection.

OPERATING MODES

Almost every amateur has one mode of operation that he prefers over all others. Once this was a simple choice between phone and code, but today's picture is more complex. The voice operator on the v.h.f. bands can use amplitude modulation, suppressed-carrier single-sideband or double-sideband, or frequency modulation, either wide or narrow-band. The code man can employ conventional c.w., keyed tone modulation or frequency-shift keying. Other modes include television (slow-scan or wide-band), radio teletype (with either audio- or radio-frequency shift), facsimile and pulse. Because some of these are wide-band modes, taking up more space than would be permissible in crowded lower bands, the choice open to the inhabitant of the world above 50 Mc. is wider than for any other amateur.

Though many of us tend to concentrate on one mode, there is much to be said for versatility. With foresight in planning his station, the v.h.f. enthusiast can incorporate several modes of operation with little difficulty. Though a.m. phone is heavily entrenched in v.h.f. work, other modes have much to offer. The most effective system for long-distance v.h.f. work is keyed c.w., and it is also the simplest of all communications systems in its transmitter requirements. There is no valid reason why c.w. capability should not be built into every v.h.f. station, yet a surprising number of v.h.f. men make no use of it. This is lamentable, for as shown in Fig. 2-10 the improvement gained through intelligent use of c.w. can double or triple the effective operating range of any v.h.f. station. Probably by no other means can v.h.f. coverage be extended so easily.



An appreciable improvement in consistent range with voice can be achieved by going to single-sideband, as compared with a.m. phone. Both c.w. and sideband are more effective than other modes mainly because they occupy a narrower band of frequencies. C.w. requires essentially no space at all, and thus it permits almost infinite receiver selectivity when suitable techniques are employed. Sideband requires less than half the spectrum space of other voice modes. No power is wasted in transmitting a carrier, this being supplied, in effect, by the receiver's beat oscillator. The b.f.o. is also an important factor in the effectiveness of c.w.

Frequency modulation, all but ignored by a generation of v.h.f. men, has great potential worth. With suitable receiving techniques, f.m. provides almost totally noise-free communication within its service area. Either wide-band or narrow-band f.m. is easily incorporated in a v.h.f. transmitter, and because it adds nothing to the power that must be dissipated by the tubes, or to the voltage that components must be capable of withstanding, the full c.w. ratings of all parts of the transmitter apply to f.m. as well. Perhaps most important of all, f.m. eliminates practically all chance of audio-type interference in TV receivers, broadcast sets, hearing aids and other audio amplifiers, and so is invaluable in solving interference problems for the amateur in densely-populated areas. Not widely appreciated is the fact that, with proper receiving techniques, f.m. is at least the equivalent of a.m. in reliable v.h.f. coverage. Only c.w. and s.s.b. are superior.

Tone modulation (A2) is a simple means of sending code on the v.h.f. bands with any voice transmitter. It is only slightly better than a.m. phone for weak-signal work, but it is fine for code practice, and for use on frequencies where either the receiver or transmitter may not be stable enough for c.w., or where the receiver does not have a b.f.o.

Use of frequency-shift keying (f.s.k.) is confined mainly to radio-teletype work, but it could be employed for c.w. With today's voltage-variable capacitors, it is a simple matter to shift the frequency for f.s.k..

Slow-scan television has no special v.h.f. connotation. Conventional high-quality TV requires a very wide band of frequencies, so it is confined to bands from 420 Mc. up. Nearly all amateur TV is currently on the 420-Mc. band, between 436 and 450 Mc.

Pulse is a very wide mode inherently and is permitted only on certain bands above 2300 Mc. Though employed only to a limited extent in amateur communication, it has interesting possibilities on frequencies where its wide-band nature can be accommodated.[25]

Another emission is usable on all amateur frequencies above 51 Mc.: unmodulated carrier, AØ. It has many uses, not the least being remote control systems for model aircraft, boats and the like. Much neglected by v.h.f. men, an interesting adaptation of AØ is duplex phone. In using AØ the operator need identify only once every ten minutes or less, and he is not *required* to have any intelligence on the signal in between identifications. This permits running the transmitter while tuning for other signals, either on the same band as the transmitter, or on another band. The 50- and 144-Mc. bands tend to become congested at their low ends, so it is well to use duplex technique only in the less-occupied upper portions. The U.S. Regulations allow AØ only above 51 Mc. for this reason, and it is well to check the frequencies to be used for AØ to see that they are unoccupied, before embarking on duplex work.

Use of low power, separate antennas for transmission and reception, and earphones in place of speakers are aids to effective duplex operation. Especially where signals are reasonably strong, as in local work, duplex provides a ready give-and-take exchange that can be far more pleasant and efficient than the one-way monologue that conventional methods entail.

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Reception Above 50 Mc.

V.H.F. RECEIVER CHARACTERISTICS

Nowhere in amateur radio is topnotch reception so important as on the frequencies above 50 Mc. Selectivity, stability, sensitivity, smooth tuning, ability to reject unwanted signals—all considered necessary in h.f. receivers—are even more desirable in v.h.f. work. While receiver problems are similar, to some extent, on all frequencies used by amateurs, ideal solutions vary greatly for various bands and modes of operation. Thus, whether you buy your receiver or build it, an understanding of basic receiver principles and consideration of your primary operating interests are necessary, if you would make an intelligent choice. We will have a look at each of the above qualities, as it relates to v.h.f. reception.

Selectivity

On lower frequencies selectivity is desirable mainly to prevent interference between stations in our crowded bands. It serves this end for v.h.f. men, too, but it is also important in achieving the best possible signal-to-noise ratio. Receiver noise output, for a given input signal level, is related to bandwidth; the wider the bandwidth the greater the noise. This is not important in reception of strong signals, but for best readability of weak signals no more bandwidth should be used than is necessary to pass the intelligence on a signal. This kind of selectivity is mainly determined by the i.f. system in a superheterodyne receiver, and the means for achieving it do not change much with signal frequency.

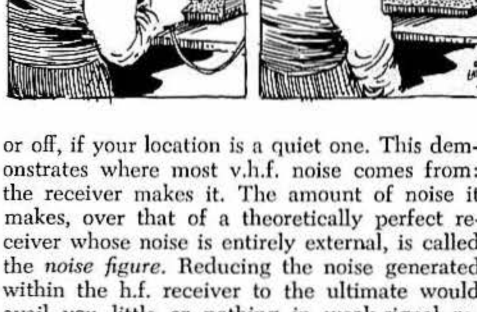
Stability

As selectivity is increased the need for good stability rises, and as we go higher in frequency the difficulty of achieving satisfactory stability also rises. Thus it is almost standard procedure to use crystal-controlled frequency converters in v.h.f. reception, doing the tuning at some lower or *intermediate frequency* where stability is more readily attained. Like selectivity, stability is then mainly a matter of the design of the lower-frequency components of the v.h.f. receiving system.

Sensitivity and Noise Figure

Here we reach the parting of the ways with our h.f. brethren. Reception of weak v.h.f. signals is limited by factors quite different from those affecting bands below 30 Mc. On frequencies up to the lower portion of the v.h.f. spectrum, reception is limited almost entirely by noise picked up by the antenna. This may be man-made (ignition, power-line noise, electrical noise from motors, neon signs and the like) or natural, such as galactic or solar noise and atmospherics. With modern tubes, transistors and circuits the noise contribution by the receiver itself is inconsequential. To prove this for yourself, tune your receiver to some spot below 30 Mc. where no signals are heard. Turn off the a.v.c. and advance the gain controls until noise is heard. Now remove the antenna. If your receiver is stable and well shielded it should go dead quiet, or nearly so. The noise was practically all coming in on the antenna.

Now try the test with a v.h.f. receiver, on 144 Mc. or higher frequencies. Chances are that the noise will change hardly at all, antenna on



or off, if your location is a quiet one. This demonstrates where most v.h.f. noise comes from: the receiver makes it. The amount of noise it makes, over that of a theoretically perfect receiver whose noise is entirely external, is called the *noise figure*. Reducing the noise generated within the h.f. receiver to the ultimate would avail you little or nothing in weak-signal reception on bands up through 30 Mc., but you can improve reception in the v.h.f. range markedly if you can amplify incoming signals faster than you build up the noise generated internally. Hence the emphasis on low-noise amplifiers in v.h.f. receiver design.

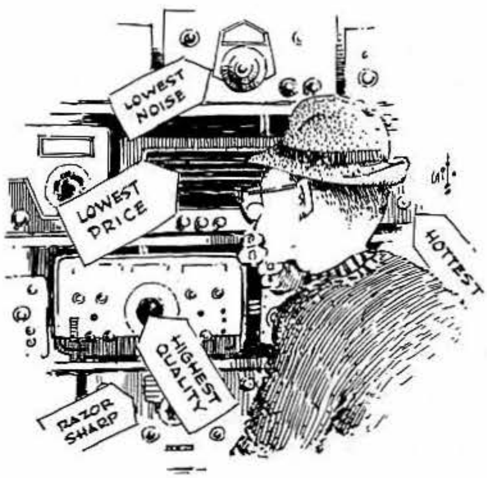
The noise output of a receiver, by itself, is of no importance at any frequency. How much a signal stands out above the receiver noise (called *signal-to-noise ratio*) is important on any frequency, and it is compounded of many factors, including selectivity as well as noise figure. Noise figure, on the other hand, is almost entirely a matter of the design and adjustment of the first stages of a v.h.f. receiving system, and it is *independent of bandwidth*. We'll look into this complex business of low-noise reception in more detail later on, but it is important to keep the above facts in mind.

Mechanical Considerations

The best selectivity, stability and sensitivity are of little use if you cannot tune the receiver effectively. Nothing is more disconcerting than a receiver that tunes too rapidly, or in sloppy fashion. Backlash in tuning mechanisms is very annoying, and the higher the selectivity the more troublesome it becomes. These are mechanical problems, but don't under-estimate their importance in building or choosing a receiver for v.h.f. work. Few receivers are entirely satisfactory in these respects, and many low-priced ones are all but useless. The would-be v.h.f. enthusiast will do well to check the mechanical qualities of a receiving system with great care.

Rejection of Unwanted Signals

A v.h.f. receiver could score high in all the above categories and still be unsatisfactory if it responds to signals of other services near our bands, or overloads readily when near neighbors come on the air in or near the band we're trying to use. No receiver is completely free of spurious responses, so we may have a difficult choice here. In areas of high population density, where there may be v.h.f. men in every

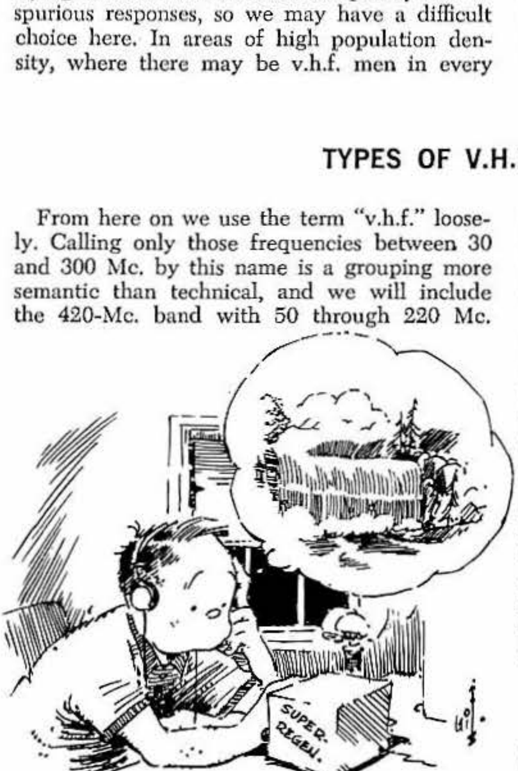


block, and TV, police, aircraft, f.m. and other v.h.f. services on almost every available kilocycle, ability of a receiver to reject unwanted signals may have to take precedence over other desirable characteristics, particularly low noise figure. We may have to give the receiver help in the form of a filter of some sort. See Chapter 12.

It should be obvious from what has been said thus far that there is no one "perfect receiver." Even with unlimited resources and design skill at our disposal, we still must examine our own particular set of operating circumstances and objectives, and select equipment or techniques that offer the best over-all hope for success.

TYPES OF V.H.F. RECEIVERS

From here on we use the term "v.h.f." loosely. Calling only those frequencies between 30 and 300 Mc. by this name is a grouping more semantic than technical, and we will include the 420-Mc. band with 50 through 220 Mc.



more often than not. In the light of present techniques, the logical dividing line between v.h.f. and u.h.f. methods lies somewhat above our 420-Mc. band, rather than at 300 Mc.

Reception on 50 through 450 Mc. can be accomplished in many ways, but v.h.f. receivers are of two principal types: the superregenerator and the superheterodyne. The first may be very simple—one tube or transistor and little else other than an antenna, a tuned circuit and headphones. Common additions are one or more audio stages to operate a speaker, and an r.f. amplifier stage, to improve performance and reduce detector radiation. The superheterodyne may be complete in itself, usually with four or more tubes or transistors, or it may be a combination of a v.h.f. converter and a communications receiver intended for use on lower frequencies.

The Superregenerative Detector

Today's newcomer may not be too familiar with this wonderful device (and probably it's just as well!) but it was almost standard equipment in early v.h.f. work. To give the Devil his

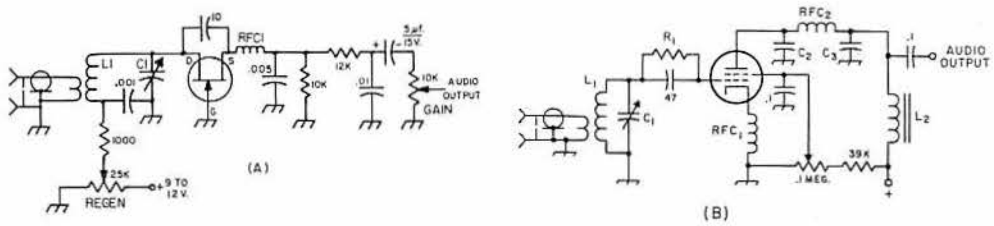


Fig. 3-1—Circuits of typical superregenerative detectors using an FET transistor, A, and a tetrode tube, B. Regeneration is controlled by varying the drain voltage on the detector in the transistor circuit, and the screen voltage in the tetrode or pentode. Values of L_1 and C_1 should be adjusted for the frequency involved, as should the size of the r.f. choke, RFC_1.

C_2, C_3—0.001-μf. disk ceramic. Try different values up to 0.005 for desired audio quality.
R_1—2 to 10 megohms.
L_1—To suit frequency.
L_2—Small audio or filter choke; not critical.
RFC_1—Single-layer r.f. choke, to suit frequency.
RFC_2—85-mh. r.f. choke.

due, the "rushbox" was a potent factor in popularization of the v.h.f. bands—and for good reason. Nothing of comparable simplicity has been found to equal its ability to detect weak signals, but like all simple devices the superregen has serious limitations. It provides little selectivity, has a high and rather unpleasant background noise level, and it radiates a broad interfering signal around its receiving frequency. While various refinements may minimize these objectionable features, the superregenerative receiver is used today mainly where its small size, light weight and low power drain are all-important, as in short-range portable work.

Most tubes and transistors that do well in other v.h.f. receiving applications make good superregenerative detectors. With tubes, tetrode or pentode types are favored at 50 or 144 Mc., as variation of their screen voltage offers smooth control of regeneration. Triodes are better for higher frequencies. Typical circuits are shown in Fig. 3-1. If no r.f. amplifier is used, operating conditions and the coupling to the antenna should be adjusted so as to permit superregeneration with the lowest satisfactory power input, to hold down detector radiation. An r.f. amplifier stage ahead of the detector will reduce or eliminate radiation, and by isolating the detector from the antenna will make control of regeneration less critical. It will also add some gain and selectivity.

Superheterodyne Receivers

Because amplification is more efficient at low frequencies than high, it is standard v.h.f. practice to use only as much r.f. amplification as may be needed for good noise figure, and then convert the signal to a lower or intermediate frequency, to be amplified and detected. This is the basic superheterodyne principle, used in nearly all radio reception today.

The simplest superhet receiver for 50 Mc. is shown in Fig. 3-2A. The antenna feeds a mixer stage operating at the signal frequency, in this case 50 to 54 Mc. A tunable oscillator, usually at some lower frequency (though it can be higher) supplies energy to beat with the signal and produce an intermediate frequency (i.f.) which is then amplified and detected. In our example the oscillator is 5 Mc. below the signal frequency, and the i.f. is, of course, 5 Mc. You could build the simple receiver with as few as two dual-purpose tubes, but its gain would be low and its selectivity poor. You would not be happy with it for long.

Gain can be increased with more i.f. amplifier stages, but a simple 5-Mc. amplifier, as in 2A, is not sufficiently selective, so we go to what is known as double conversion, Fig. 3-2B. Here a second oscillator and mixer convert the 5-Mc. signal to 455 kc., where gain and selectivity come much easier. Also added here is an r.f. amplifier stage, for improved noise figure and better sensitivity.

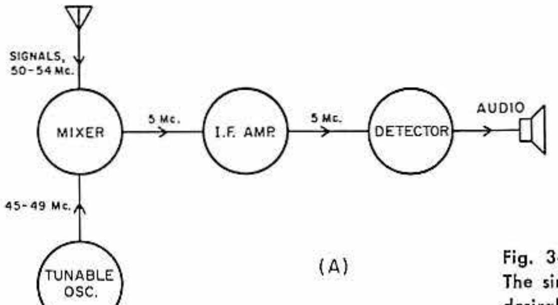
You could go right to 455 kc. in the first conversion, by suitable choice of frequency range for the tunable oscillator, but this gives rise to a serious image problem. Suppose we want to listen to someone on 51 Mc. With an i.f. of 455 kc., the oscillator would then be on 50.545 Mc., 455 kc. away from the signal frequency. The mixer responds to signals 455 kc. on *either* side of the oscillator frequency, so someone on 50.090 Mc. would interfere with the desired signal on 51 Mc. All superhets have this problem, but when a high i.f. is used (roughly 10 percent of the signal frequency is common) the selectivity of the first tuned circuits of the receiver is sufficient to reject the unwanted image signal. Thus, for good image rejection, most v.h.f. receivers employ double- or triple-conversion circuits.

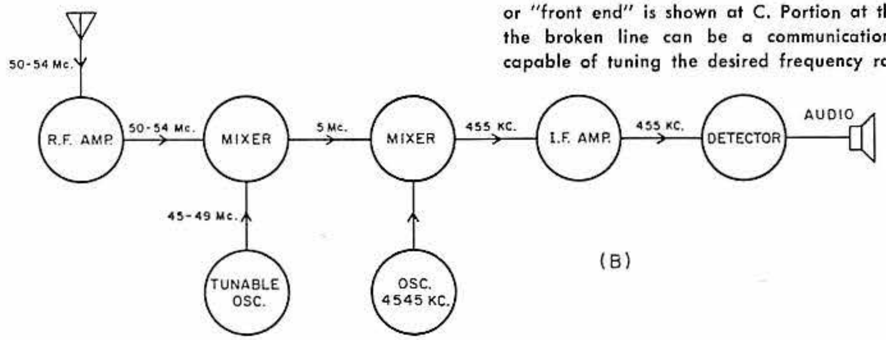
So far we have used a tunable oscillator and fixed intermediate-frequency amplifiers. In the v.h.f. range, however, tunable oscillators may not be stable enough for narrow-band reception, so we more often use a crystal-controlled source for the first conversion, as in Fig. 3-2C. Here the first intermediate frequency is variable, so the i.f., detector and audio system (portion to the right of the vertical broken line) may take the form of a communications receiver that tunes the desired frequency range, in this example 14 to 18 Mc. Our r.f. amplifier, first mixer and crystal oscillator are usually built in a single unit called a crystal-controlled converter. This converter-receiver combination is the most common approach to amateur v.h.f. reception in use today.

There are uses for the single-conversion receiver however, especially where simplicity, low cost and small size are more important than high selectivity. Also, by use of advanced i.f. design techniques, particularly involving the crystal-lattice filter, it is possible to develop excellent selectivity at high intermediate frequencies. Though fairly expensive, the high-frequency crystal filter has much to recommend it. Requiring only a single conversion for good selectivity, it reduces the possibility of unwanted signals being heard, and it makes possible optimum performance with fewer stages and circuits than multiple-conversion systems.

A promising over-all receiving system for the v.h.f. man is the use of such a single-conversion setup for the 50-Mc. band as the basic receiving unit of the station. This would have a tunable oscillator; converters for 144 Mc. and higher bands would be crystal-controlled, with 50 to 54 Mc. as the tunable first intermediate frequency.

It should be emphasized that the frequencies given in the above discussion are examples only. Almost any combination of oscillator and intermediate frequencies can be used and many factors govern the choice. These will be taken up later.





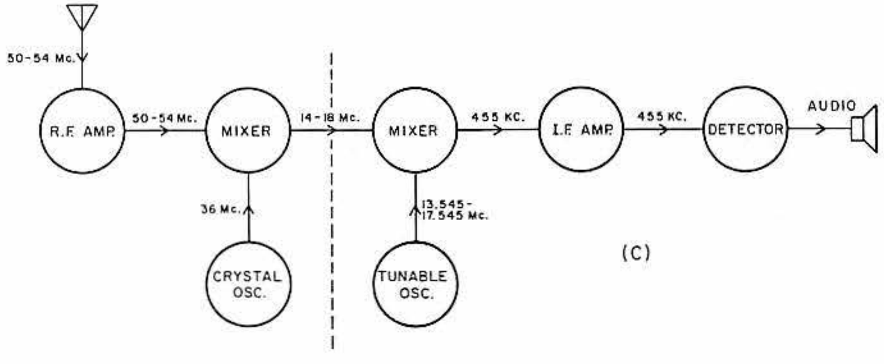


Fig. 3-2—Development of the v.h.f. superheterodyne. The simplest receiver of this type, A, would lack most desirable qualities. B shows a double-conversion system, with r.f. and a second conversion added. A double conversion system with crystal-controlled converter, or "front end" is shown at C. Portion at the right of the broken line can be a communications receiver capable of tuning the desired frequency range.

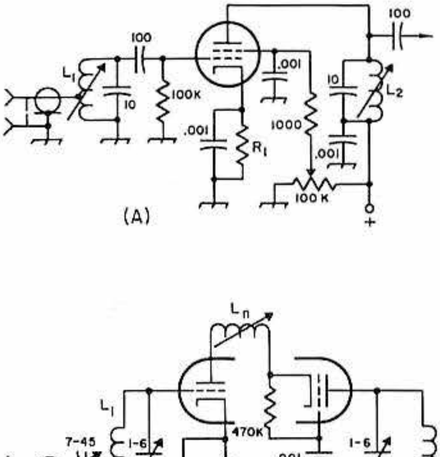
V.H.F. CONVERTER DESIGN

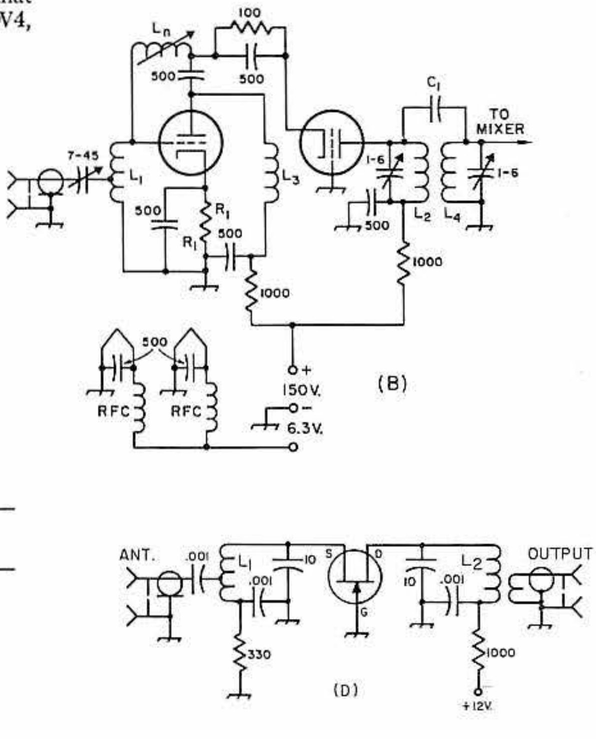
The relation between the noise generated in the first r.f. amplifier stage and the gain of that stage is the principal factor in the weak-signal sensitivity of v.h.f. receivers. Low noise figure is mainly a matter of r.f. amplifier tube or transistor design, though circuitry and adjusment are factors. Modern tubes and transistors, and the circuits to go with them, have made possible a marked improvement in v.h.f. and u.h.f. reception in recent years.

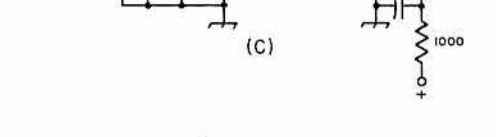
Low-Noise R.f. Amplifier Tubes and Circuits

Tubes developed for r.f. amplifier service above 30 Mc. are of compact design, to hold down lead inductance, tube capacitance, and electron transit time. They are mechanically and electrically constructed to have high transconductance. The better triodes are rated at 10,000 micromhos or more, and some costly types reach 50,000. Inexpensive triodes that work well in v.h.f. amplifiers include the 6CW4, 6DS4, 6AM4, 6BC4 and several others no longer being made, but still available in some quarters. Several dual triodes work well, including the 6BQ7, 6BZ7, 6BC8, 6BS8 and others.

At 50 Mc., where internal noise is not yet a dominant factor, pentode r.f. amplifiers may give more gain per stage, with somewhat better freedom from overloading, than triodes. The 6AK5 and similar tubes work well, and more expensive high-transconductance types such as the 7788 are very good at 50 Mc. and usable at 144 Mc. Noise figure at this and higher frequencies will not equal the better triodes, however. A typical pentode r.f. amplifier circuit is given at A in Fig. 3-3. An advantage of the pentode is ease of gain control, through variation of screen voltage. This may be an important consideration in preventing mixer overloading, in areas of high v.h.f. population density.







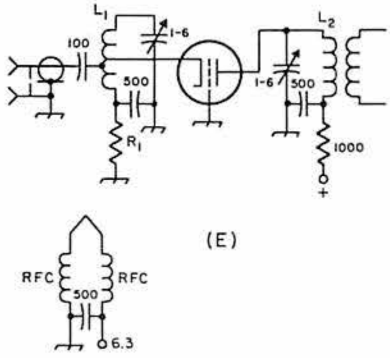


Fig. 3-3—Typical r.f. amplifier circuits for v.h.f. converter or preamplifier service. Decimal values of capacitance are in uf.; others in p.f. Good quality button-mica or ceramic feed-through capacitors should be used for by-passing at 144 Mc. and higher. Disk ceramics are usable at 50 Mc. Values of inductance in coils and r.f. chokes depend on frequency. Cathode resistor values should be chosen to suit the tube type used.

Circuit A is a simple pentode or tetrode amplifier, useful mainly below 100 Mc. B and C are versions of the cascode circuit. A grounded-gate FET amplifier is shown at D, and its tube counterpart at E.

Triode r.f. amplifiers require some form of neutralization. This can be capacitive, as commonly used in transmitter circuits, or inductive, as shown in Fig. 3-3B and C. An alternative to neutralization is isolation of the input and output circuits through use of a grounded-gate or grounded-grid amplifier, Fig. 3-3D and E.

Two examples of the commonly used cascode r.f. amplifier circuit are shown. B and C are basically the same, in that each has a neutralized triode working into a grounded-grid triode. The grounded-grid stage, having low input impedance, places a heavy load on the first amplifier, and may prevent it from oscillating, even without neutralization. Better noise figure is possible with neutralization, however. This is accomplished with the coil L_n, which is resonated broadly at the midpoint of the intended tuning range.

The simplified cascode, Fig. 3-3C, is often used with dual-triode tubes. Here the triodes are operated in series as far as the plate supply is concerned, and a few less components are required than with the circuit at B. With any except the highest-transconductance tubes there is little choice between these two circuits. With the best tubes, particularly when reliable means for measuring performance are available, circuit B is preferable.

Amateur workers unfamiliar with the cascode circuit occasionally encounter instability trouble that is not solved by neutralization of the first stage. In such instances the trouble may lie in the grounded-grid portion, rather than in the neutralized stage. The grid (or gate in the FET transistor form) must be at true ground potential for r.f. Where the element is connected directly to ground, as in circuits B, D and E, the connection should be as short as possible, to the chassis or to a grounded shield plate mounted across the socket. In the by-passed version, 3-3C, the capacitor should be the button-mica or ceramic feed-through type. Disk ceramics are seldom satisfactory above 50 Mc., unless the value and lead lengths are selected to make the capacitor series-resonant at the operating frequency. (See Chapter 13.)

The grounded-grid or grounded-gate amplifier, Fig. 3-3D and E, has the virtue of simplicity, yet it is capable of good noise figure, and adjustment may be simpler than with neutralized circuits. Stage gain is likely to be lower, however, and two stages may be required for best performance. The input element here is the tube's cathode or the FET's source. The input impedance is low, so better results may be obtained if the element is tapped down the input coil, L_1, as in E. The antenna lead is also tapped down on the input circuit, and the positions of the taps should be adjusted for optimum performance, though they will not be found to be particularly critical. True grounding of the grid or gate element is important, as with the cascode circuit discussed above.

To achieve the best possible noise figure in grounded-grid amplifiers at 144 Mc. and higher, it may be important to keep the cathode *above* ground for r.f. Heater chokes (RFC in circuits B and E) may be helpful. Ferrite beads slipped over the heater lead right at the socket work very well in place of the heater chokes shown.

Adjusting R. F. Stages

The first step in adjusting an r.f. amplifier is to set the tuned circuits at the approximate frequency. This can be done with the aid of a dip meter, or by peaking the circuits for maximum response while actually receiving. In circuit A, the only additional step is to adjust the tap position on L_1 for best noise figure or signal-to-noise ratio. Noise figure adjustments are best made with a noise generator, as detailed in Chapter 11. The work can be done on a weak steady signal, if care is used to get the *greatest margin of signal over noise*, rather than merely maximum gain.

In adjusting antenna coupling, best noise figure will be found as the coupling is increased a bit *beyond* the point where maximum signal is obtained. The gain and signal level may drop slightly, but the noise falls off faster at this point, so signal-to-noise ratio actually improves at the start of over-coupling. Gain can be made up anywhere in the receiver, but noise figure is set by the first stage, and predominantly in the first tuned circuit. This cardinal principle of v.h.f. receiver design should be borne in mind at all times.

Procedure is similar for neutralized stages, except for the work with L_n. In circuit B the plate or heater voltage can be removed from the first tube, and L_n adjusted for *minimum* response on a strong signal. The other circuits should be peaked for maximum meanwhile. There will be interlocking between the various circuits, so repeat this operation several times. The first tube may then be put back into normal operation, and final adjustment for lowest noise figure made by adjusting the input coupling and the inductance of L_n as outlined above. If separate tubes are used in Circuit C the same procedure may be followed, by removing heater voltage from the first tube. With dual triodes having a single heater the adjustment of neutralization must be by the cut-and-try method.

The grounded-grid stage, E, can be used with untuned input, requiring only tuning of the plate circuit for best results. The tuned input circuit in amplifier E is adjusted for best noise figure, rather than maximum gain.

Two output coupling arrangements for working into a following mixer are shown. Capacitive coupling, as in A, is simple, but with only the single tuned circuit, L_2, rejection of unwanted frequencies may be poor. Better band-pass characteristics are obtainable with the double-tuned circuit in amplifier B. Its coupling capacitor, C_1, can be varied to provide different degrees of coupling and selectivity. Usually values of 1 to 3 pf. are used, and these are readily obtained by twisting insulated leads of hookup wire together for a half inch or so.

The low-impedance output coupling shown in circuit D can be used wherever a stage is to serve as a preamplifier ahead of an existing v.h.f. converter or receiver. A simplified version of circuit E, with the antenna connected directly to the cathode through a capacitor, and the tuned circuit omitted, may be quite helpful in boosting the 50-Mc. performance of some multiband communications receivers that cover this band. Often these receivers are only slightly deficient in gain and noise figure at 50 Mc., and a simple preamplifier may be all that is needed.

V.h.f. Mixers

Because the noise figure of a receiver having a good r.f. amplifier is set by its first stage, the mixer in a v.h.f. converter can be designed with other objectives in mind. These may include freedom from overloading, and desired bandpass characteristics. Where the oscillator is tunable, reaction on the oscillator frequency by mixer circuits may be a factor to be considered.

The mixer can be a pentode, a triode, a transistor, or even a crystal diode. Pentodes generally provide higher conversion gain and require less oscillator injection than triodes. Properly designed, they offer good overload characteristics. Their upper limit of frequency for good operation is somewhat lower than for triodes. The triode mixer is simple, and it works well at any frequency in the v.h.f. range. Transistor mixers may use either FET or bipolar transistors, but the latter are more susceptible to overloading.

Crystal-diode mixers are common at 420 Mc. and are standard practice for all higher amateur bands. When suitable diodes and circuits are used, the noise figure obtainable with a crystal mixer in u.h.f. work is better than all but the very best vacuum-tube circuits. The crystal mixer with no preceding r.f. amplifier is an effective u.h.f. receiving device, provided that the first i.f. stage following it is of low-noise design. Low-noise design information in the section on r.f. amplifiers applies equally to i.f. stages following a diode mixer.

A v.h.f. converter using tubes or transistors nearly always has an r.f. amplifier stage ahead of the mixer, in the interest of low noise figure, but where extreme simplicity is the primary concern the amplifier may be omitted. The noise figure of the mixer then becomes important, requiring different operating conditions from those employed when an amplifier is used.

In the pentode mixer circuit, Fig. 3-4A, the optimum value of screen resistor depends on whether or not an amplifier precedes the mixer. Best noise figure is obtained if the plate current is held to the lowest usable value, so the screen resistor, R_1, will be around 1 megohm in a simple converter with no r.f. stage. Low plate current makes for easy overloading, however, so the tube is allowed to draw more current in a converter having a good r.f. amplifier ahead of the mixer. Usual values are 10,000 to 47,000 ohms, depending on plate supply voltage. Any of several miniature pentodes (6AK5, 6CB6 and others) will work well as mixers, up through the 220-Mc. band.

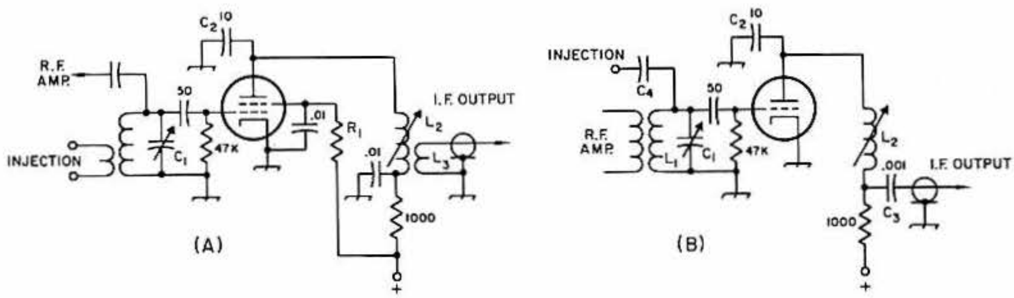
A typical triode mixer circuit is shown at B in Fig. 3-4. Any v.h.f. triode is suitable for use to at least 300 Mc. Some of the better types (6AM4, 6CW4, 6J6 etc.) will work well up to about 500 Mc., though crystal mixers tend to take over in amateur service above 420 Mc.

Various arrangements are shown for input and output coupling, and for feeding injection from the oscillator to the mixer. These are given as examples, rather than to show a circuit that is desirable for the type of tube in question. Capacitive coupling from the r.f. stage, as in A, is simple; but inductive coupling, as in B, can be used for desirable band-pass effect and better rejection of unwanted frequencies. There is little choice between the output coupling methods given in A and B, though A may be slightly better in cases where leak-through of signals at the intermediate frequency is troublesome. Injection coupling can be inductive, as in A, or capacitive, as in B. Capacitor C_4 can be made variable by using the ends of insulated hookup wire, twisted together for a turn or two. Either coupling method should be adjusted to the lowest value that will give satisfactory conversion gain. Excessive injection may increase the likelihood of overloading and spurious responses.

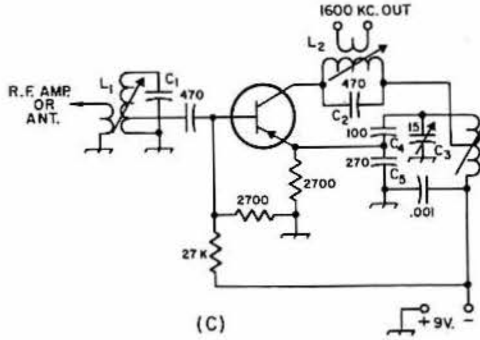
The mixer plate circuit L_2C_2 should have low Q to prevent humped response at one portion of the desired tuning range. Capacitor C_2 should be a low value, and preferably connected directly between the plate and cathode terminals of the mixer tube socket. Connected in this way, C_2 helps to prevent parasitic resonance near the signal frequency in the mixer plate circuit, which could result in oscillation. This trouble is more common in triode mixers than with pentodes, but the above is good oscillation insurance with any tube mixer.

Dual tubes are often used in mixer-oscillator service. Where there is a common heater and cathode, as in the 6J6 and 6X8, injection coupling through the tube elements may be sufficient. In others the coupling may be as shown in A or B. In the case of triode-pentodes it is customary to use the pentode as the mixer and the triode as the oscillator.

A very simple mixer-oscillator using a single transistor is given in Fig. 3-4C. As with circuits A and B, L_1C_1 resonates at the signal frequency. The oscillator tuned circuit is L_3C_3. In a 50-Mc. mixer this combination tunes in the 24-Mc. region, its second harmonic beating with incoming signals in the 50-Mc. band to provide an i.f. of 1600 kc. The output circuit, L_2C_2, may be coupled inductively to the "loopstick" antenna of a small pocket broadcast receiver, or it may be link-coupled to a receiver having a low-



Figs. 3-4—Mixer circuits for v.h.f. converters. A pentode mixer, A, may be adjusted for good noise figure or maximum resistance to overloading by varying the value of R_1. Pentode mixers are used mainly for the lower v.h.f. region. The triode mixer, B, works well on any frequency up to about 500 Mc. A very simple transistor mixer-oscillator circuit for use with broadcast receivers is shown at C. For other transistor mixers, see Chapter 4.



impedance input circuit. Capacitor C_2 "looks like" a bypass to the 24-Mc. energy in L_3. This mixer-oscillator is subject to spurious responses of many kinds, but it is useful in simple v.h.f. converters where ideal performance is not important. Further details and an application will be found in Edition 1 of this Manual, in the 44th and 45th Editions of the ARRL Handbook, and November, 1964, *QST*, page 26.

Where a crystal mixer is used without a preceeding low-noise r.f. amplifier, two points are important if low-noise performance is to be achieved. First, injection to the mixer should be as pure as possible. Care should be taken to eliminate unwanted products of the oscillator-multiplier stages, as injection of other than the proper heterodyning frequency will add to the mixer noise and tend to produce responses to other than the desired signal frequencies. Second, coupling between the injection stages and the mixer should be loose. Otherwise, much of the signal may be lost through the injection circuits.

Oscillators for V.h.f. Converters

Most v.h.f. converters used today are the crystal-controlled variety, but the tunable-oscillator converter still has its uses. It is good where the receiver or i.f. system with which the converter is used does not have adequate tuning facilities. A tunable converter is less subject to interference from signals at its intermediate frequency, because this frequency can be selected for freedom from i.f. interference and then left set. If the mixer and r.f. stages are gang-tuned with the oscillator they can be made as selective as good engineering permits, thus effecting a considerable reduction in interference potential from out-of-band signals.

The tunable oscillator presents stability problems, however. The components must be mechanically and electrically selected with this in mind. The tuning capacitor must be solidly built, and preferably the double-ended type. Split-stator variable capacitors designed for v.h.f. service, having rugged mounting brackets and possibly ball-bearing end plates, are well worth their extra cost. Leads should be stiff wire. Coils should preferably be wound on ceramic forms. If air-wound they should be supported rigidly in some manner, and not suspended on their leads. The oscillator should run at very low input, and its plate voltage should not be removed during transmitting periods.

Tunable oscillators for v.h.f. converter service are shown in Fig. 3-5. Circuit A may be used with any v.h.f. triode, the 6AF4 and 6CW4 being favored. The transistor circuit, B, is useful mainly

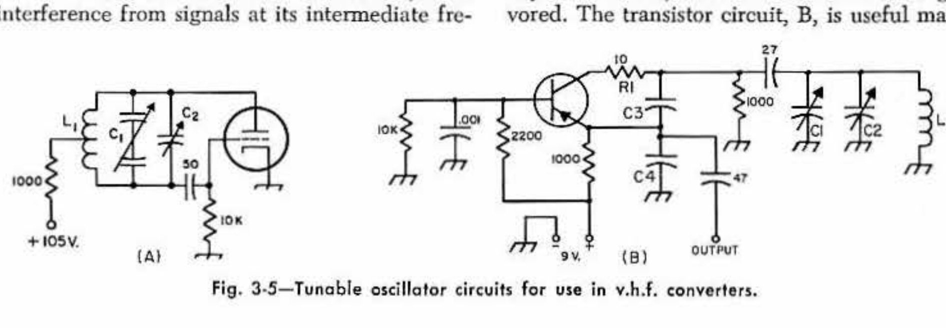


Fig. 3-5—Tunable oscillator circuits for use in v.h.f. converters.

for 50-Mc. converters. Any v.h.f. transistor should do. In both circuits the values for L_1, C_1, C_2 depend on frequency. In B, C_3 and C_4 should be as high as can be used and still sustain oscillation; 180 and 220 pf. are typical. R_1 maybe omitted if there is no problem with parasitic oscillation. The padder capacitor, C_2, is used to set the tuning range to the proper frequency, and to spread out the tuning on C_1. It can be temperature compensated to correct drift due to heating. Output can be taken off through a link around L_1, or by capacitive coupling to any portion of the tuned circuit.

In a crystal-controlled converter the oscillator can make use of any of the circuits shown in Chapter 5. Usually it will be an overtone oscillator, Fig. 5-4, and the same considerations apply as for transmitting. Most converter designers mount their oscillator and multiplier stages in the same assembly as the r.f. portions of the converter, but there are good arguments for building the injection stages separately. Particularly where multiplier stages are needed, as in most converters for 144 Mc. and higher, a separate injection unit makes complete isolation more feasible, preventing unwanted frequencies from reaching the mixer. This may be helpful in eliminating interference from TV, f.m. and other services operating in the v.h.f. region.

Multiplier stages in a converter should operate at as low a power level as possible. A convenient way to obtain high orders of frequency multiplication is the crystal-diode multiplier, shown in several of the converters described in Chapter 4.

COMMUNICATIONS RECEIVER PROBLEMS

Most of the information in this chapter has to do with crystal-controlled converters, since this is the most common approach to v.h.f. reception. The converter is only part of the problem, however. Unless the h.f. receiver with which the converter is to be used is satisfactory in the qualities discussed in the first paragraphs of this chapter, the best v.h.f. converter design will be largely wasted. Let us consider the communications receiver, therefore, from the point of view of the v.h.f. converter user.

Selectivity: If a receiver is satisfactory for the h.f. amateur bands it will be selective enough for v.h.f. service. Most communications receivers will satisfy all but the most critical users in this respect.

Stability: Tuning with converters imposes no special stability problems. Some receivers, particularly older or inexpensive ones, may have progressively poorer stability on each higher-frequency band, so this may be a factor in choosing the converter output frequency range that will be most desirable with a given receiver.

Sensitivity: The ability of a v.h.f. receiving setup to detect weak signals is determined almost entirely by the first stage of the v.h.f. converter. Any communications receiver worthy of the name will have more gain and sensitivity than you'll ever need in using a converter ahead of it.

Mechanical Qualities: These probably rate first in choosing a receiver for v.h.f. converter use. In general-coverage receivers the tuning rate of the dial system is very important to the v.h.f. man, whereas it may be only a minor consideration to other users. Such receivers ordinarily cover from the broadcast band through 30 Mc., usually in four to six ranges. On inexpensive receivers having four bandswitch positions, each range covers such a wide frequency spread that the number of kilocycles per rotation of the tuning knob is almost certain to be excessive. Receivers having five bands may be better, but usually the tuning rate will be too high on frequencies above about 10 Mc. or so. This may be a factor in selecting the most desirable converter output frequency range. Some of the better general-coverage receivers have six or more bandswitch positions. Only these give adequate tuning smoothness for the v.h.f. man, ordinarily.

There are ways around this problem. One is to mount a vernier mechanism in place of the knob on the general-coverage dial. Usually a mounting arrangement can be worked out that will not disfigure the receiver in any way. Two examples are shown in Figs. 3-6 and 7. The first utilizes a 5-to-1 vernier drive, the other a two-speed drive.

An alternative approach with two-dial receivers is to do the tuning with the bandspread dial, resetting the general-coverage dial as each additional swing of the bandspread mechanism is made. This gives good tuning rate, but after the first swing the calibration is lost. Because the two are connected in parallel, changing the setting of the general-coverage capacitor changes the number of kilocycles a given number of

Fig. 3-6—Typical mounting for a vernier dial to be used on a general-coverage receiver. Mechanical arrangements can be worked out to fit most receivers without the necessity for drilling holes or otherwise permanently disfiguring the receiver. Adjacent control shafts and nuts provide convenient anchorages.

Fig. 3-7—Example of use of a two-speed planetary mechanism to slow down the tuning rate on the general-coverage dial of an inexpensive communications receiver.

revolutions of the other capacitor will cover.

Rejection of Unwanted Signals

Image rejection is the principal concern here. Single-conversion receivers with 455-kc. i.f. systems have relatively poor image rejection, so there is a tendency for strong signals to be repeated 910 kc. away from the desired response. This trouble is worse at the high end of the receiver's coverage than at the low. Up through about 10 Mc. most receivers are selective enough in their r.f. circuits so that the image is not bothersome, so a converter output frequency range beginning at 7 Mc., for example, may be preferable to one starting at 14 Mc.

Most receivers made prior to the late 1940s were single-conversion models. Except for the image problem, many such older receivers are quite satisfactory for v.h.f. converter service, so this is an important factor in selecting the converter output frequency. An old but good receiver may be a better value for the money than an inexpensive newer model, since the v.h.f. man is primarily concerned with receiver qualities that have not changed greatly in many years of receiver development. Some receivers dating back even to before World War II still serve the v.h.f. man's needs quite well, if the converter output frequency is kept low.

Some receivers may not be completely shielded, with the result that stations operating in the converter output frequency range may be heard along with the desired v.h.f. signals. This is a converter design problem as well, but corrective measures may have to be applied at the receiver. The three-terminal antenna connection plate used on many communications receivers is shown in Fig. 3-8. Usually Terminal 1 is connected to the chassis inside the receiver. This connection may act as a coupling loop for i.f. signals. Removing the internal connection and grounding terminal 1 on the outside of the chassis may correct this.

A better solution is to install a coaxial fitting on the rear wall of the chassis. When this is done, the lead normally connected to Terminal 2 inside the receiver should be permanently grounded, and the lead that went to Terminal 3 should be connected to the inner conductor of the coaxial fitting.

Receivers may pick up signals through the a.c. line to some extent. Bypassing the a.c. lead inside the chassis should take care of this source of i.f. leak-through. Some receivers, such as early models of the HRO, have separate power supplies. Power cable leads may pick up signals, and also radiate harmonics of the receiver oscillator that may show up in the tuning range of the v.h.f. converter. Filtering and bypassing power supply leads is required in these instances, much as would be done in TVI-prevention work on a transmitter. See Chapter 12 for examples.

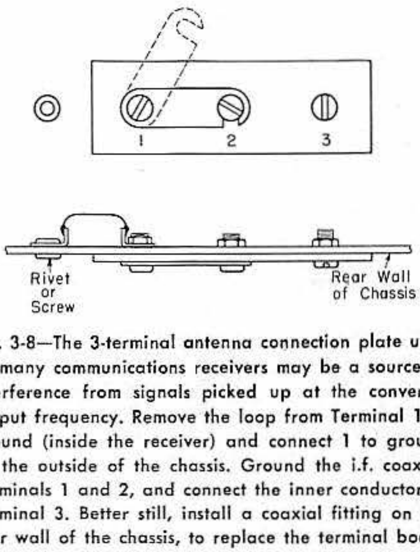


Fig. 3-8—The 3-terminal antenna connection plate used on many communications receivers may be a source of interference from signals picked up at the converter output frequency. Remove the loop from Terminal 1 to ground (inside the receiver) and connect 1 to ground on the outside of the chassis. Ground the i.f. coax to Terminals 1 and 2, and connect the inner conductor to Terminal 3. Better still, install a coaxial fitting on the rear wall of the chassis, to replace the terminal board entirely.

To tell whether signals at the intermediate frequency are coming through because of receiver deficiencies or directly through the v.h.f. converter, try running the receiver alone with its antenna terminals shorted. If no signals are heard in this condition it can be assumed that the trouble lies in the converter. The problem may be one of improper bonding between the receiver and converter chassis. The outer conductor of the coax interconnecting the units may not be enough. Place the converter close to the receiver, and then bond the two together with a heavy braid or strap of copper. Often this will eliminate the leak-through trouble. Effective bypassing of the converter power circuits is another step if the bonding fails. Use 0.01-μf. disk capacitors for this purpose. Lower values may not provide complete bypassing at frequencies under about 10 Mc.

General-Coverage vs. Amateur-Band Receivers

Except for a few models that have a special tuning range for v.h.f. converters, receivers designed for amateur-band service exclusively do not meet the v.h.f. man's tuning needs fully. The obvious solution to this problem is to use more than one crystal in the v.h.f. converter. Two crystals will give the required 4-megacycle spread with receivers that tune 28 to 30 Mc., for example. This may not be entirely satisfactory, however, as the performance of some receivers is poor in the 28-Mc. range compared to lower frequencies.

V.h.f. men who want the best reception with a minimum of crystal changing may decide to forfeit coverage of some parts of the v.h.f. bands in favor of those that can be covered readily on ranges where their receivers work best. This might mean using a receiver's 400-kc. coverage at 14 Mc. with two converter crystals, for example, to give 144.0 to 144.4 Mc. and 145.0 to 145.4 Mc., or any other segments that may happen to fit a local activity picture.

Some communications receivers themselves use crystal-controlled oscillators with a tunable i.f. system. Crystals are usually supplied to cover the amateur bands from 3.5 to 30 Mc. with such receivers. In some instances (as for example the Collins S-Line) crystals can be obtained to extend the continuous coverage. Continuous coverage from 14 to 15.8 Mc. is possible with a 75S-1 or S-3, for example, by substituting suitable crystals in the 21- and 28-Mc. positions, for those supplied with the receiver.

RECEIVING F.M.

Effective use of frequency modulation demands the right receiving techniques. The r.f. end of the v.h.f. receiving setup is the same for f.m. as for other modes; it is in the i.f. and detector circuits that the f.m. receiver differs. The fact that our receivers are almost invariably designed with c.w., a.m. and s.s.b. requirements in mind, but with no provision for good reception of f.m., is probably the greatest deterrent to general acceptance of f.m. in amateur communication. In many respects our neglect of f.m. is unfortunate, for it could contribute much to the v.h.f. scene. It has been said, with some wisdom, that if f.m. had been invented first and had become well established before other voice modes came along, it almost certainly would be the standard method of voice communication all through the v.h.f. and u.h.f. spectrum today.

The f.m. receiver problem is twofold. A type of detector circuit unsuited to other modes must be employed if the full potential of f.m. is to be realized, and the bandwidth of the receiver must match that of the transmitter more precisely than with other modes. As used today, f.m. may be wide-band or narrow-band in nature. The former is not intelligible when tuned in on a receiver of the usual communications bandwidth. We will discuss the special nature of the wide-band f.m. receiver later.

Narrow-Band F.M. Detection

Frequency-modulated signals of the narrow-band variety can be received after a fashion on the conventional communications receiver having an a.m. detector, by what is called *slope detection*. In Fig. 3-9A we see a signal tuned in part-way down one side of the selectivity curve. Either side will do on most receivers, but the carrier strength is always less than the maximum possible with normal tuning "on the nose." When the frequency of the signal varies with modulation, it swings up and down the curve slope, resulting in an amplitude variation in accordance with the modulation on the signal. If this swing is kept within the portion of the selectivity curve that is linear, this results in an f.m.-to-a.m. conversion that permits the signal to be handled by the a.m. detector.

It can be seen that slope detection is inherently a makeshift method of reception, never very efficient. If the receiver selectivity curve is

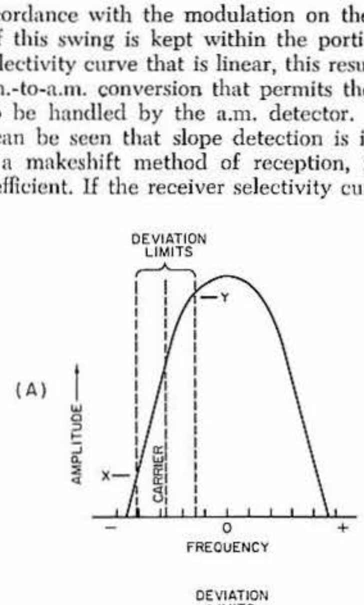


Fig. 3-9—F.m. or p.m. detection characteristics. A—"Slope detection," using the sloping side of the receiver's selectivity curve to convert f.m. or p.m. to a.m. for subsequent rectification. B—Typical discriminator characteristic. The straight portion of this curve between the two peaks is the useful region. The peaks should always lie outside the pass band of the receiver's selectivity curve.

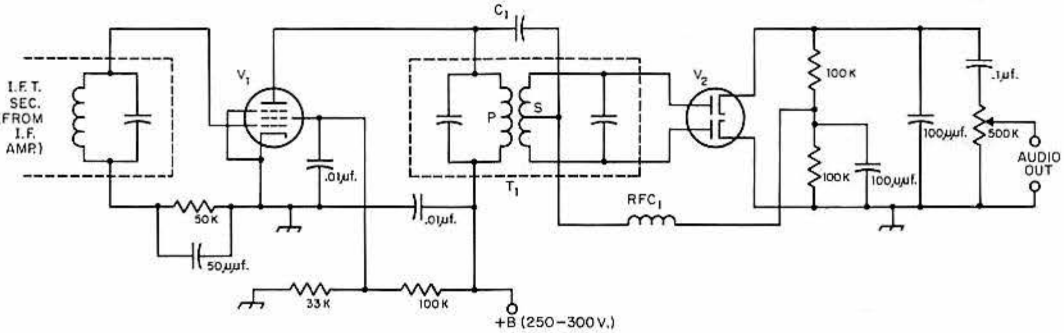


Fig. 3-10—Limiter-discriminator circuit frequently used at 455 k.c. in the form of an "adapter" for communications receivers, for reception of narrow-band f.m. signals.

C_1—App. 100 pf. for 455-kc. i.f.; 50 pf. for higher frequencies.

T_1—Discriminator transformer for i.f. used. Push-pull diode transformer may be substituted.

RFC_1—10 mh. r.f. choke for 455-kc. i.f.; 2.5 mh. satisfactory for frequencies above 3 Mc.

V_1—6AU6 or equivalent.

V_2—6AL5 or equivalent.

symmetrical and not too steep-sided, the signal may sound quite satisfactory, if it is strong. Some v.h.f. operators may not notice immediately that the signal is frequency modulated. There is a narrow segment of distortion at the center, but since the signal is usually approached from one side or the other the operator stops tuning before the distortion point is reached, and thus is quite happy with what he hears. But if the signal is weak he will notice that the audio level appears low compared with an a.m. signal of the same strength. If he does not recognize the f.m. for what it is, he is sure to complain of "low modulation."

If the receiver has the very steep-sided response that the mechanical filter and the crystal-lattice filter provide (and these are used in many of our better h.f. receivers) f.m. detection by the slope method produces considerable distortion and is inefficient. In any case, the frequency deviation of the transmitter must be adjusted to fit the receiver response exactly, if maximum audio recovery is to be achieved. Since there is wide variation in i.f. characteristics of communications receivers currently in use, this optimum matching is seldom achieved except by accident or individual experiment.

If the receiver bandwidth is appreciable, as in nearly all small a.m. transceivers for the v.h.f. bands, the operator will hear little audio on an f.m. signal where the deviation has been adjusted for satisfactory audio on a communications receiver-converter setup. If the transmitter deviation is increased to give usable audio level with the unselective transceiver, the result is an unintelligible jumble to the user of a communications receiver. Wide-band f.m. occupies more space than can be spared in crowded bands, so it is not permitted below 52.5 Mc.

Wide-Band F.M. Reception

Wide deviation is capable of producing a remarkably noise-free signal in a receiver specifically designed for it. The true f.m. detector will have a characteristic similar to that shown in Fig. 3-9B. The output is zero when the unmodulated carrier is tuned to the center, *O*, of the characteristic. When the frequency swings higher, the rectified output amplitude increases in the positive direction (in this example), and when the frequency swings lower the output amplitude increases in the negative direction. Over the range in which the characteristic is a straight line the conversion from f.m. to a.m. is linear and there is no distortion. One type of detector that operates in this way is the *frequency discriminator*, which combines the f.m.-to-a.m. conversion with rectification to give an audio-frequency output from the frequency-modulated r.f. signal.

Limiter and Discriminator

A practical discriminator circuit is shown in Fig. 3-10. The f.m.-to-a.m. conversion takes place in transformer T_1, which operates at the intermediate frequency of the receiver. The voltage induced in the transformer secondary, S, is 90 degrees out of phase with the primary current. The primary voltage is introduced at the center tap on the secondary through C_1 and combines with the secondary voltages on each side of the center tap in such a way that the resultant voltage on one side of the secondary leads the primary voltage and the voltage on the other side lags by the same phase angle, when the circuits are resonated to the unmodulated carrier frequency. When rectified, these two voltages are equal and of opposite polarity. If the frequency changes, there is a shift in the relative phase of the voltage components that results in an increase in output amplitude on one side of the secondary and a corresponding decrease in amplitude on the other side. Thus the voltage applied to one diode of V_2 increases while the voltage applied to the other diode decreases. The difference between these two voltages, after rectification, is the audio-frequency output of the detector.

The output amplitude of a simple discriminator depends on the amplitude of the input r.f.

signal, which is undesirable because the noise-reducing benefits of f.m. are not secured if the receiving system is sensitive to amplitude variations. A discriminator is always preceded by some form of amplitude limiting, therefore. The conventional type of limiter also is shown in Fig. 3-10. It is simply a pentode i.f. amplifier, V_1, with its operating conditions chosen so that it "saturates" on a relatively small signal voltage. The limiting action is aided by grid rectification, with grid-leak bias developed in the 50,000-ohm resistor in the grid circuit. Another contributing factor is low screen voltage, the screen voltage-divider constants being chosen to result in about 50 volts on the screen.

In tuning a signal with a receiver having a discriminator or other type of f.m. detector the tuning controls should be adjusted to center the carrier on the detector characteristic. At this point the noise suppression is most marked, so the proper setting is easily recognized. An amplitude-modulated signal tuned at the same point will have its modulation "washed off" if the signal is completely limited in amplitude and the discriminator alignment is symmetrical. With either f.m. or a.m. signals, there will be a distorted audio-frequency output if the receiver is tuned "off-center." With slope detection of f.m. the situation is reversed. Minimum distortion is achieved with the signal detuned to one side of center, and distortion appears at the middle.

We have described slope detection as a narrow-band technique, and the f.m. discriminator-limiter as wide-band, but either method can be used with any bandwidth. It is mainly the widespread use of communications receivers of the a.m. type that makes slope detection primarily a narrow-band approach. When wide-band f.m. is received on the typical v.h.f. transceiver, it is by slope detection. Conversely, many communications receivers can be equipped with outboard f.m. limiter-discriminator circuits, which make them efficient detectors of narrow-band f.m.

It is important to remember that there is no one receiving system that works with both wide- and narrow-band f.m. transmission. But when the best possible receiving method is used, f.m. of any deviation is capable of producing results in voice communication that compare favorably with the best a.m. detection systems.

Frequency modulation has shown its worth to the extent that it is almost standard for all v.h.f. communication outside the amateur bands. Police, fire, commercial mobile services and the like are not bothered by the receiver problems discussed above, as their work is between various station units all designed to work together. Transmitters and receivers are both crystal-controlled, and are set up on a prescribed channel and left there for the life of the equipment. The demand for frequencies for this use soon exceeded the supply, and channel widths were cut down by law, to permit assignment of more channels in a given portion of the spectrum, making large quantities of used but serviceable equipment available for conversion to amateur service.

Fixed-frequency operation may be foreign to a v.h.f. operator's concepts of what amateur communication should be, but it has many uses and the availability of equipment at moderate cost has produced something approaching a boom in f.m. in the 50-, 144- and 420-Mc. bands. Reliable emergency communication, radio club or net liaison, extension of local coverage through strategically-located repeater stations—these are just a few of the possibilities opened up by the f.m. equipment bonanza.

Technically, this equipment is classed as "wide-band." Unfortunately, it is rather in between true wide-band and narrow-band f.m. The deviation is too great to permit reception on typical a.m. gear used by amateurs, and too narrow to work well with i.f. bandwidths used in f.m. broadcasting and in TV sound. It is very effective where both receiver and transmitter from this class of service are used.

Wide-Band F.M. with Simple Gear

An application for true wide-band f.m. that seems not to have occurred to most v.h.f. enthusiasts is its use in connection with simple modulated-oscillator transmitters. The transmitting aspect of this is discussed in Chapters 5 and 6, but the important point is that in modulating a v.h.f. oscillator the first thing you get is f.m. Swinging the plate voltage on a simple oscillator results in changes in frequency, before there is appreciable amplitude modulation. The resulting signal may not be simon-pure f.m., and it is out of place in a crowded band, but it sounds better than you'd think on a receiver of the bandwidth used in f.m. broadcast reception.

This suggests a simple communications system for frequencies where there is room for it: oscillator or oscillator-amplifier transmitters of simple design and converters working into f.m. broadcast receivers. Purists who insist on T-9 notes and narrow-band reception will frown on this approach, and admittedly it is far from the ultimate in technical development, but it does work extremely well in a practical and low-cost way. The upper portions of the 144- and 220-Mc. bands could support quite a bit of this sort of thing without bothering anyone, and nearly all the vast expanse of the 420-Mc. band except a small segment beginning at 432 Mc. is fair game for it. If there is an effective "simple way" left to ham radio in the world above 50 Mc., this is it.

Reception can be handled in several ways. The nearest thing we know of to a universal receiving system for the v.h.f. man is a converter of conventional design working into a communications receiver having provision for wide-band f.m. reception. The Hallicrafters S-240 is one such receiver made today, and several useful models are on the used-receiver market. The best for our purposes are the Hallicrafters SX-42, SX-62 and SX-43. These have two i.f. channels: one for the usual communications bandwidths and the other for wide-band f.m. and a.m. The wide channel comes into play on all frequencies above 30 Mc. and is optional for 27 to 30 Mc. A converter working into the receiver at 27 Mc. thus may be used for conventional narrow-band work, or for wide-band f.m., at the flip of a switch.

A somewhat less versatile model is the Hallicrafters S-27, or its later version, the S-36. This series has no narrow-band channel, but is equally good for wide-band work. All these receivers, incidentally, are fine for f.m. broadcast reception. They are also handy for checking the v.h.f. range for skip propagation, since their coverage is continuous above 27 Mc.

Reaching farther back into history, the f.m. broadcasting band was once 42 to 50 Mc. There are a few receivers for this range still around, and they work well for our purpose if used with converters for 144, 220 or 420 Mc. The more modern f.m. receivers, for the band at 88 to 108 Mc. may also be used in this way. Tunable converters are practical for wide-band f.m., because the small amount of drift and mechanical instability inherent in the average good design is not enough to cause trouble in this receiving application.

Other i.f. systems may be pressed into service with tunable converters. The sound i.f. system of a defunct TV receiver is a possibility. Deviation in TV sound transmission is 25 kc., whereas f.m. broadcasting is 75 kc. A complete receiver for the low-band commercial mobile channels, or just the i.f. system and audio of any mobile f.m. unit, could be used with a tunable converter. It should be remembered that the bandwidth of these receivers is 30 kc. (transmitter deviation 15 kc.) and their use will impose strict requirements on oscillator stability in both transmitter and receiver. Building a complete receiver for wide-band f.m. is not a particularly difficult job. All the necessary i.f. components are readily available, or can be salvaged from used gear. Front-end tuning is uncritical mechanically and electrically, because of the greater bandwidth involved. Inexpensive portable f.m. receivers might offer promise as tuners for portable f.m. transceivers.

Whatever system is employed, the results are likely to be a pleasant surprise to anyone accustomed to the high noise level and restricted audio ranges that are characteristic of most voice communication. Wideband f.m., even with relatively simple equipment, is capable of excellent voice quality and almost complete elimination of noise. In practical communications range it is approximately equal to the best a.m.

TUBES OR TRANSISTORS?

It is no news to most readers that v.h.f. reception today is swinging to transistors and other solid-state devices, to the almost complete abandonment of vacuum tubes. It is the fashion to state that anyone who uses tubes today is hopelessly out of date. What is the real truth of the transistor-tube controversy?

In favor of transistors it can be said that they have opened up a whole new field for small, lightweight, low-cost portable v.h.f. gear. They reign supreme here, giving performance better than was ever possible with tubes, with battery economy and portability never before approached. In recent years, transistors have also made possible r.f. amplifiers and complete v.h.f. converters having extremely low noise figure. Particularly at 144, 220 and 420 Mc., solid-state devices equal or exceed tubes in low-noise reception capabilities. But with the common bipolar transistors, which until recently have yielded the best noise figures, we have had to accept limitations along with the gains. Being small-signal diodes, bipolar transistors are very fussy as to overloading.

The field-effect transistor (FET) now promises to correct much of this, having dynamic qualities much like the better tubes. The FET is now available in types that work very well up to at least 500 Mc., and in the amateur v.h.f. bands they offer many advantages. But where performance is the only criterion, there may be no marked advantage in *conversion* to solid-state techniques.

The amateur who has been in the game for some years will have accumulated a considerable inventory of equipment designed around vacuum tubes. Where he has the power supplies and other accessories for tube operation, he may stand to gain little by converting to transistors, and have nothing to lose by continuing with tubes. The newcomer making a fresh start on the v.h.f. bands might, on the other hand, wish to avoid use of vacuum tubes for everything except high-power amplifiers in his transmitters. In receiving, at least, he should give the solid-state approach serious consideration.

V.h.f. Receivers, Converters and Preamplifiers

This chapter will deal with the actual construction of receiving equipment for the bands from 50 through 450 Mc. To conserve space and present the greatest possible variety of practical examples for the home-builder of v.h.f. gear, the material to follow will lean heavily on Chapter 3 to explain why and how certain circuits are used. Our text will assume that the reader is familiar with what has gone before. If you have not already done so, it would be well to read Chapter 3 carefully before starting construction of any of the units to be described. It might save you some time and trouble in the end.

Though "v.h.f." is the label given to frequencies from 30 to 300 Mc., amateur practice tends to include the 420-Mc. band in the v.h.f. picture. Thus examples of gear for that band will be found occasionally in our v.h.f. chapters. Where the equipment design is essentially the same as that employed on lower frequencies we will lump 420 with 50 through 225 Mc. in our descriptive material. Examples of the use of true u.h.f. circuit techniques for 420-Mc. gear will be found in Chapter 10.

AN EASY-TO-BUILD TUNER FOR 14 TO 18 MC.*

Some means of listening is the first requirement of the newcomer to amateur radio. It may be one of the thorniest problems for the v.h.f. beginner. It is almost standard practice in v.h.f. circles to use a converter ahead of a communications receiver, but as explained in some detail in the preceding chapter, many receivers that are passable on lower bands are not well suited for use with v.h.f. converters. The v.h.f. newcomer thus may be in the market for a tuner covering 14 to 18 Mc. that will permit him to use converters effectively, without the expense of buying or building a first-class communications receiver.

The tuner of Fig. 4-1 was designed to fill this need. It can be built easily at moderate cost, and it works with converters fairly well. Building it gives the beginner good receiver experience, yet his investment is low enough so that he can afford to replace the tuner when he is ready for something better. Even then, it remains useful as a standby receiver, and for portable work.

The Nuvistor converters described later were built for use with this tuner as the receiving portion of a complete v.h.f. station pictured in Chapter 7. They are topnotch equipment, capable of the best in v.h.f. reception with good communications receivers. The investment in them is a long-term one. The tuner will enable you to get started in v.h.f. work with the converters, in an entirely practical way, and at low cost.

Tuner Design

You can listen on the 14-Mc. amateur band, and to various commercial and broadcasting services between the top end of that band and 18 Mc. with the tuner, so it makes an interesting project on its own. Consulting the circuit diagram, Fig. 4-3, it will be seen that the tuner uses two 6CB6s as i.f. amplifier and detector, followed by a two-stage audio amplifier using a 6CX8 triode-pentode. Power is obtained by plugging into the side of the modulator and power supply unit directly, or through a 4-wire cable of any convenient length. If the power supply has not yet been built, the tuner may be tested on any supply capable of delivering 150 to 200 volts d.c., at a few milliamperes, and 6.3 volts a.c. or d.c. at about 1½ amperes. The supply should be well-filtered, and have fairly good d.c. regulation, for best results with the type of detector used. If hum is encountered, try adding 4 μf. or more, from pin 6 to ground.

The detector tuning capacitor, C_1, is attached to a vernier dial (National Type AM-7). The actual tuning range is from just below 14 to just above 18 Mc., but the white dial scale taped to the front panel shows the equivalent v.h.f. ranges, 50 to 54 and 144 to 148 Mc. The calibrated scales can be added after the receiver is completed, and you have the range where you want it on the dial. Controls below the dial are i.f. gain, left, audio gain, center, and regeneration at the right.

Regeneration is the means by which we achieve a fair measure of performance from so simple a receiver. Three tubes may not seem like much in these days of umpteen-tube chrome-plated monsters, but this receiver is not unlike those that were in general use not too long ago. A regenerative or superregenerative detector is a marvelous device when properly

*A full-size template for drilling the tuner chassis is available from ARRL. Please send 25 cents and a stamped self-addressed envelope, and give the publication, edition, page number and name of the item for which the template is requested.

controlled, and with the tubes we have today they can be made to work much better than the blooper receivers our predecessors made out with in the '20s, and even in the '30s. Such a receiver requires a bit of skill and patience in tuning, but when you learn how to ride it, the regenerative detector will take you a long, long way! As a tuner following a crystal-controlled v.h.f. converter, its equal would be hard to find in the low-priced communications receiver category.

The amplifier stage, V_1, preceding the detector provides gain, but more important it isolates the detector from the converter stages, and makes control of regeneration a relatively simple matter. The gain control, R_1, allows the operator to feed signals to the detector at the optimum level for all types of reception, and while this makes for two-handed tuning and the need for a bit of juggling now and then, it helps the simple receiver to do its job in an effective manner.

The detector may be operated in three different conditions by varying the screen voltage (regeneration) control, R_2. At low screen voltages the detector works at low sensitivity, but in a completely uncritical manner, making it fine for strong local signals. As the voltage is turned up you hear the noise rise as the detector nears the oscillation point. Sensitivity and selectivity pick up here, and if the detector is adjusted carefully just below the point of oscillation, the sensitivity on modulated signals is very good. Condition 2 is reached when the detector goes into oscillation. In tuning through a signal you hear a beat note, just as with a communications receiver with its beat oscillator on. This is the c.w. or s.s.b. mode, and highest sensitivity is found just on the high side of the point where oscillation stops.

Condition 3, superregeneration, occurs at higher screen voltage, and is characterized by a loud "rushing" noise when no signals are being received. Only modulated signals can be copied with a superregenerative detector, for there is no audible beat with the incoming carrier; only a drop in the background noise when the signal is tuned in. The degree of quieting is dependent on signal strength, and the stronger signals (locals and some DX) quiet the noise almost completely. In superregeneration the detector is not easily overloaded, and tuning is uncritical. It is markedly insensitive to ignition and other impulse noise. Audio quality is inferior to other modes of detection, however, and the rushing noise takes some getting used to. Old-timers in the v.h.f. game will tell you that there is no music as sweet as the rush of a smooth superregen, but you will not love it that much, at first, if you're new to v.h.f. hamming!

Building the Tuner

Parts arrangement in the tuner is not fussy, but a layout template is available for drilling the chassis if you want it, as already mentioned. This is most useful if you use components mechanically similar to those in the original, a restriction that is not too important otherwise. Probably the only critical item is the main tuning capacitor, C_1. A double-bearing model with mounting feet front and rear is desirable here,

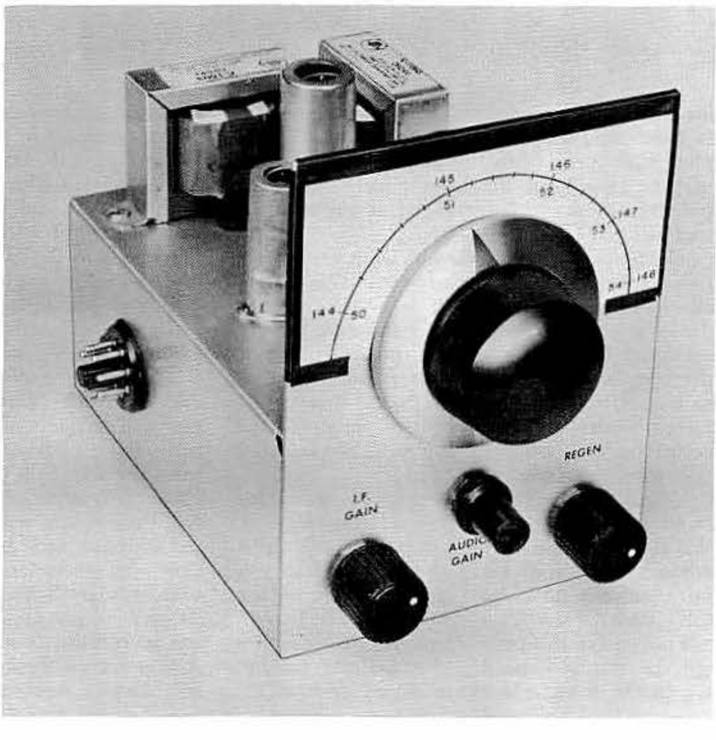


Fig. 4-1—The simple tuner is calibrated for the v.h.f. bands, though it actually covers 14 to 18 Mc. The calibration is drawn on white paper and taped to the area around the vernier dial. Controls below the dial are the i.f. gain at the left, the regeneration at the right, and audio gain, center.

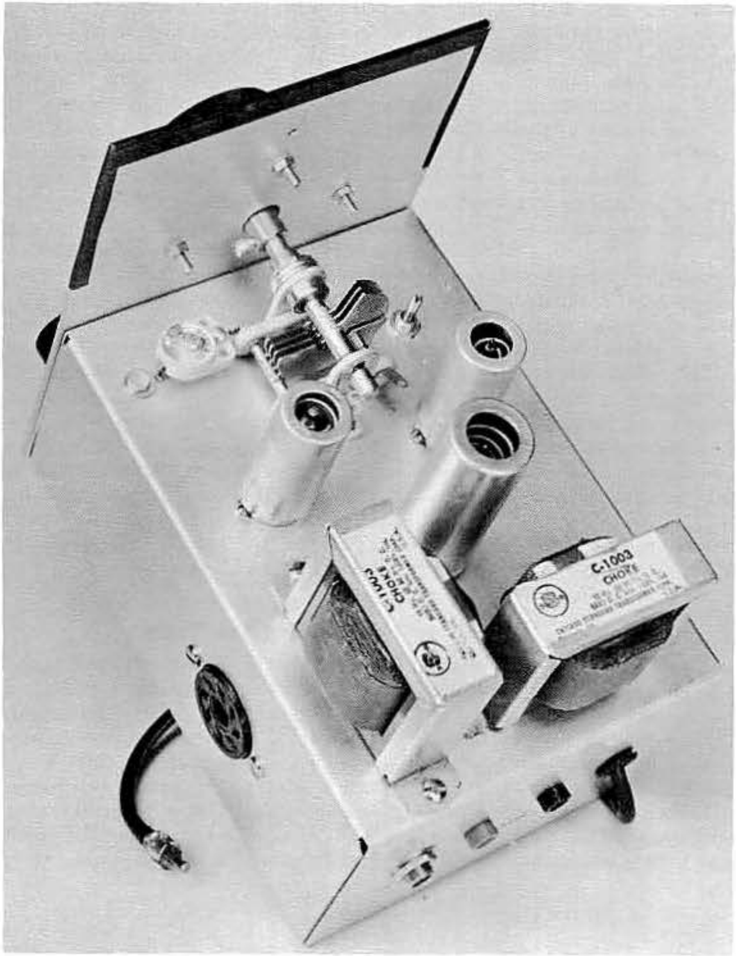


Fig. 4-2—Rear view of the tuner. Note that a double-bearing capacitor is used for tuning the detector circuit. The ceramic padder to the left is C_2. The detector is the left of the two smaller tubes. Asymmetrical arrangement of the two audio chokes is for minimum hum pickup.

for there may be a slight amount of backlash in the tuning with single-bearing types. A template that comes with the National dial can be used in laying out the front panel. The three potentiometers can be arranged in any convenient manner.

The power and audio circuits were wired with Belden Type 8885 shielded wire. This is not absolutely necessary, but it is a great aid in doing a neat job. Shielded leads can be any necessary length, and can be run in corners of the chassis or wherever convenience dictates, so long as their shields are bonded together at intervals with solder and held in place with an occasional grounding lug. But don't use shielded wire for any circuits carrying r.f.!

Use of insulated tie-point strips for mounting small parts also makes for a neat wiring job. When you assemble the tuner put these strips adjacent to each socket. Use whatever lugs you need and clip off unused ones when the wiring job is done.

Looking at the tuner from the top rear you see the tuning capacitor, C_1, and its padder, C_2, at the front. The ceramic padder is at the left, with its rotor lug clamped under a washer and its stator lug soldered to the front stator bar of C_1. At the right of C_1 is the tuning screw for the i.f. amplifier coil, L_1-L_2. A feedthrough bushing (National TPB) is mounted directly in back of the left-side stator lug of C_1. The 10-pf. fixed padder, the 50-pf. grid capacitor, and the top end of L_4 are connected to the underside of this feedthrough.

The detector tube, V_2, is at the left, and the i.f. amplifier, V_1, is at the right, just in back of C_1. The dual audio amplifier, V_3, is near the middle of the chassis. The two chokes at the rear and left side of the chassis are L_5 and L_6, respectively. These are used instead of audio transformers, and just about any small filter choke will serve. Audio transformers are also OK, though somewhat more expensive. The output coupling arrangement, L_6, the .05-μf. capacitor, and phone and speaker jacks, are for use with ordinary headphones or a speaker that has its own output transformer (a transformer for use with ordinary audio output tubes will do). A connection directly to the voice coil (low impedance) will not work with this coupling.

With speaker leads plugged into the tip jacks, J_1 and J_2, the speaker is connected automatically when the phones are removed from J_3. The position of the jacks and the hole for bringing out the coaxial input lead, on the back wall of the chassis, is not critical.

The male power plug, J_4, on the left side of the tuner when viewed from the front, fits into a matching socket on the right side of the modulator unit. To use the tuner at some distance from the modulator, a cable of the required length may be made with an Amphenol 78-S8 socket at the tuner end and an 86-CP8 plug at the modulator end. These should be covered with Amphenol 3-13 plug caps. Placement of the plugs and sockets in the side walls of the various components of the station is not important, so long as they all will match up.

Except for the i.f. and detector coils, L_2 and L_4, placement of other parts is not critical, and considerable variation from the original can no doubt be made without affecting results. The i.f. coil is a standard Miller slug-tuned unit of approximately 7-μh. inductance. The primary coil, L_1, is wound over the bottom turns of L_2. It is wound on, cemented in place, and left to dry while other work is done. The turns are in the same direction as the secondary, and the bottom ends of both windings are connected to ground. The top end of L_1 is brought to a tie point, where the coaxial line is connected to it.

The detector coils, L_3 and L_4, are made from a single piece of B & W Miniductor. Start with a piece having at least 20 turns. This fine-pitch coil stock can be cut readily if a sharp knife is held against the plastic supporting strips and heated with a soldering iron pressed against the top edge of the knife. When the blade nears the melting point of the plastic, the ribs can be cut easily. At the sixth turn press the wire down toward the axis of the coil. It may then be cut or broken. Thread the ends back out and unwind a half turn each side of the cut. Now unwind the outside turns until there is left a coil of 4 turns and one of 10½. The tap is made by pushing down the third turn up from the inner end of the larger winding. This makes a point that can be soldered to for the lead to the cathode of the 6CB6.

This assembly is mounted in a horizontal position supported on tie points by its leads, as shown at the upper right in the bottom-view photograph. The outer end of the larger coil goes to the feed-through bushing, the outer end of the smaller to the plate of V_1. The balance of the assembling and wiring is almost completely uncritical, though neatness and ease of adjustment will be served if leads are kept short, particularly in the circuits of the amplifier and detector tubes.

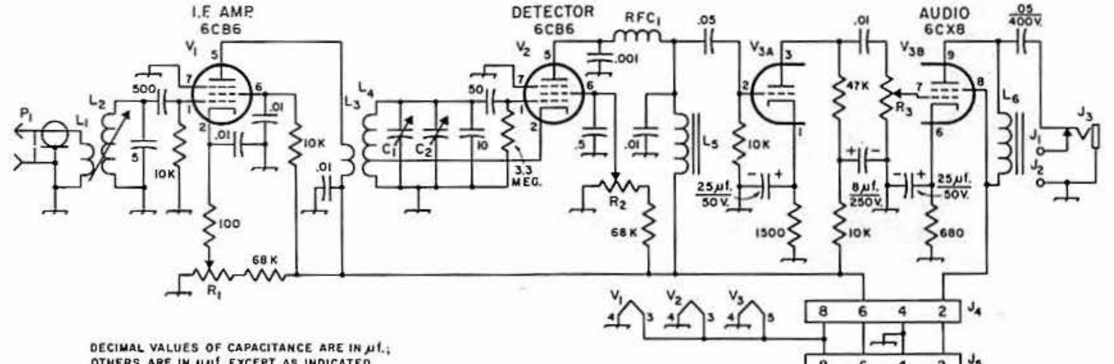


Fig. 4-3—Circuit diagram and parts information for the 14- to 18-Mc. tuner. Capacitors marked with polarity are electrolytic. Others are paper-tubular, ceramic or mica, 200 volts or more, unless marked. The .01- and .001-μf. are ceramic disk. Resistances are in ohms, resistors are ½ watt unless specified.

- C_1—50-pf. ($\mu\mu$f.) double-bearing variable (Hammarlund MC-50-S).
- C_2—4-30-pf. ($\mu\mu$f.) ceramic trimmer (Mallory ST554N or Centralab 822-EN).
- J_1, J_2—Insulated tip jack.
- J_3—Closed-circuit phone jack.
- J_4—8-pin male chassis fitting (Amphenol 86-CP8). Goes on left side of chassis.
- J_5—Octal socket (Amphenol 77-MIP-8).
- L_1—3 turns No. 24 insulated wire wound over low end of L_2.
- L_2—4½- to 10-μh. iron-slug coil (Miller 21A826RB1).
- L_3—4 turns No. 24 tinned, 32 t.p.i., ½-inch diam.
- L_4—10½ turns like L_3. Both are made from single piece of B & W Miniductor No. 3004. See text. Tap at third turn from inner end.
- L_5 L_6—16-hy. 50-ma. filter choke (Stancor C-1003).
- P_1—Shielded phono plug, attached to 18-inch length of small-diameter 52- or 75-ohm coaxial cable.
- R_1, R_2—20,000-ohm control (25,000-ohm also suitable).
- R_3—500,000-ohm control, audio taper.
- RFC_1—100-μh. r.f. choke.
- V_1, V_2—6CB6.
- V_3—6CX8.

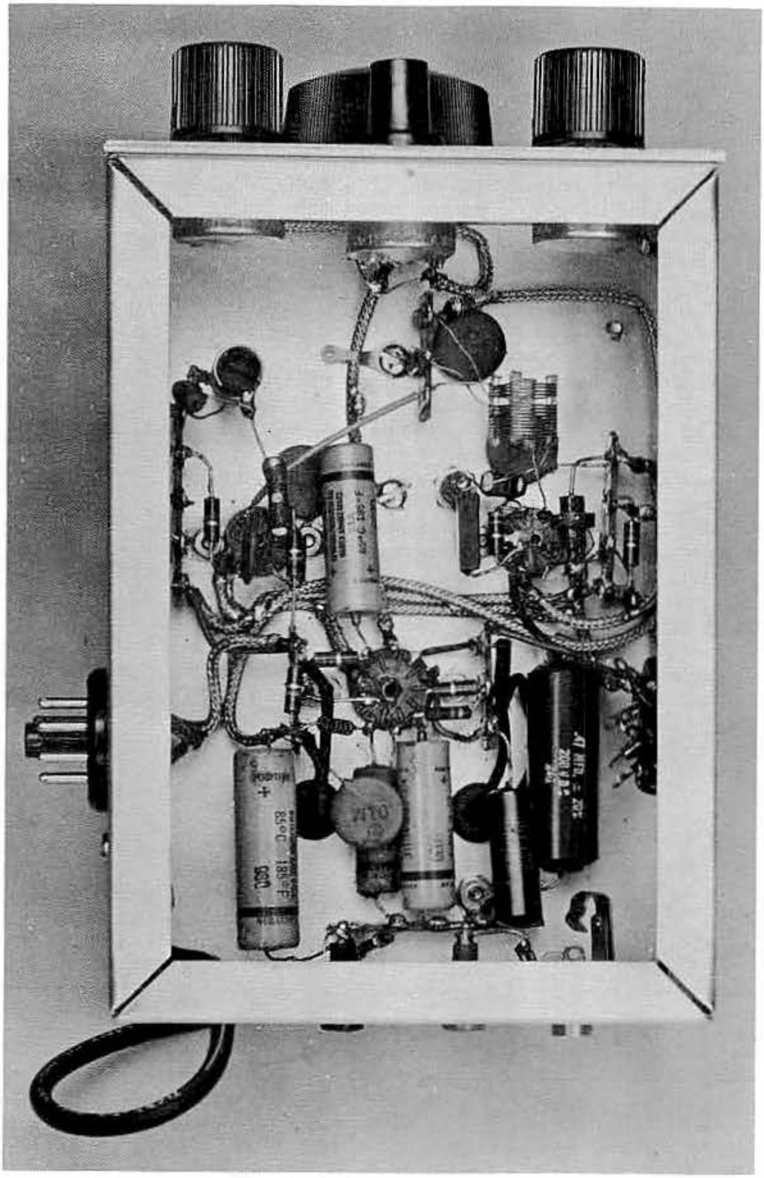


Fig. 4-4—Bottom view of the tuner. Arrangement of parts, other than the i.f. amplifier coil, L_2 (upper left) and the detector coil, L_4 (upper right), is not particularly critical. Power and audio circuits are wired with shielded wire.

Adjustment and Operation

If the tuner has been wired correctly it should be possible to hear signals of some sort on it almost at once. Apply 6.3 volts a.c. for the heaters between Pins 4 and 8 of J_4. Temporarily connect Pins 2 and 6 together and apply plate voltage, preferably not much over 150 volts at first, positive to Pins 2 and 6, negative to Pin 4. Plug in the phones or speaker. Have all three potentiometers turned down. First try the audio gain control, R_3. Turning it up should bring up the level of noise, and possibly hum. Set it at a comfortable level, and turn the i.f. gain, R_1, about three-fourths on. Turn up the regeneration control, R_2, until a rushing sound is heard. Attach a few feet of wire to the tip of the plug, P_1, and turn the dial slowly.

Some signals should be heard, unless you have hit one of those rare times when the 14-Mc. range is completely dead. When you find a signal, experiment with the setting of the i.f. gain control, R_1, and the regeneration control, R_2. If you have never used a regenerative or super-regenerative detector before, the intricacies of adjusting it properly will take some learning. Practice with various signals, trying the three conditions mentioned earlier. If you have trouble with a high-pitched audio whistle, try a lower value of detector grid leak: 1 to 2 megohms, in place of the 3.3 megohms shown.

Now, you're ready to peak things up, and get the dial calibration around to what you want. The tuning capacitor, C_1, has the ceramic capacitor, C_2, connected across it, so the setting of the latter will markedly affect the tuning range of your vernier dial. If you have made your coil correctly, setting C_2 to near maximum capacitance will place the 14-Mc. amateur band near the maximum-capacitance end of the tuning range of C_1. If you succeed in locating the amateur band you will find c.w. signals at the low frequency edge, and phone signals above them. Adjust C_2 gradually until the lowest-frequency amateur c.w. signal comes in with the dial close to its maximum-capacitance setting. A good signal to look for now is WWV or WWVH on 15 Mc. One or the other of those stations, perhaps both, will be receivable at least part of the time almost anywhere in the United States. With these and the low end of the amateur 14-Mc. band, you have the first megacycle of your tuning range well marked.

Note that an indicating pointer for the dial is made by sticking a triangular-shaped piece of black plastic tape to the nickel-plated rim. Put the capacitor at the maximum setting, and then attach the pointer to the rim so that it is bisected by the left side of an imaginary horizontal line drawn through the center of the dial. When you turn the dial around to bring the capacitor plates all out, the mark will be at the right side. If you have used components similar to the original, you can set the padder, C_2, so that 14 Mc. is just above the horizontal point at the left, and 15 Mc. will come just a bit to the left of vertical. The next megacycle of tuning, to 16 Mc., will occupy slightly less space, and the third and fourth megacycles (to 17 and 18 Mc.) progressively less. This is a fortunate result of the plate shape in the tuning capacitor: the more active lower halves of the v.h.f. bands you will eventually be tuning will be spread out more than the less-occupied frequencies at the high ends.

When you get your converters working, 14 Mc. will be 50 or 144 Mc., 15 Mc. will be 51 or 145 and so on. The tuner will operate almost exactly the same when working with the converters as it now does on 14 to 18 Mc., except that the frequencies covered will be in the v.h.f. range. For the moment, you can tune 14 to 18 Mc., and there is a lot going on there most of the time. It won't do any harm to practice tuning, for one of the prices of performance with simple equipment is some trickiness in operation. There is more to running this one than turning the dial!

With the tuner plugged directly into the power supply you may find that the hum level is too high to suit you. This is the result of inductive pickup from the power-supply components by the chokes in the tuner audio circuits. The position of the chokes was adjusted for minimum hum pickup, but it is still considerable at high audio levels. Running the tuner with even a short cable between it and the power supply will bring down the hum level markedly. Use of completely shielded chokes or audio transformers also reduces the hum level, but at higher cost.

CRYSTAL-CONTROLLED CONVERTERS FOR 50 THROUGH 436 MC.*

The tuner just described and the converters for 50 and 144 Mc. in this series are parts of a complete station for these bands shown in Chapter 7. The 144-Mc. model worked so well that it was decided to make one for 220 Mc. As more experience with Nuvistors accumulated they were found to be very effective for 420 Mc. as well, so a converter for 432 to 436 Mc. was worked out along similar lines.

All converters of the set work well with the simple tuner, Fig. 4-1, or with communications receivers tuning the 14-Mc. range. Coil and crystal information is supplied for use of other i.f. ranges as well. Power connections are set up for the "Two-Band Station," but anything delivering around 150 volts d.c. at 30 ma., and 6.3 volts a.c. at 1 amp. will do for the models for 50 through 220 Mc. The 432-Mc. converter requires about 50 ma.

In Fig. 4-5 the 50- and 144-Mc. converters are shown together, with the 50-Mc. one at the right. The 220-Mc. model is identical in external appearance to the 144-Mc. one, so it is not shown except in the interior view. Construction and adjustment of each will be treated separately, insofar as there are differences between them. The Nuvistors shown are all 6CW4s, but 6DS4s may be used interchangeably, both as to operating conditions and results.

Construction

The converters for 50, 144 and 220 Mc. are built in aluminum Miniboxes, 3 by 4 by 5 inches in size. The Nuvistor sockets have small metal tabs that are bent down against the underside of the chassis to provide grounding. These are clamped under washers by 4-40 screws and nuts on opposite sides of the sockets. The socket hole should be ½-inch diameter, with small notches filed out for the tabs. The ceramic trimmers in the 144- and 220-Mc. converters, C_1, C_2, and C_3, also require notched holes.

Leads in r.f. circuits should be as short as possible. Power wiring can be placed for neatness, but keep insulated power leads close to the chassis. Use terminal strips for holding resistors in place, and lugs bolted to the chassis for grounding.

*Full-size templates for drilling the top surfaces of the converters for 50, 144 and 220 Mc. are available from ARRL. Please send 25 cents and a stamped self-addressed envelope, and give the publication, edition, page number and name of the item for which the template is requested.

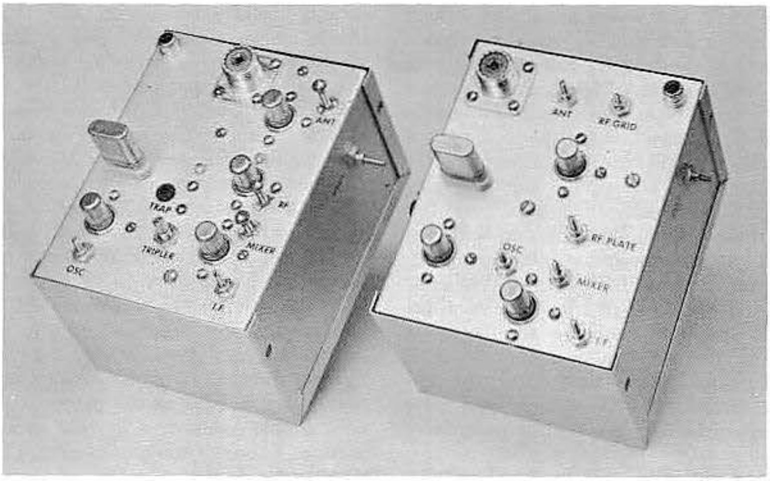


Fig. 4-5—The 50- and 144-Mc. converters are built in standard aluminum boxes, and fitted with plugs that line up with the power connectors in the tuner of Fig. 4-1 or the power unit shown in Chapter 7. The 50-Mc. converter is at the right.

Checking the Oscillators

The crystal oscillator is checked first. The meter in the bridge unit described in Chapter 11 or any other 1-ma. meter, may be used to measure oscillator plate voltage, or a voltmeter will serve if you have one for the 100-volt d.c. range. To use a 1-ma. meter, connect a 100,000-ohm resistor in the positive lead and ground the negative lead. It is not important for this purpose that the 1000-ohm resistor shown in the bridge unit, be included.

Working on the converters is easier if a 3-wire power cable with suitable plugs is used, rather than plugging the converters directly into the tuner or power unit. Tests may be made with all tubes in their sockets, as the dropping resistors in the plate leads prevent excessive current. Apply power to the converter. Touch the free lead of the 100,000-ohm resistor to the B-plus end of the oscillator plate coil. The meter indicates 100 volts d.c. for full scale. The voltage reading obtained will depend on whether the tube is oscillating or not. The oscillator current runs through a 10,000-ohm resistor, so the more current the tube draws the lower the voltage will be. When the circuit oscillates, plate current drops, and the indicated voltage rises.

Use of Ohm's Law will tell you what the plate current is, though this need not be found except as a matter of interest. With the core stud all the way up, the circuit probably will oscillate, and the meter indication will be around 0.7 (70 volts). Turn the stud into the coil, watching the meter. It will rise to around 0.9 (90 volts) and then drop suddenly as oscillation stops, to around 0.5 (50 volts). These represent actual plate currents of 8, 6, and 10 ma., respectively.

Readings may vary considerably from the above, due to differences in crystals and other parts. The important points are the gradual rise (increasing vigor of oscillation) and then the sudden dip as oscillation ceases. Set the slug for the highest reading (lowest oscillator plate current) at which the oscillator will start each time power is applied. The frequency can be checked with a calibrated wavemeter or grid-dip meter. It should be the frequency marked on the crystal, and no other.

The 50-Mc. Converter

Three 6CW4s are used in the 50-Mc. converter. The first, a neutralized r.f. amplifier, is in the upper center of the converter as shown in Fig. 4-5. At the bottom right is the mixer tube, and to its left is the crystal oscillator. The 36-Mc. crystal is in the left center, and above it is the antenna connector.

Turn now to the circuit diagram, Fig. 4-6. The tuned circuits L_2 and L_3, with the small coupling capacitor, C_2, are used to give some selectivity in the r.f. amplifier grid circuit. The tuning screws for the coils are visible at the top of the first photograph. Similar circuits are used between the amplifier plate and mixer grid (L_5, L_6 and C_3) and these are at the right side of the top view. The oscillator coil, L_8, is in the lower center. The mixer plate coil is in the lower right corner of Fig. 4-7. The neutralizing coil, L_4, is mounted horizontally, with its adjusting stud coming out of the side of the box. The i.f. output connector is in the upper right corner of the top view.

The trap circuit, L_1C_1, is optional. Its purpose is to absorb Channel-2 video signals that might cause interference to 50-Mc. reception, as the result of the second harmonic of the oscillator (72 Mc.) beating with a Channel-2 TV signal. $(72 - 14 = 58)$ Unless you are near a Channel-2 TV station you will not need the trap, and the connection from J_1 can be made directly to the tap on L_2.

The bottom view of the converter, Fig. 4-7, is inverted vertically from the top view. The antenna connector and the trap circuit are in the lower left corner. To the right are the coils L_2 and L_3, and the i.f. output connector. Near the middle is the r.f. amplifier socket, and in line with it at the top is the mixer socket. The crystal oscillator tube socket is at the upper left. The oscillator plate coil, L_8, and the mixer grid coil, L_6, are in the same plane to the right. Directly below L_6 is the r.f. plate coil, L_5. The i.f. output coil, L_7, is in the upper right corner, connected through a shielded lead to the output connector in the lower right. The neutralizing coil, L_4, is just above the latter, with its tuning screw projecting through the side of the box.

The coupling capacitors, C_2, C_3 and C_4, are made by twisting insulated wires together to form small capacitances where needed. This is a convenient and inexpensive way of doing the job, and since the values are not particularly critical, the twisted wires serve just as well as would a fixed or variable capacitor of conventional design.

Power is taken from the 150-volt and 6.3-volt sources in the power supply described in Chapter 7. The 8-pin power plug, J_3, is mounted in the side of the converter case. It should be positioned so that it will line up with the socket on the side of the tuner, Fig. 4-1, or the similar socket on the modulator, if the tuner is not used.

The 144-Mc. Converter

The 144-Mc. r.f. amplifier uses two 6CW4s instead of one, and an oscillator-multiplier system is needed for developing the injection voltage for the mixer. Hand-wound coils are used in the r.f. circuits, instead of slug-tuned coils. The first amplifier, a neutralized triode stage, is followed by a grounded-grid stage, in the manner of the familiar series-cascode v.h.f. amplifier. The crystal oscillator works on 43.333 Mc., and drives a crystal-diode frequency tripler to 130 Mc. This injection frequency beats with signals at 144 to 148 Mc. in the mixer, producing an i.f. of 14 to 18 Mc., as before.

Looking at the top view, Fig. 4-5, we see the r.f. amplifier and mixer tubes in line vertically at the right side of the converter. The crystal oscillator is at the lower left. The capacitor C_5, which tunes the diode tripler circuit, is in the lower center of the picture. Just above is a grommet inserted in the hole over the trap capacitor, C_4, of which more later. The antenna connector is in the middle of the top portion, and the i.f. output connector is in the upper left.

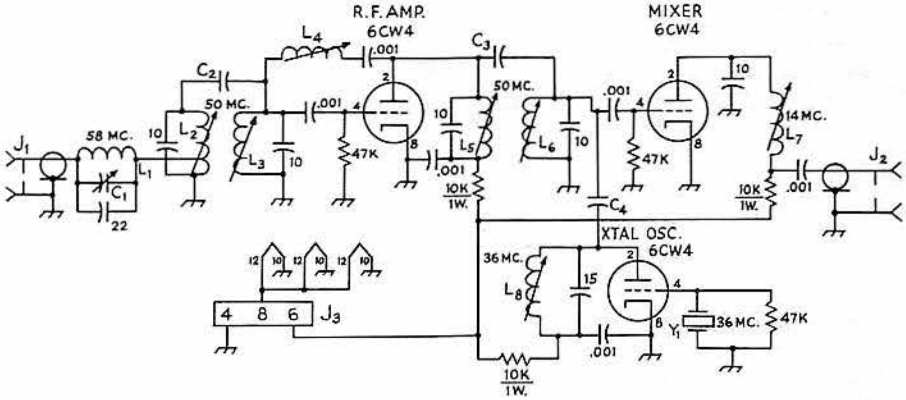


Fig. 4-6—Schematic diagram and parts information for the 50-Mc. converter. Resistors ½ watt unless specified. Fixed capacitors are ceramic; decimal values in μf., others in pf.

C_1—3–30-pf. mica trimmer.

C_2, C_3—No. 22 insulated hookup wires 2 inches long, twisted together for approximately 1¼ inches.

C_4—Same, but 1-inch wires twisted for ½ inch.

J_1—Coaxial connector, SO-239.

J_2—Phono jack.

J_3—8-pin plug (Amphenol 86-RCP8).

L_1—5 turns No. 18, ½-inch diam., 8 t.p.i. (B & W No. 3002).

L_2—10 turns No. 28 enam., close-wound on ¼-inch iron-slug phenolic form, tapped at 3 turns, 0.65 to 1.3 μh. (Miller form No. 20A000RBI).

L_3, L_5, L_6—8 turns No. 28 enam., close-wound on ¼-inch iron-slug phenolic form. Range 0.43 to 0.85 μh. L_3 set for 0.64 μh., L_5 for 0.66, L_6 for 0.73 μh. (Miller coils No. 20A687RBI). L_2 and L_3 are ⅞ inch apart c. to c. L_5 to L_6 is ¾ inch; L_7 to L_8 is ⅞ inch.

L_4—No. 32 enam., close-wound ⅛ inch on ¼-inch iron-slug phenolic form; 3.8 to 8.5 μh., set for 6.9 μh. (Miller coil No. 20A686RBI).

L_7—Universal-wound coil, 4.7 to 10 μh., set for 7.9 μh. (Miller coil No. 20A826RBI).

L_8—8 turns No. 32 enam., close-wound on ¼ inch iron-slug phenolic form; 0.67 to 1.25 μh., set for 0.94 μh. (Miller coil No. 20A106RBI).

Y_1—36-Mc. crystal (International Crystal Mfg. Co. F-605).

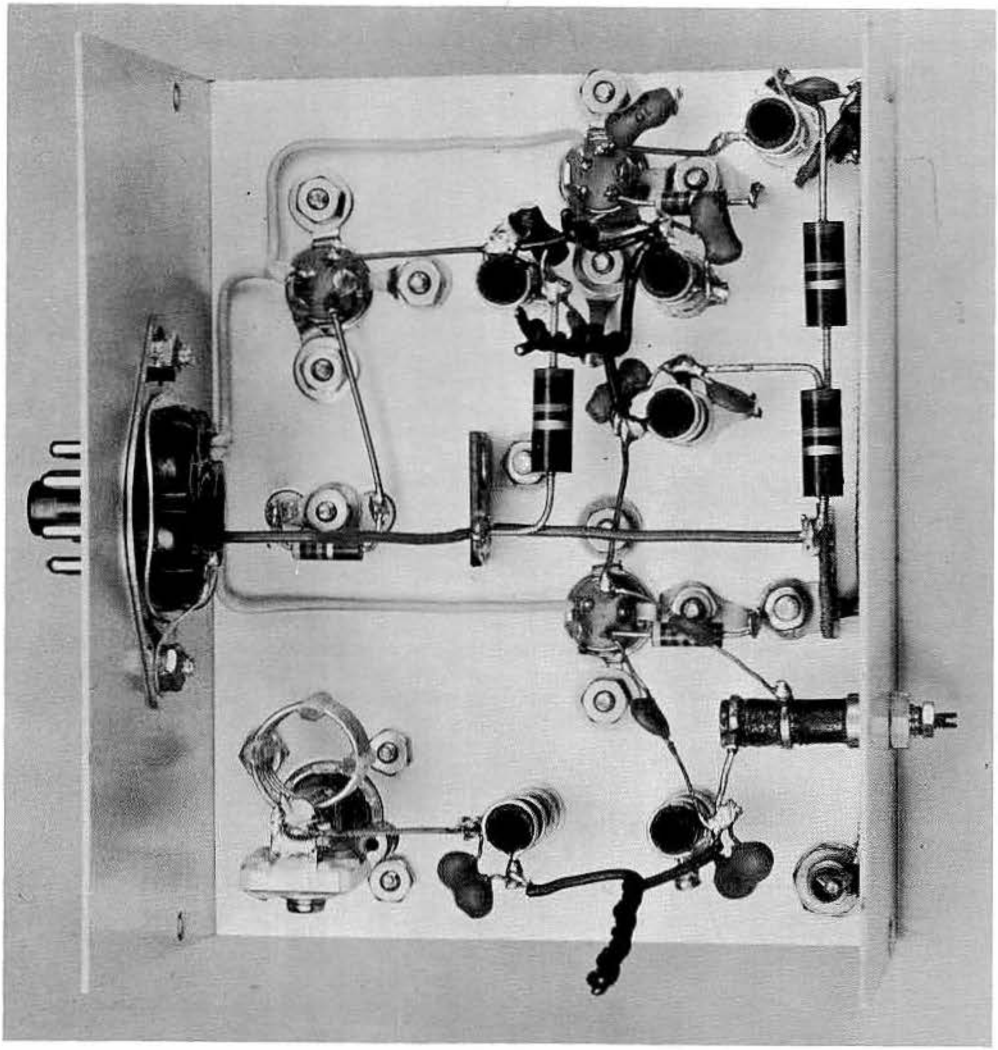


Fig. 4-7—Bottom view of the 50-Mc. converter. The antenna connector and trap circuit are in the lower left corner. The neutralizing coil, L_4, is mounted horizontally at lower right.

The bottom view was made by rotating the unit vertically, so the antenna connector appears at the bottom. The first amplifier grid circuit, L_1C_1, is in the lower right corner. Above it is the neutralizing coil, L_2, mounted on the side of the box. The two tinned-wire coils side by side just above and to the right of center are for the amplifier plate, L_3, and mixer grid, L_4. To their left is the trap circuit, C_4L_9, tuned to the second harmonic of the oscillator, 86.67 Mc. The coil with its axis at right angles to these is L_8. It is tuned to 130 Mc. by C_5, which appears in the upper center of the picture. The oscillator plate coil, L_6, and the mixer plate coil, L_5, are in the upper left and right corners, respectively.

The Diode Multiplier and Trap Circuits

Frequency multiplication with crystal diodes may be new to some readers, but it is a simple and effective way of developing injection voltage in the v.h.f. range. Diodes do the job easily, and at less cost than a vacuum tube. The diode works at low impedance, so it is connected between a loop (L_7) around the oscillator coil and a tap on the tuned circuit L_8C_5. The latter should be fairly high-C, so that the desired harmonic, in this instance the third, will be accentuated, and other harmonics of 43.3 Mc. suppressed.

There will be some energy at unwanted harmonic frequencies passed on to the mixer grid circuit. The trap, L_9C_4, is inserted in the lead to L_4 to suppress the second harmonic, 86.6 Mc. This trap circuit need be included only if local interference makes it necessary. In the Hartford area an f.m. station just above 100 Mc. rode through around 14.2 Mc. ($100.8 - 86.6 =$

14.2), but the trap removed the interfering signal completely when tuned to twice the crystal frequency. Removing the offending harmonic from the mixer circuit was the best way of handling the problem. A trap in the antenna circuit to absorb the interfering signal was tried but it resulted in a slight deterioration of the converter noise figure at 144 Mc.

50-Mc. Adjustments

The 50-Mc. converter is now ready to receive strong signals, as soon as it is connected to the receiver or tuner. The latter has a cable and plug for connection to the i.f. output jack, J_2. To use a communications receiver, make up a cable of any small coax, putting a phono-pin plug on one end. The other end connects to the receiver antenna terminals. This may require a coax fitting for some receivers, but most have screw terminals. Connect the inner conductor to the antenna terminal and the outer sheath to the ground terminal or the receiver chassis. Do this with the shortest possible leads, to keep down pickup of signals at 14 Mc. See Chapter 3 for hints on reducing i.f. pick up.

Now a 50-Mc. signal is needed. This can be from a grid-dip oscillator, a nearby 50-Mc. station, the harmonic of your transmitter, or ideally, a good signal generator. For any except the last, connect some kind of antenna to J_1. A short piece of wire will do at first, and the length can be varied to suit the strength of the signal. Set the stud in L_4 at about the middle of its range. Next, peak the screws in L_2, L_3, L_5, L_6 and L_7 for maximum signal strength. Now disable the r.f. amplifier stage by disconnecting the 10,000-ohm resistor from L_5, or by removing the heater lead from Pin 12 of the socket. Adjust L_4 for *minimum* signal. Replace the heater or plate voltage and readjust all coils except L_4 for maximum signal again.

The converter should be close to optimum performance if everything has been done properly to this point. If the Channel 2 trap is used, adjust it so that no interference is heard from the local TV station. If the station is very near by, it may still be heard as long as the cover is off the converter case. It should dis-

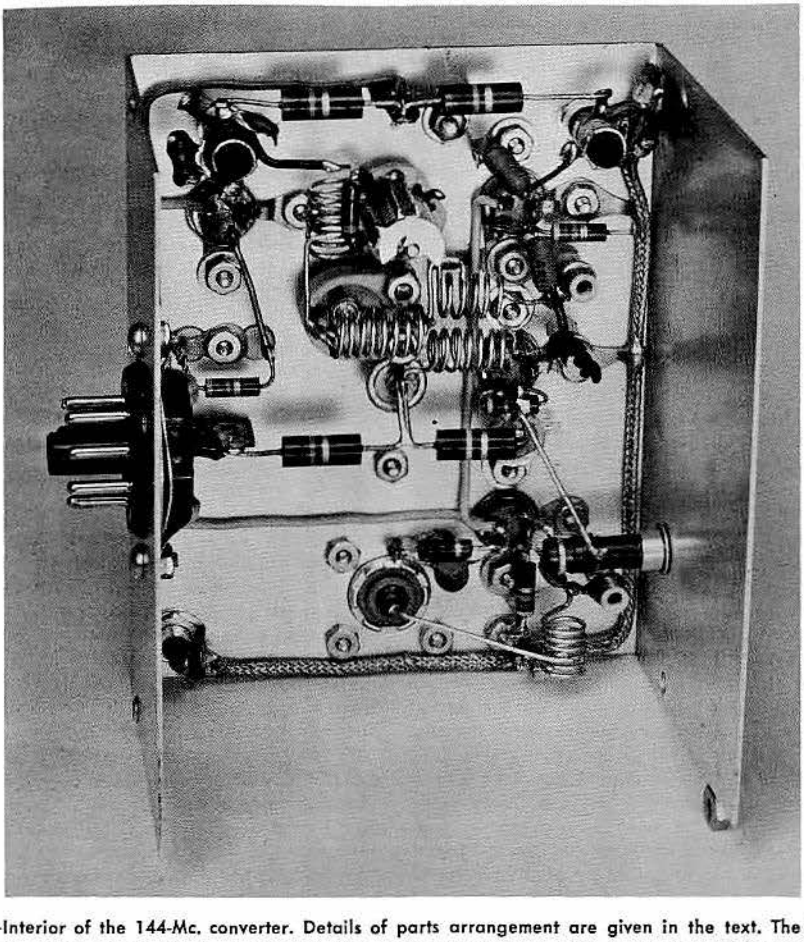


Fig. 4-8—Interior of the 144-Mc. converter. Details of parts arrangement are given in the text. The i.f. output from the mixer plate coil, L_5, is brought through a shielded lead from the upper right corner, down the side of the picture and across the bottom, to the output connector, J_2, at the lower left.

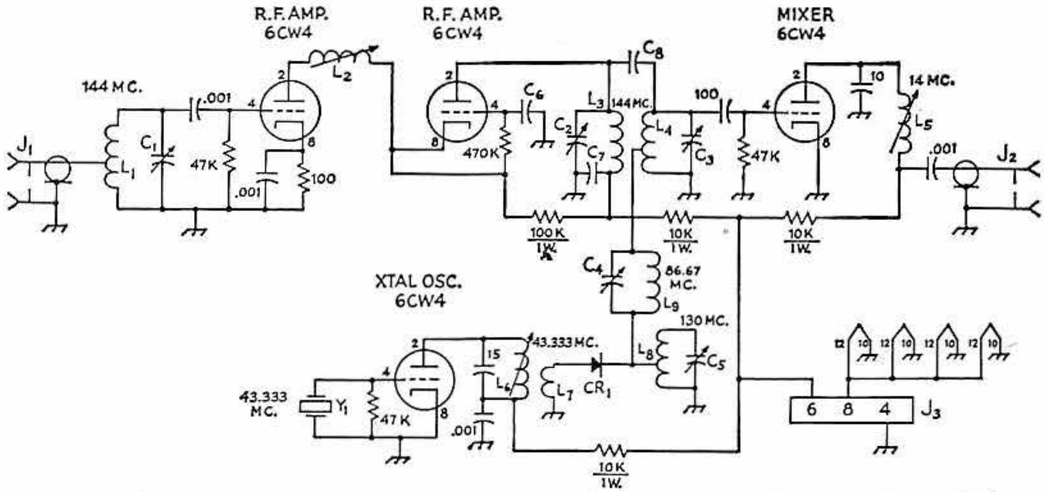


Fig. 4-9—Schematic diagram and parts information for the 144-Mc. converter. Resistors ½ watt unless specified. Fixed capacitors are ceramic unless specified. Decimal values in μf., others in pf.

C_1, C_2, C_3—1–7.5-pf. ceramic trimmer (Centralab 829-7).

C_4—4–30 pf. ceramic trimmer (Mallory ST-554-N).

C_5—20 pf. miniature variable (Hammarlund MAC-20).

C_6, C_7—0.001- or 0.0005-μf. button-type bypass (Centralab ZA-102 or Erie CB11RD471K). Do not use other wire-lead capacitors for these points.

C_8—No. 22 insulated hookup wires 1¼ inches long, twisted together for approximately 1 inch.

CR_1—Crystal-diode rectifier; 1N82.

J_1—Coaxial connector, SO-239.

J_2—Phono jack.

J_3—8-pin plug (Amphenol 86-RCP8).

L_1, L_8—6 turns No. 18, ¼-inch diam. ½ inch long. Tap at 2½ turns.

L_2—5 turns No. 28 enamel, close-wound on ¼-inch iron-slug form. Range 0.24 to 0.41 μh., set for 0.33 μh. (Miller coil No. 20A337RBI).

L_3—6½ turns No. 18, ¼-inch diam., ⅝ inch long.

L_4—5 turns like L_3, ½ inch long, tapped at 2 turns. L_3 and L_4 are parallel, ⅜ inch apart, c. to c.

L_5—Universal-wound coil, 4.7 to 10 μh., set for 7.9 μh. (Miller coil No. 20A826RBI).

L_6—9 turns No. 28 enamel, close-wound on ¼-inch iron-slug form. Range 0.58 to 1 μh., set fo 0.82 μh. (Miller coil No. 20A827RBI).

L_7—1½ turns insulated hookup wire around L_6.

L_9—8 turns No. 18, ¼-inch diam., ⅝ inch long.

Y_1—43.333-Mc. crystal (International Crystal Mfg. Co. F-605).

appear when the case is assembled. Recheck the adjustment of L_2 and L_3 after final adjustment of the trap. A coaxial filter (see Chapter 12) may be needed in extreme cases.

Further work to improve weak-signal reception should be done with a noise generator, though satisfactory results can be obtained on weak signals if the work is done with care. The aim should be better signal-to-noise ratio, rather than merely greater signal strength. This will not be noticeable with the simple tuner, but it can be achieved with a communications receiver as the i.f. system. Using the receiver S meter, or the audio sound of a weak signal, tune for maximum signal *with respect to noise.*

As a final check, put a 50-ohm resistor across J_1. Observe the noise level. Now remove the resistor and put on an antenna system with 50-ohm feed. If the noise rises appreciably, you are hearing the external noise that limits your v.h.f. reception. The only improvement you can make from here on is to put up a bigger or higher antenna, or move to a quieter location.

144-Mc. Adjustment

Adjustment of the 144-Mc. converter is similar, except that the multiplier tank circuit, L_8C_5, should be adjusted for maximum signal. External noise may not be discernible in quiet locations on 144 Mc., and the antenna check outlined for 50 Mc. may be inconclusive. Adjustment of all r.f. circuits should be made carefully for greatest margin of signal over noise, using weak signals. The minimum-signal method of adjusting the neutralizing coil, L_2, should be followed initially, but readjustment for optimum signal-to-noise ratio (or lowest noise figure, using a noise generator) may produce a worthwhile improvement. Do not use the second-harmonic trap, L_9C_4, unless it is necessary to eliminate f.m. interference, as this circuit introduces one more variable to complicate the adjustment procedure.

In most areas 2-meter activity is spread over more of the band than is the case with 50 Mc. The converter response can be made uniform across most or all of the band by tuning the i.f. output coil, L_5, for maximum response near the high end or middle of the band. This coil affects only the gain of the converter; detuning it does not reduce the signal-to-noise ratio. The r.f. amplifier plate and mixer grid circuits, C_2-L_3 and C_3-L_4 have only a minor effect on noise

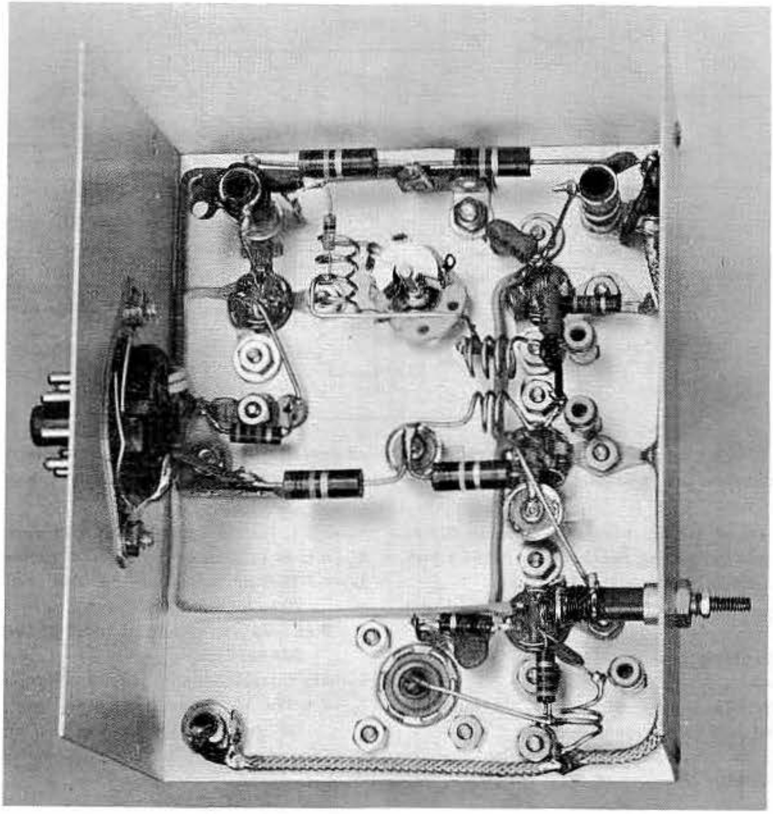


Fig. 4-10—The interior of the 220-Mc. converter is very similar to that of the 144-Mc. model, except for the smaller coils and the elimination of the trap for f.m. interference.

figure, so they can also be "stagger-tuned" to some extent to achieve uniform response.

A fair final check on the 144-Mc. converter performance is to detune the diode multiplier circuit, L_8C_5, and note its effect on the signal-to-noise ratio. If the r.f. amplifier is working properly it should be possible to detune this circuit so that the gain drops an S unit or two, before there is any effect on the signal-to-noise ratio observable on weak signals.

The 220-Mc. Converter

The 220-Mc. converter, Figs. 4-10 and 4-11, is similar to the 144-Mc. converter in both construction and circuitry. A diode frequency quadrupler is used to furnish a 206-Mc. local-oscillator signal from a 51.5-Mc. crystal oscillator. Two tuned circuits are used between r.f. stage and mixer, coupled by a small capacitance. Because the 220-Mc. band is 5 Mc. wide, the receiver following this converter must tune from 14 to 19 Mc. if the entire band is to be covered.

The inductors L_1, L_3, L_4 and L_8 are first wound on a ¼-inch diameter rod or drill and then spaced to meet the specifications. They are supported by soldering the ends directly to tube pins, ground lugs or capacitor terminals.

The adjustment of the converter is quite similar to that of the 144-Mc. converter, and the instructions given earlier apply equally as well to the 220-Mc. band. Depending upon the local operating habits, it may be desirable to peak the circuits for a particular portion of the band. In areas where TV sets are tuned to Channel 7, there may be substantial TV-receiver local-oscillator radiation that will mess up the first megacycle or two of the band, and consequently the amateur activity will peak around 222 or 223 Mc. Both a grid-dip oscillator or signal generator, and a noise generator will be found to be very useful in getting best results from the converter.

If a good noise-figure measuring setup is available the converters should show up about as follows, when adjusted correctly: 50 Mc.—3 db. (far lower than is needed, in view of external noise problems on this band); 144 Mc. –3.5 to 4 db.; 220 Mc.–4.5 to 5 db.

432-Mc. Converter

The crystal-controlled converter for 432 to 436 Mc. shown in Figs. 4-12 and 4-13 uses two grounded-grid r.f. stages and a grounded-grid mixer. This proved to be a more stable arrangement at this frequency, and easier to dupli-

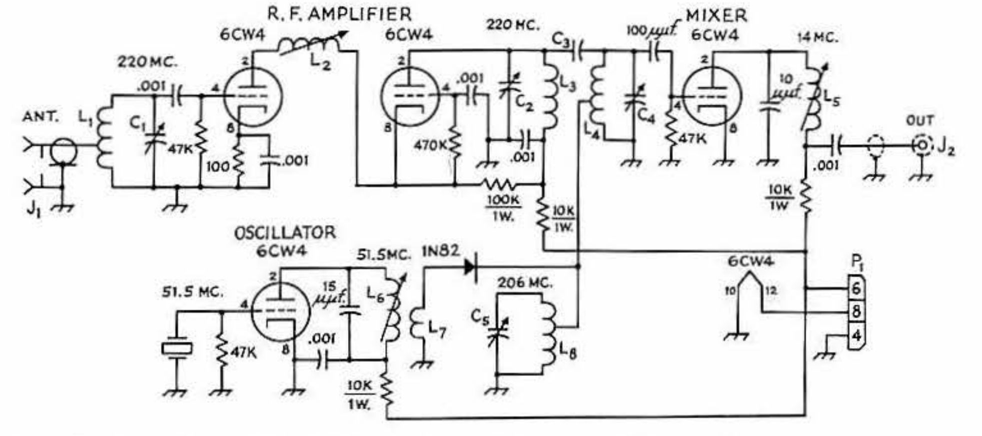


Fig. 4-11—Circuit diagram of the 220-Mc. crystal-controlled converter. Unless specified otherwise, resistors are ½ watt, resistances are in ohms. Fixed capacitors are ceramic unless specified. Decimal values in μf., others in pf.

C_1, C_2, C_4—1-6 pf. tubular trimmer (Centralab 829-6).
C_3—2 pf., made by twisting two insulated wires 1 inch.
C_5—15-pf. variable (Hammarlund MAC-15).
C_6, C_7—.001-μf. button-mica. Disk ceramics not suitable.
J_1—Chassis-mounting coaxial receptacle (SO-239).
J_2—Phono jack.
L_1—2¼ t. No. 18 spaced wire diam., ¼ inch i.d., tapped ¾ t. from ground end.
L_2—0.12-0.19 μh. adjustable inductor (Miller 20A157RBI).
L_3—2¾ t. No. 18 spaced twice wire diam., ¼ inch i.d.
L_4—4 t. as L_3, tapped 1 turn from ground end.
L_5—4.7-10.0 μh. adjustable inductor (Miller 20A826RBI).
L_6—0.43-0.85 μh. adjustable inductor (Miller 20A687RBI).
L_7—1½ t. insulated wire wound on ground end of L_6.
L_8—4 t. No. 18 spaced three times wire diam., ¼ inch i.d., tapped 1¼ t. from ground end.
P_1—Chassis-mounting octal plug (Amphenol 86-RCP8).

cate than one with grounded-cathode stages. A major source of over-all feed-back is the heater connections, and more elaborate heater-line filtering will be found in this unit than in the companion units for lower bands. The local-oscillator signal at 418 Mc. is obtained by tripling twice from a 46.44-Mc. crystal oscillator, once in a triode section of a 6J6 and once through a 1N82 diode.

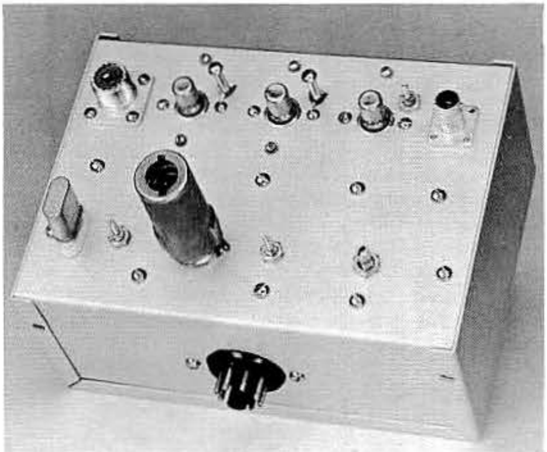


Fig. 4-12—The 432-Mc. converter is built in a 5 x 7 x 3-inch Minibox. At the top in this view, from left to right: input jack, r.f. amplifier, r.f. amplifier, mixer, output jack. The tube (shielded) is a 6J6, used as crystal oscillator and frequency multiplier.

Referring to the circuit diagram in Fig. 4-14, the circuitry through the oscillator-multiplier chain is similar to that shown for the lower-frequency converters described earlier in this chapter. Adjustable inductors tune the circuits on the lower frequencies, and at 418 Mc. a variable capacitor, C_3, tunes a half-turn coil, L_7 (see Figs. 4-13 and 4-15). In the signal channel, 6CW4s are used as grounded-grid amplifiers and mixer, and the coupling circuits are ceramic trimmers and half-turn inductors. The B+ leads are filtered heavily as are the heater leads. For over-all good stability, numerous ceramic feed-through and button bypass capacitors are used.

Study of Figs. 4-12 and 4-13 will give a good idea of the location of the various components. The shield partition is built from a 3¼-inch wide strip of aluminum, and a ⅝-inch lip is bent on one edge for mounting on the chassis. The line of screws holding the shield bisects the chassis.

Practically all of the components will be supported by tube-socket pins, button or feed-through capacitors, ground lugs or tie points. The exception is the diode multiplier, which is supported at one end by the point on L_7 to which it is soldered and at the other by an end of L_8.

A regulated 105-volt power supply is recom-

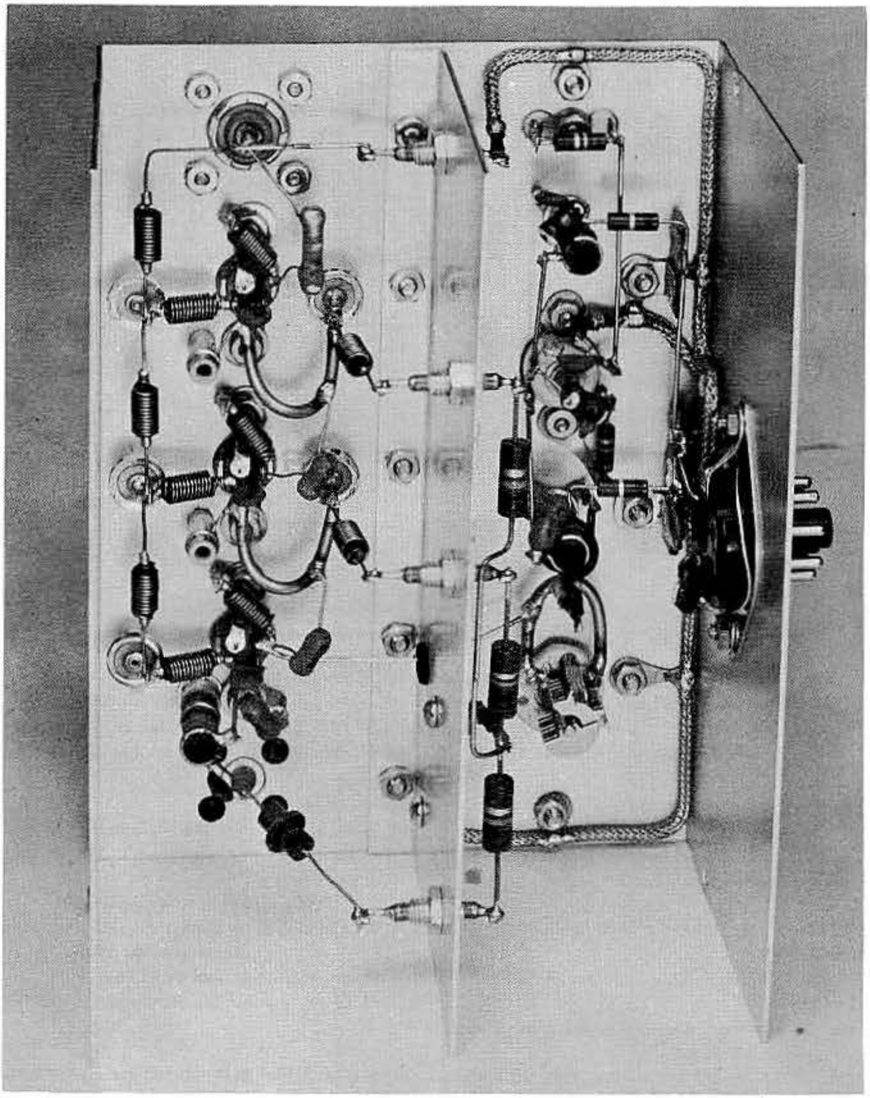


Fig. 4-13—View underneath the 432-Mc. converter shows the partition separating the oscillator and multiplier (right) from the r.f. and mixer. R.f. chokes mounted on button bypass capacitors (left) provide filtering for heater wiring. Feedthrough bypass capacitors in partition are 2 inches from chassis, except heater feedthrough (top) which is ½ inch from chassis. Oscillator injection lead from L_7 to the cathode of the mixer runs through rubber grommet in partition.

mended for use with the converter. The crystal oscillator should be checked first, by measuring the voltage drop across the 1000-ohm resistor to L_4. The drop across this resistor will be greater when the stage is not oscillating than when it is, and the slug of L_4 should be set at a lower-inductance value than that which gives minimum voltage drop across the resistor, to insure proper starting and operation. If a wavemeter is available, the frequency of output should be checked. The slug of L_5 should be set for minimum voltage drop across the 1000-ohm resistor connected to L_5, and again the frequency should be checked with a wavemeter if one is available.

The third harmonic of a 144-Mc. signal source can be used to align the r.f. stages, after which C_3 and L_5 should be peaked for maximum signal.

If any instability is experienced in the signal circuits, as evidenced by regeneration or oscillations at some setting of C_1 and C_2, look for poor connections or poor grounds. In some cases the instability may also be caused by having the mixer tap too high above ground on L_7.

Using Other Intermediate Frequencies

The Nuvistor converters just described were designed for use with receivers capable of tun-

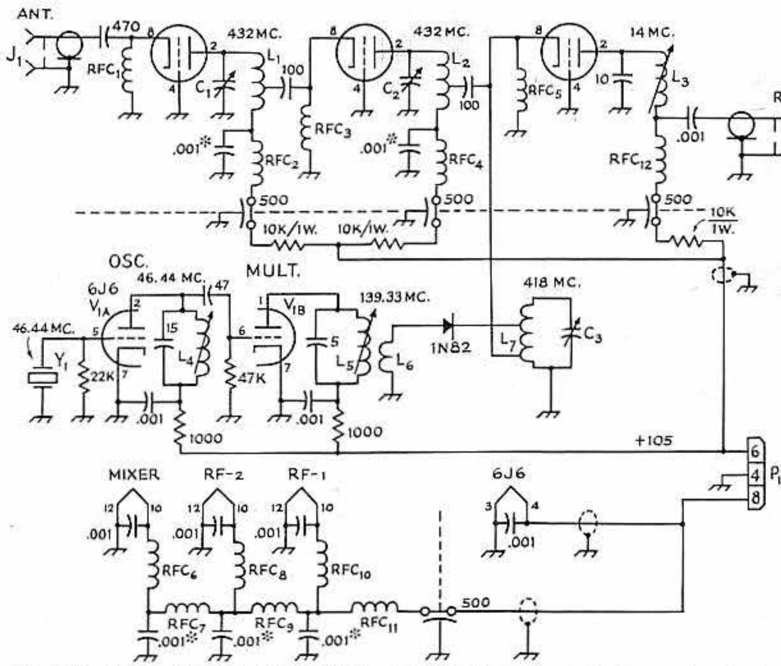


Fig. 4-14—Wiring diagram of the 432-Mc. converter. Capacitance values are in pf., except decimal values which are in μf. Resistors are ½ watt unless specified otherwise. All 0.001-μf. capacitors marked * are button-type (Centralab ZA-102, Erie CB11RD471K or equiv.); other 0.001-μf. capacitors are disk ceramic. All 500-pf. feedthrough capacitors are Centralab FT-500.

C_1, C_2—6-pf. ceramic variable (Centralab 829-6).
C_3—15-pf. variable (Hammarlund MAC-15).
L_1, L_2, L_7—See Fig. 4-15.
L_3—4.7—10.0-μh. adjustable inductor (Miller 20A826RBI).
L_4—0.43—0.85-μh. adjustable inductor (Miller 20A687RBI).
L_5—0.119—0.187-μh. adjustable inductor (Miller 20A157RBI).
L_6—2 t. No. 20 insulated, wound around ground end of L_5.
RFC_1—RFC_{11}—8 inches No. 22 enam. cleaned ½ inch each end and wound on 10K or higher ½-watt resistor.
RFC_{12}—50-μh. r.f. choke (Millen 34300-50).
J_1—Coaxial chassis receptacle (SO-239).
J_2—Coaxial chassis receptacle (UG-290A/U).
P_1—Octal plug (Amphenol *86-RCP8).*
Y_1—46.44-Mc. crystal (International Crystal F-605).

ing 14 to 18 Mc. (14 to 19 for the 220-Mc. model.) While this is by far the most popular i.f. range for v.h.f. converter service, it is not always usable or desirable. The amateur-bands-only receiver has only 500 kilocycles tuning range, at the most, beginning at 14 Mc., and often less. Some low-priced receivers and many older but otherwise desirable models work better at 7 Mc. than at 14. Finally, several receivers have tuning ranges designed especially for v.h.f. converter work, some beginning at 22 Mc. and others at 30.5 Mc. These include the Mohawk, NC-300, NC-303 and SX-101.

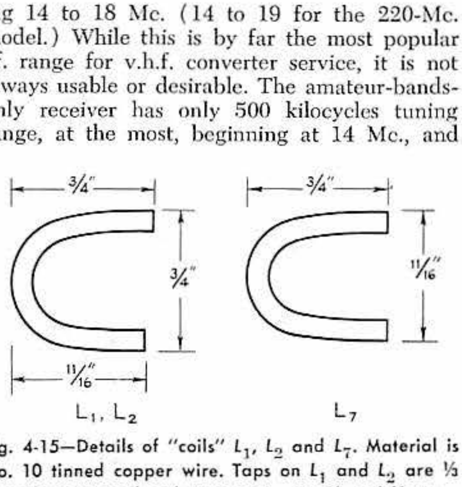


Fig. 4-15—Details of "coils" L_1, L_2 and L_7. Material is No. 10 tinned copper wire. Taps on L_1 and L_2 are ⅓ up from ground end; taps on L_7 are ⅓ and ½ up from ground end.

The best solution with the amateur-band receiver is usually to use the 10-meter range, and more than one crystal, if full band coverage is wanted on the v.h.f. bands. But even when this is done the actual starting frequency will vary from one type of receiver to another, as the low end of the 10-meter range may be 26, 27 or 28 Mc., or some frequency in between these. With small changes in i.f. range the only important consideration is the converter crystal. With a 50-Mc. converter the crystal frequency is 50 minus the i.f., or 24, 23 or 22 Mc. for the three examples mentioned just above. For a 144-Mc. converter the *injection* frequency is arrived at the same way. It would be 118, 117

TABLE 4-I
Crystal frequencies recommended for use with common intermediate frequencies in v.h.f. converters.

Band,		*Crystal frequency for i.f. beginning at*				
Mc.	*7 Mc.*	*22 Mc.*	*26 Mc.*	*27 Mc.*	*28 Mc.*	*30.5 Mc.*
50	43.0	28.0	24.0	23.0	22.0	19.5
144	45.667	40.667	39.333	39.0	38.667	37.833
220	53.25	49.5	48.5	48.25	48.0	47.375
432	——	45.555	45.111	45.0	44.9	44.611

Injection frequency is 3 times the crystal frequency for 144-Mc. converters, 4 times for 220 Mc. and 9 times for 432. Many other combinations can be used. An i.f. less than 14 Mc. is not recommended for 432 Mc.

or 116 Mc. for these same three examples. The crystal is some fraction of this frequency, those in Table 4-I being merely typical examples for the common intermediate-frequency ranges. Table 4-II gives the nominal inductance of the slug-tuned coils needed with these frequencies.

Tuning of the mixer plate circuit is not critical. If the circuit will not peak with the coil value selected, the capacitance connected from the mixer plate to ground can be varied to bring the resonant frequency into the desired range.

TABLE 4-II
Nominal inductance of slug-tuned coils for frequencies listed in Table I. In the multiplier circuits the desired frequency ranges can be covered by adjustment of the associated variable capacitor, and by adjusting turn spacing of air-wound coils. Use number of turns specified for 14-Mc. i.f.

Freq., Mc.	*Nom. L, μh.*	*Freq. Mc.*	*Nom. L, μh.*
7-11	27	36-40	1
19.5	4.5	40-45	0.8
22-26	3	45-50	0.7
26-28	2.5	50-55	0.6
30.5	1.5	136-139	0.15

BANDSWITCHING CONVERTER USING A TV TUNER

Tuners used in home TV receivers are, in effect, multiband v.h.f. converters, so where the tuner is made as a removable subassembly it is a relatively simple matter to convert the tuned circuits for the amateur v.h.f. bands. A conversion of the familiar Standard Coil Tuner for use on 28, 50, 144 and 220 Mc. is described below. The tuner is an Admiral type 94C18-4. Two earlier models, the 94C8-2 and 94A8-2 may be used similarly. Similar tuners are available under many other numbers, as the same basic tuner was used by several large receiver manufacturers.

The tuner has a pentode r.f. amplifier and a 6J6 mixer-oscillator, and the i.f. output frequency is roughly 20 to 24 Mc. Many other tube lineups are found in TV tuners, and mechanical and electrical designs abound in infinite variety. The information given is correct only for the tuners mentioned above. We do not have conversion data for any other tuner by this or other manufacturers, though the following pro-

TABLE 4-III
Coil Information for the Converted TV Tuner

	28 Mc.	*50 Mc.*	*144 Mc.*	*220 Mc.*
Antenna, L_{101A}	20 t. No. 26 d.s.c. center-tapped. Wind over L_{101B}.	10 t. No. 22 e. close-wound, or Channel 2 with no change.	4 t. No. 26 d.s.c. interwound in L_{101B}, center-tapped, or Channel 7 with no change.	2 t. No. 22 e., center-tapped. Mount at center of L_{101B}.
R.F. Grid, L_{101B}	No. 30 d.s.c., 1³⁄₁₆ in. long.	28 t. No. 26 d.s.c. or add 4 t. to each end of Channel 2.	8 t. No. 22 e., ⅝ in. long, or add 1 t. each end of Channel 7.	4 t. No. 22 e., ³⁄₁₆-inch dia., 1-inch long.
R.F. Plate, L_{102A}	No. 32 e., ¹³⁄₃₂ inch long.	19t. No. 26 d.s.c. or add 2 turns to outside end of Channel 2.	5 t. No. 22 e., ¼ inch long, or add 1 turn to Channel 7.	2 t. No. 22 e. sp. 1 dia., or remove 1 turn from Channel 13.
Mixer Grid, L_{102B}	No. 32 e., ¹¹⁄₃₂ inch long.	15 t. No. 26 d.s.c. or add 2 turns to Channel 2.	4 t. No. 22 e., ¼ inch long, or add 1 turn to Channel 7.	2 t. ¹⁄₁₆-inch copper strip, ⅛ inch apart, or spread turns of Channel 13.
Osc., L_{102C}	No. 32 e., ¼ inch long, with slug.	14 t. No. 26 d.s.c. or add 2 turns to Channel 2.	5 t. No. 22 e., ⁵⁄₁₆ inch long, or add 1 turn to Channel 7.	2¼ t. No. 22 e., sp. 1 dia., or remove ¾ turn from Channel 13, with slug.

cedure should contain useful ideas for anyone wishing to attempt the job with other models. Many Standard Coil Tuners are available with dual-triode front end designs. These should give slightly better noise figure than the pentode model shown, but coil values must be found by experiment.

Mounting the Turret

The cylindrical turret has individually removable coils. With the exception of those for 28 Mc., all the amateur-band coils, Table 4-III were made by modifying existing coils. Inexpensive replacement units are available from TV distributors, so the constructor can experiment without fear. The tuner schematic, as supplied by the manufacturer is given in Fig. 4-17. The maker's parts identifications are shown, and will be used throughout this description. Layout, as viewed from the top, is given in Fig. 4-20. It is well to become familiar with the unit from these drawings, and from careful inspection, before operating on it. In fact, it is a good idea to set it up for actual operation and try it out in reception of TV signals, in order to be completely familiar with it, and to know that it is in working order as received.

The converter was assembled on a 5 by 7 by 2-inch chassis, equipped with a VR tube and socket to drop any available voltage to 105, regulated, for the converter tubes. An i.f. output coaxial connector is mounted on the rear wall. A crystal socket for 300-ohm balanced input is mounted on a small bracket fastened to the same wall. A small aluminum panel was provided. Any large knob can be modified for the fine tuning. It should be drilled to pass the bandswitch, whose shaft is coaxial with the tuning sleeve. A smaller knob is affixed to the switch shaft.

A tuner with 21-Mc. output (the old-type i.f.) is preferable to late-model types, as it can be used directly with any communications receiver tuning 20 to 24 Mc. If the receiver tunes only up to 18 Mc., the i.f. winding, L_{103}, can be padded down to about 16 Mc. This is desirable in any case, particularly if 28-Mc. reception is to be included, as it will help prevent mixer troubles on this band. The padder should be about 30 pf., connected in parallel with C_{112}.

The fine tuning will not cover an entire band, so the converter may be used in the same manner as a crystal-controlled one: by leaving

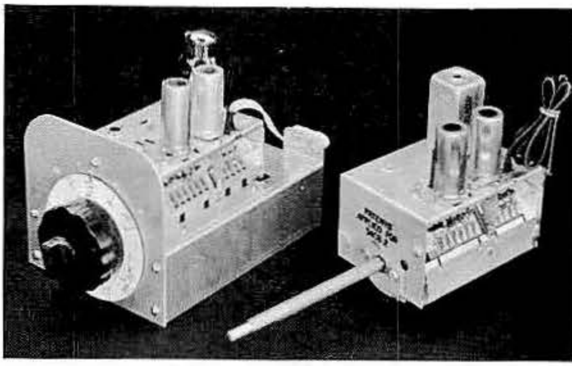


Fig. 4-16—A 4-band v.h.f. converter made from a Standard TV Tuner. Earlier model of the tuner is shown at the right. Eight TV channels may be left in use, for harmonic-checking purposes.

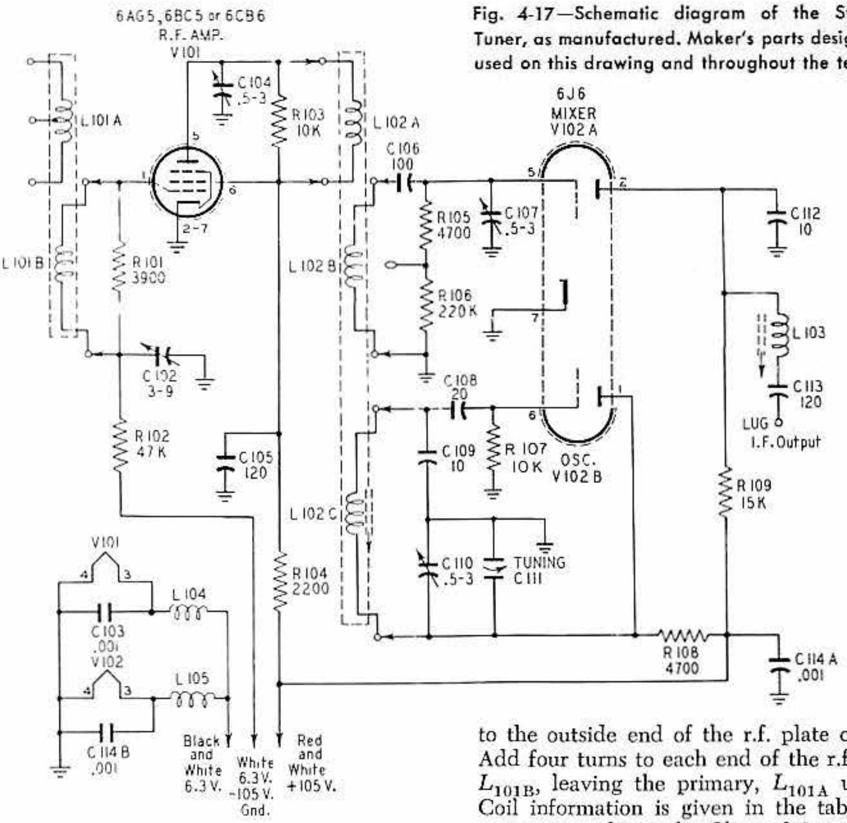


Fig. 4-17—Schematic diagram of the Standard TV Tuner, as manufactured. Maker's parts designations are used on this drawing and throughout the text.

the fine tuning set in one position and doing all tuning with the communications receiver. If only a narrow tuning range is wanted, the fine tuning serves very well. Coverage is roughly 600 kc. on 28 Mc., 900 on 50, 2 Mc. on 144 and 4 Mc. on 220. To go higher, shift the i.f. *lower*.

Adjustment and Use

Apply power and tune the receiver for maximum noise. The exact intermediate frequency is unimportant, as the oscillator range can be shifted as needed, by changing the oscillator coil inductance. The actual tuning range of the oscillator with the fine tuning will vary with the setting of the oscillator padder, C_{110}. Remember that this trimmer and all the others, C_{102}, C_{104} and C_{107}, affect all tuner ranges, so coils must be adjusted in inductance so that the setting of the four tuner trimmers is the same for each coil range.

The 50-Mc. range is easy if the Channel 2 coils are modified for this band. The oscillator winding, L_{102C}, can be used by removing the brass slug and adding two turns to the inside end of the winding. To use the brass slug for inductance adjustment, use at least four turns. It is recommended that the slug be removed, and turns spread or closed together for inductance adjustment. Add two turns to the adjacent end of the mixer grid winding, L_{102B}, and to the outside end of the r.f. plate coil, L_{102A}. Add four turns to each end of the r.f. grid coil, L_{101B}, leaving the primary, L_{101A} unchanged. Coil information is given in the table, in case you want to leave the Channel 2 coils unmodified.

To set up for reception on 50 Mc., put the fine tuning near its maximum-capacity setting, and peak the i.f. coil, L_{103} or its padder mentioned earlier, for maximum noise. Peak trim-

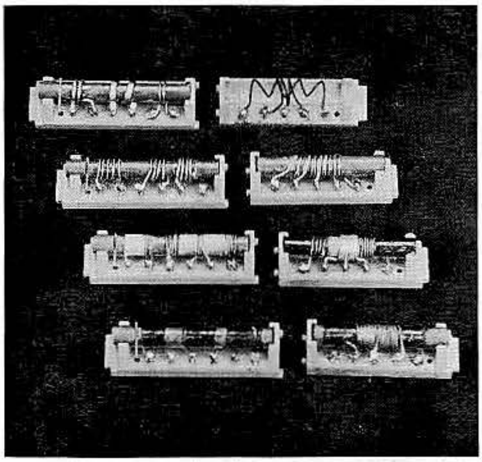


Fig. 4-18—Amateur band coils for the converted TV tuner. Each assembly at the left has three windings: oscillator, mixer grid and r.f. plate, in that order. The two-winding coils, right, are the r.f. grid and antenna-coupling coils. In order of frequency, reading up from the bottom, they are for 28, 50, 144 and 220 Mc.

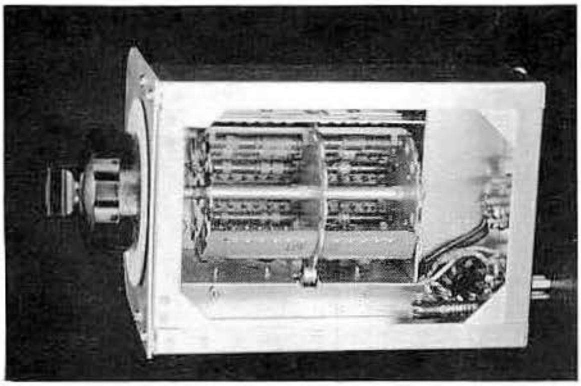


Fig. 4-19—Bottom view of the converter, showing the tuner with several sets of coils removed. Each channel has separate assemblies, which may be removed readily.

mers C_{107}, $_{104}$ and $_{102}$ for maximum noise, and you're ready to go. If any of these padders peaks near maximum or minimum setting, it is well to adjust the coil inductance to bring them near the middle of their range, as they will need to be that way for the other bands, too. Checking with them is handy in the adjustment phase, however, to indicate whether coil inductance must be raised or lowered. The oscillator coil should be set so that 50 Mc. comes near the low end of the tuning range (maximum setting of the fine-tuning capacitor).

Procedure for other bands is similar. Remember, the objective is to get all coils to peak with the tuner trimmers at the same setting. If you're interested in only one band this is, of course, unimportant. If you find that response varies across a band, the coil inductances can be staggered to level off the peaks. Again, the tuner trimmers will show which way to change the inductance.

If a signal generator is available the process is simpler, but peaking on noise and checking coverage by means of signals is not difficult. If you have a v.f.o. or a good supply of crystals the tuning range can be checked readily enough. Performance is good, but not quite the best obtainable with modern crystal-controlled converters. The oscillator stability is far from perfect, especially on the high bands, though it is good enough for voice reception with good quality. With a receiver of relatively broad i.f. response, such as the surplus BC-312 and 342, the stability is adequate on all ranges.

Sensitivity on 28 or 50 Mc. is as good as you can ever use. Less than 0.5 microvolt input gives a 10-db. signal-to-noise ratio. About 1 microvolt is required on 144 and 2 on 220. A low-noise preamplifier will help on these two bands, but one will not be needed for optimum weak-signal reception on 50 or 28.

The converter is also useful in hunting down harmonics, or in listening for TV signals as indications of band conditions. Several TV channel coils can be left in for these purposes. Video signals make a fearful racket, but the f.m. audio can be understood fairly well by slope tuning, with all but the most selective receivers.

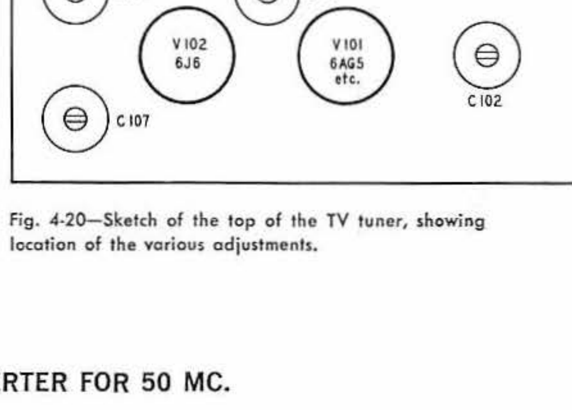


Fig. 4-20—Sketch of the top of the TV tuner, showing location of the various adjustments.

SIMPLE FET CONVERTER FOR 50 MC.

The FET (field-effect transistor) provides freedom from overloading comparable to the better vaccuum-tube r.f. amplifiers. Many inexpensive FETs are now available that give excellent noise figure and gain in the v.h.f. range. The FET is quite similar in characteristics to a tube amplifier, and it will be seen from a comparison of the schematic diagrams that the simple 50-Mc. converter of Fig. 4-21 is not unlike the Nuvistor converter previously described. It should be its equal in performance, and it has a considerable edge in overall simplicity, since it can operate directly from a 9- or 12-volt battery.

Circuit Details

In the interest of bandpass response and rejection of out-of-band signals, double-tuned circuits are used for the input and interstage coupling. Back-to-back diodes, CR_1 and CR_2, between the antenna connection and ground, protect the first stage from r.f. leakage from the transmitter, or from transient voltages such as might result from nearby lightning discharges.

The r.f. and mixer transistors may be any of several v.h.f. FETs. The inexpensive MPF-102 shown is by Motorola, and similar types are available from other sources. Any type rated for r.f. amplifier service to 100 Mc. or so should be more

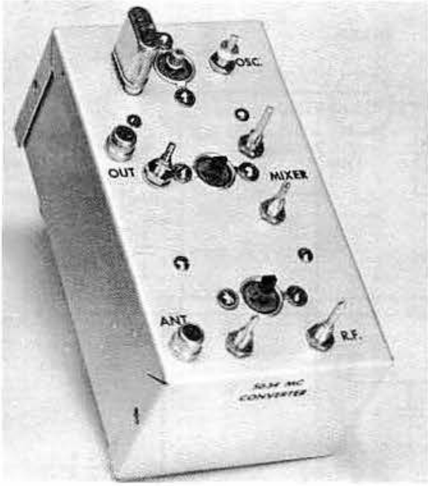


Fig. 4-21—50-Mc. converter with field effect transistors in the r.f. and mixer stages.

than adequate. The r.f. transistor is operated grounded-gate, which is the equivalent of grounded-grid use of a vaccuum tube. Such a stage is relatively low-gain, but it is stable, and does not require neutralization. At 50 Mc., at least, the gain is adequate for good reception, when the stage is used with a grounded-source mixer, as shown.

The oscillator uses a bipolar transistor with a 43-Mc. crystal. Any v.h.f. transistor will do here. The mixer output is 7 to 11 Mc., as shown, but the 14-Mc. range could be used equally well. In that case the coil and crystal information could be taken from the converter of Fig. 4-6.

Construction

The converter case is a 3 by 5¼ by 2¼-inch Minibox, with the parts mounted on the cover portion. Shields are mounted across the chassis to keep down unwanted interstage coupling. Parts layout is not particularly critical, though the approximate relative positions of the principal r.f. components should be followed, for best results. Phono connectors were used for the input and output fittings, J_1 and J_2. These work well enough, though you may prefer the better mechanical qualities of BNC or other coaxial fittings.

Colored tip jacks J_3 and J_4 on the rear wall are used to bring in the operating voltage. Small feedthrough bushings (Johnson Rib-Loc) are mounted in the interstage shields, where leads must run between sections. In Fig. 4-22 the r.f. amplifier stage is at the right side of the picture, the mixer at the center, and the oscillator at the left.

Adjustment

Check the wiring, to be sure that it is correct and complete, then connect the converter to the communications receiver input, and apply d.c. voltage through J_3 and J_4. The receiver noise

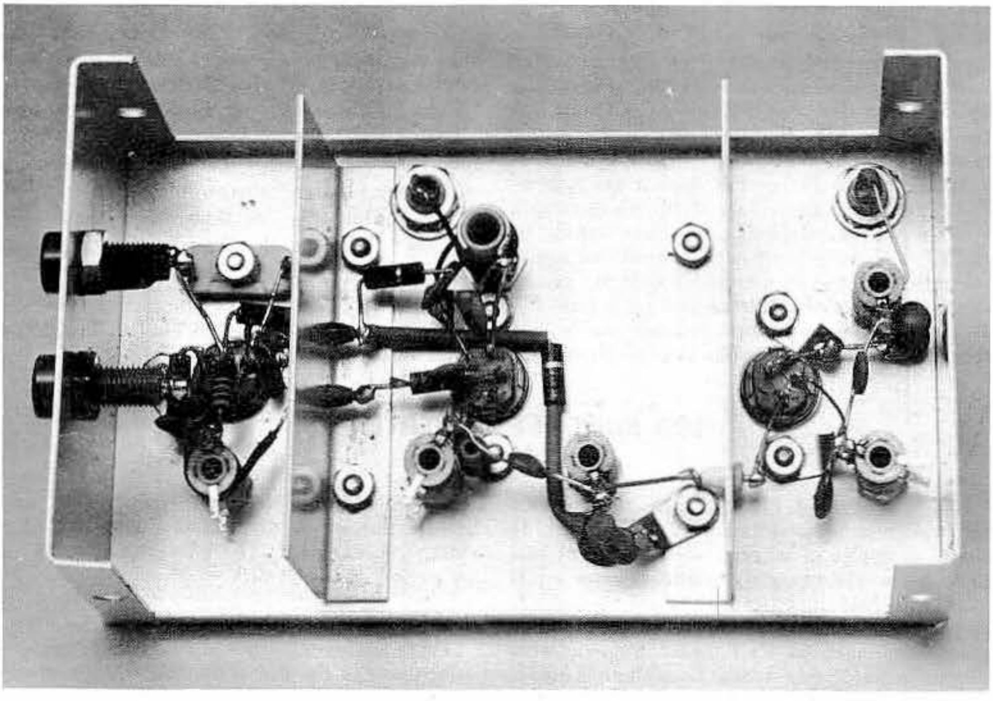


Fig. 4-22—Looking into the under side of the 6-meter converter the mixer is in the center, with the r.f. stage at the right and the oscillator at the left.

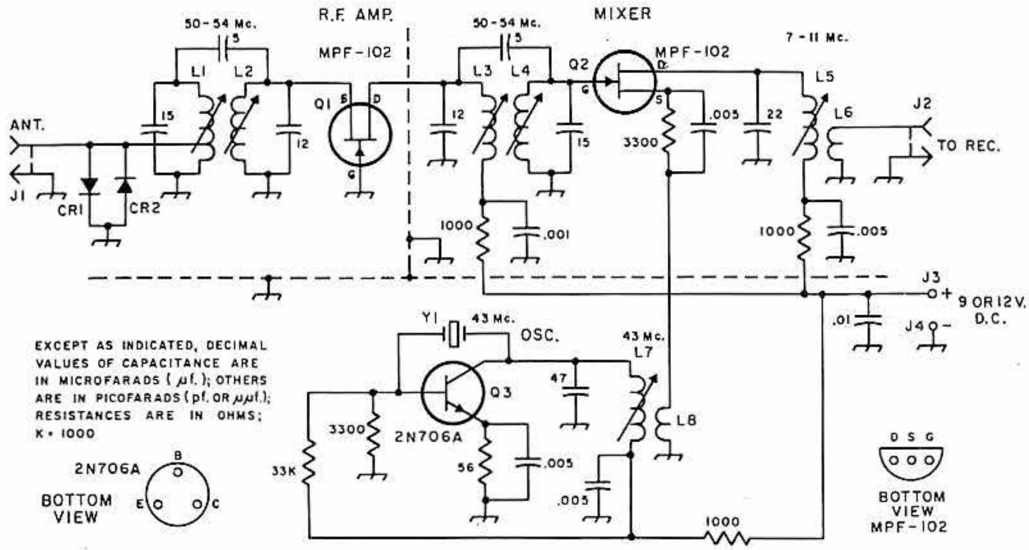


Fig. 4-23—Schematic of the 6-meter FET converter. All resistors are ½-watt composition. All capacitors are disk or tubular ceramic.

CR_1, CR_2—Germanium diode (1N34A suitable).
J_1, J_2—Phono connector.
J_3, J_4—Insulated banana jack, one red, one black.
L_1, L_4, inc.—0.68 μh., slug-tuned (Millen 69054-0.68*). L_1 has tap added at 2nd turn from ground end.
L_5—11 to 24 μh. slug-tuned (Miller 4507).
L_6—5 turns insulated wire over cold end of L_5.
L_7—0.33 μh., slug-tuned (Millen 69054-0.33*).
L_8—1 turn small-gauge insulated wire over cold end of L_7.
Y_1—43.0-Mc. third-overtone crystal (International Crystal Co. Type F-605).
*Available directly from James Millen Mfg. Co. 150 Exchange Street, Malden, Mass.

level will increase markedly, if the converter oscillator is working. Turning the slug in L_7 should bring this about, if it does not start immediately. Set the slug at a point where oscillation will occur each time the voltage is applied.

It should now be possible to hear any reasonably strong signal, actually on-the-air or from a v.h.f. signal generator. Peak all core studs in the r.f. and mixer circuits for maximum signal strength, and the adjustment should be nearly complete. It may be helpful to stagger-tune L_3, L_4 and L_5 for uniform response across the desired frequency range. This results in somewhat lower than maximum gain, but does not affect the noise figure adversely, as this is determined mainly by the first tuned circuit. L_1 and L_2 should be adjusted carefully for best signal-to-noise ratio on a weak signal, rather than for maximum gain, if there is a difference discernible.

In adjusting the coil slugs, be sure that the circuits actually peak. Occasionally there will appear to be a peak which is actually the centering of the slug in the winding. If this happens, you need more turns in the coil or more capacitance across it.

LOW-NOISE FET CONVERTER FOR 144 MC.

The converter of Fig. 4-24 was designed to provide optimum performance and flexibility. It shows examples of several techniques that can be used to advantage by constructors of v.h.f. gear. Two field-effect transistors are used as r.f. amplifiers in a transistor version of the familiar cascode circuit. An FET mixer follows, with output in the 28-Mc. band. Then there is an i.f. amplifier employing an integrated circuit, a convenient way of handling this job, since all the small parts needed are included in the IC itself. The oscillator-doubler system uses inexpensive bipolar transistors, and the oscillator collector voltage is regulated. The converter is built on an etched circuit board.

Options

There are several types of FETs that could be substituted in the first three stages, though the inexpensive Motorola MPF-102s are adequate. Maybe your receiver works well enough at 28 Mc. so that you don't need the i.f. amplifier

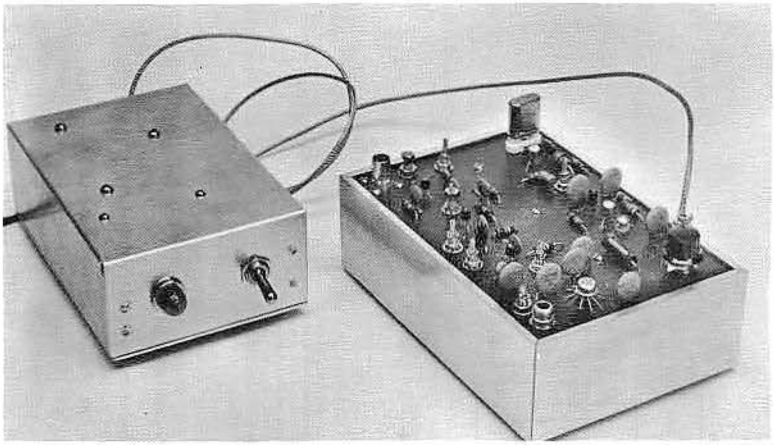


Fig. 4-24—A look at the completed 2-meter etched circuit FET/IC converter and its 12-volt a.c.-operated power supply. The converter is at the right, mounted in a Vector case. The power supply is housed in a 4 x 5 x 2-inch aluminum chassis. It has a bottom plate to which four rubber feet have been attached.

stage, but the gain-control feature makes it worth inclusion, and its input and output circuits can be stagger-tuned, to smooth out the overall response of the converter across the desired bandwidth. Regulation of the oscillator voltage may not mean much to the average operator, but if you want the ultimate in stability, for s.s.b. or weak-signal c.w. work, you'll very likely find it worthwhile. Then there's that polarity-insurance diode, CR_1. If you're sure that you'll always hook up power the right way around, this serves no useful purpose. It's pretty cheap insurance, though, especially if you decide to run the converter from a battery. All-battery operation might be attractive if you happen to have a transistorized receiver to use with the converter.

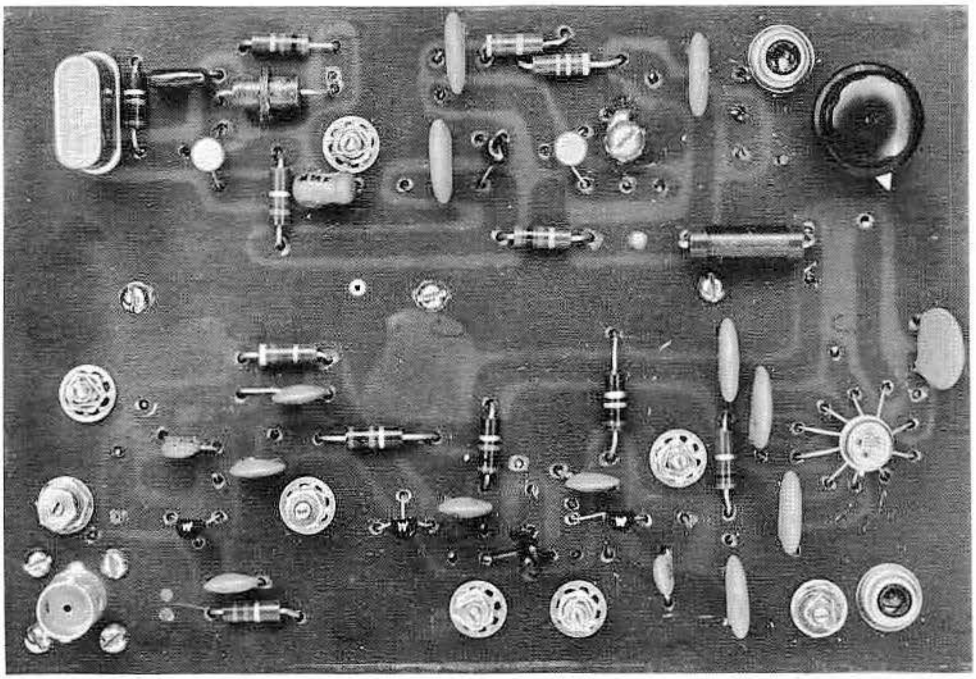


Fig. 4-25—A head-on view of the top surface of the etched circuit board. The i.f. gain control knob is at the upper right. The input jack for the 12-volt supply is just to the left of the gain control. The i.f. output jack is at the lower right, and the r.f. input jack is at the lower left on the board. The IC is located at the far right, just above the i.f. output connector.

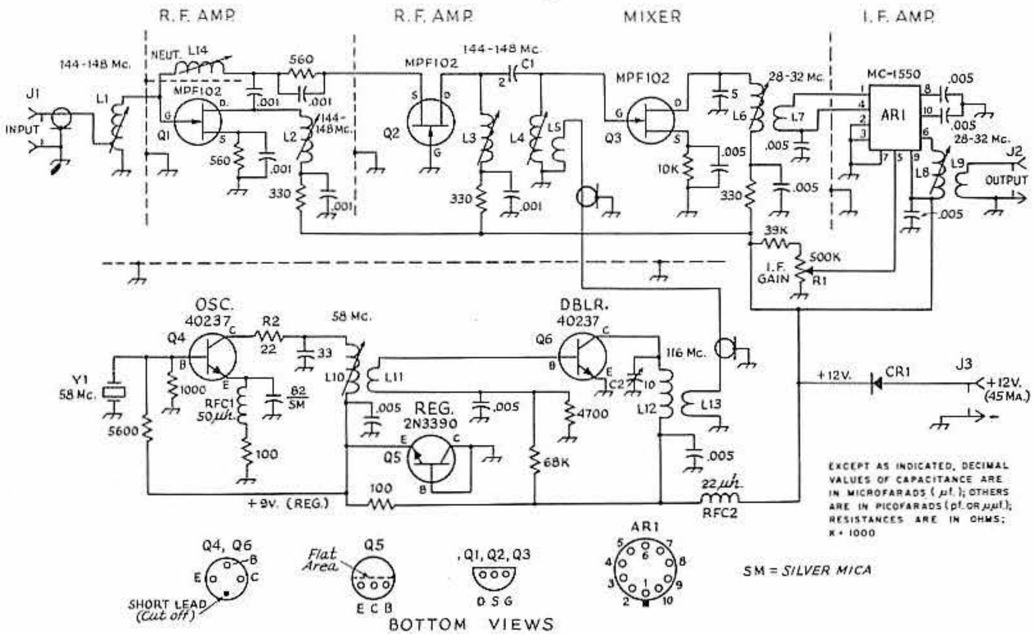


Fig. 4-26—Schematic of the converter. Fixed-value resistors are ½ watt composition. Fixed-value capacitors are disk or tubular ceramic unless stated otherwise.

AR_1—Motorola MC-1550 integrated circuit. Take output from L_7, if amplifier stage is not used.

C_1—Gimmick capacitor: two 1-inch lengths of insulated hookup wire, twisted 6 times. A 2-pf. fixed-value ceramic capacitor can be substituted.

C_2—10-pf. piston-type trimmer (Centralab 829-10).

CR_1—Silicon diode, 50 p.r.v. or greater, at 200 ma.

J_1—BNC-style chassis connector.

J_2, J_3—Phono jack.

L_1—6 turns No. 24 enam., wire to occupy ⅜ inch on slug-tuned form, ¼ in. dia.; (Miller 4500-4) tap 1¼ turns above ground end.

L_2—4 turns No. 24 enam. wire to occupy ⅜ inch on same type form as L_1.

L_3—5 turns No. 24 enam. to occupy ⅜ inch on same type form as L_1.

L_4—4 turns No. 24 enam. to occupy ⅜ inch on same style form as L_1.

L_5—2 turns insulated hookup wire over ground end of L_4.

L_6, L_8—Slug-tuned, 1.6 to 2.8 µh. (Miller No. 4503).

L_7, L_9—Three-turn link over cold ends of L_6 and L_8. Use small-diameter insulated hookup wire.

L_{10}—5 turns No. 24 enam. wire to occupy ⅜ inch on Miller 4500-4 slug-tuned form.

L_{11}—2-turn link of small-dia. wire over cold end of L_{10}.

L_{12}—5 turns No. 20 tinned copper wire (or enam.), 5/16-inch diameter, ⅜ inch long.

L_{13}—2 turns small-dia. insulated hookup wire inserted in cold end of L_{12}, ¼-inch dia.

L_{14}—9 turns No. 24 enam. wire, close-wound on same style form as L_{10}.

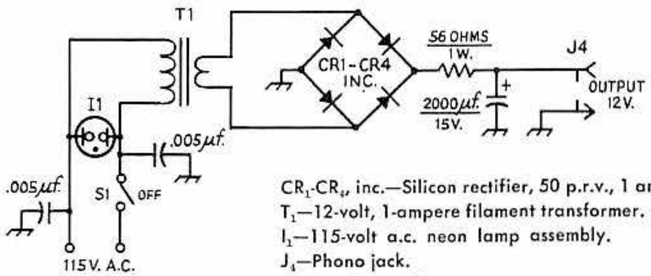
Q_1-Q_6, inc.—For text reference purposes.

R_1—500,000-ohm control, linear taper.

RFC_1—50-µh. r.f. choke (Millen J-300-50).

RFC_2—22-µh. r.f. choke (Millen J-300-22).

Y_1—58-Mc. third-overtone crystal (International Crystal type F-605).



CR_1-CR_4, inc.—Silicon rectifier, 50 p.r.v., 1 ampere.
T_1—12-volt, 1-ampere filament transformer.
I_1—115-volt a.c. neon lamp assembly.
J_4—Phono jack.
S_1—S.p.s.t. toggle.

Fig. 4-27—Schematic of the converter's power supply. The 2000-μf. capacitor is electrolytic, others are disk ceramic, 1000-volt units. The 56-ohm resistor was selected to give the proper power-supply voltage whe used with the circuit of Fig. 4-26 (12 volts d.c.)

Construction

The converter is assembled on a 4½ by 6½-inch etched circuit board (Vector CU65/45-1). A layout drawing, Fig. 4-29, will help you to make your own layout, or a template can be obtained if you wish.° Ready-made boards can be purchased.° Shields of flashing copper isolate the various sections of the converter. Where these are soldered or bolted to the circuit board it is necessary to trim away portions to prevent shorting out the circuits. The shields do not show clearly in Fig. 4-28, so their approximate location can be checked out by the dashed lines in Fig. 4-26. Locations of the key components can be determined from the layout drawing, Fig. 4-29, and from the top and bottom photographs.

The chassis is a matching Vector assembly made of two of their Fram-Loc rails 2 by 6⅝ inches (Vector SR2-6.6.062), two 2 by 4½ inches (SRI-4.6/062), and a bottom cover (PL4566). A standard chassis could be used, if the cover plate is cut out to fit the circuit board.

° Full-size template similar to Fig. 4-29 sent upon receipt of 25 cents and stamped self-addressed envelope. Address ARRL Technical Dept. Newington, Ct. 06111 and mention figure number, publication and edition number.

Readymade boards from Harris Co., 56 East Main St., Torrington, Ct. 06790. Write them for prices.

Power Supply

The converter requires about 12 volts d.c. at 45 ma. The a.c. operated 12-volt d.c. supply for

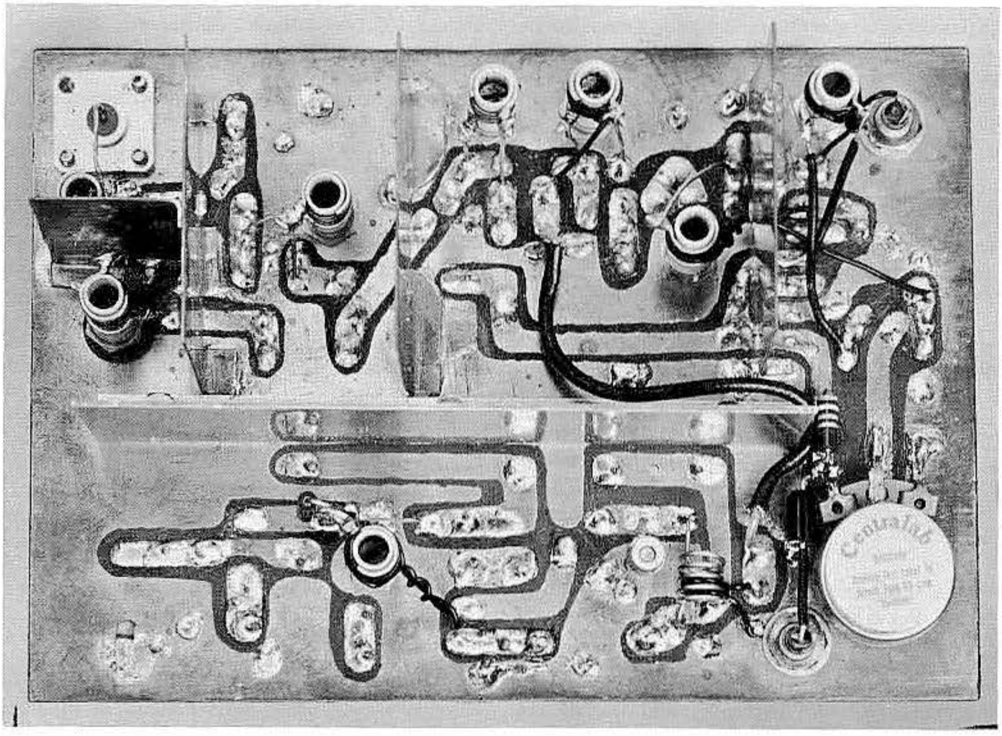


Fig. 4-28—Bottom view of the etched-circuit board converter. The i.f. gain control and 12-volt power jack are at the lower right. The input circuit and r.f. stages are at the upper left. The mixer is at the upper center, and the IC i.f. amplifier is at the upper right. The oscillator chain extends along the lower portion of the board. The interstage shields are in place, but are difficult to see in this photo.

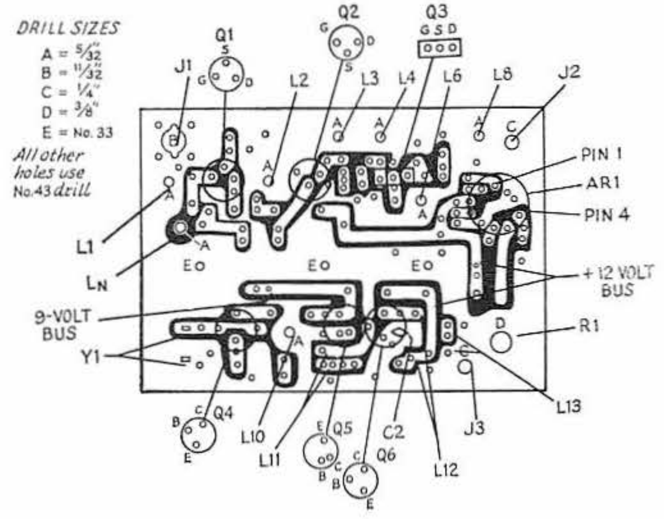


Fig. 4-29—Layout of the etched circuit board. The lines show where the key components are mounted and indicate the way the semiconductor leads are indexed. This is a bottom view of the board (copper side). The inked-in areas represent the sections of the board that have been etched away. The white areas are the copper strips that remain.

fixed-station use, Figs. 4-24 and 4-27, is built in a 4 by 5 by 2-inch aluminum box with bottom plate. For portable work the converter might be operated from the same source as a transistorized communications receiver with which it is to be used, or from a car battery. A bank of 8 D-cells will provide many hours of intermittent use.

If mobile operation is planned, it would be prudent to connect an 18-volt Zener diode across J_3, to protect the transistors from transient peaks which occur in automotive electrical systems. Under normal conditions the Zener would not conduct.

Checkout

Before applying operating voltage, make a thorough check of the soldering operations on the circuit board, to be sure that the job is complete and correct, and that there are no incidental shorts.

With a test signal (generator, or on-the-air, starting with a high level) on about 145 Mc., adjust L_1, L_2, L_3, L_4, L_6 and L_8 for maximum output. If the test signal cannot be heard, it is likely that the oscillator, Q_4, has not started. In this case, adjust L_{10} until an increase in noise occurs, indicating the start of oscillation. Detune the slug out of the coil slightly, until the oscillator will start whenever voltage is applied.

The inductance of L_1 and L_4, and the position of the tap on L_1, should be adjusted for best (lowest) noise figure, if a noise generator is available. These adjustments can be made on a weak signal, if careful observation of the margin of the signal over noise is maintained. There is interaction between these adjustments, so several resettings of each may have to be made for optimum reception. Reasonably flat response across the desired tuning range can be achieved by stagger-tuning the rest of the circuits, both r.f. and i.f., as only the input and neutralizing coils will affect noise figure measurably.

The gain control, R_1, should be set so that the noise level, with no signal, just shows on the receiver S-meter. In this way the signal readings will then be more useful than is often the case with converter-receiver combinations where no i.f. gain control is included. The setting for various receivers may vary markedly, but it should be remembered that the position of this control has no bearing on the ability of the system to respond to weak signals, if it is set high enough so that the noise output of the converter can be heard, or seen on the meter.

HIGH-PERFORMANCE 220-MC CONVERTER

The superiority of transistors over tubes becomes more marked as the upper frequency limit of the tubes concerned is approached. Thus a well-designed 220-Mc. converter using the better transistors may outperform one using anything but the most expensive and hard-to-get vaccuum tubes. The 220-Mc. converter of Fig. 4-30 is almost a duplicate of the 144-Mc. model shown earlier in this chapter. Its weak-signal sensitivity should be better than has been possible heretofor at this frequency, for anything of comparable simplicity and moderate cost. It was built by Tom McMullen, W1QVF.

To save space and avoid duplication, only those portions of the converter that are different from the 144-Mc. version are discussed here. An identical circuit board is used. The circuit, Fig. 4-31, is similar, but not identical to that of the 144-Mc. converter. The same parts designations are used insofar as possible. Self-supporting coils and cylindrical ceramic trimmers are used in the r.f. circuits. The first r.f. stage has capacitive

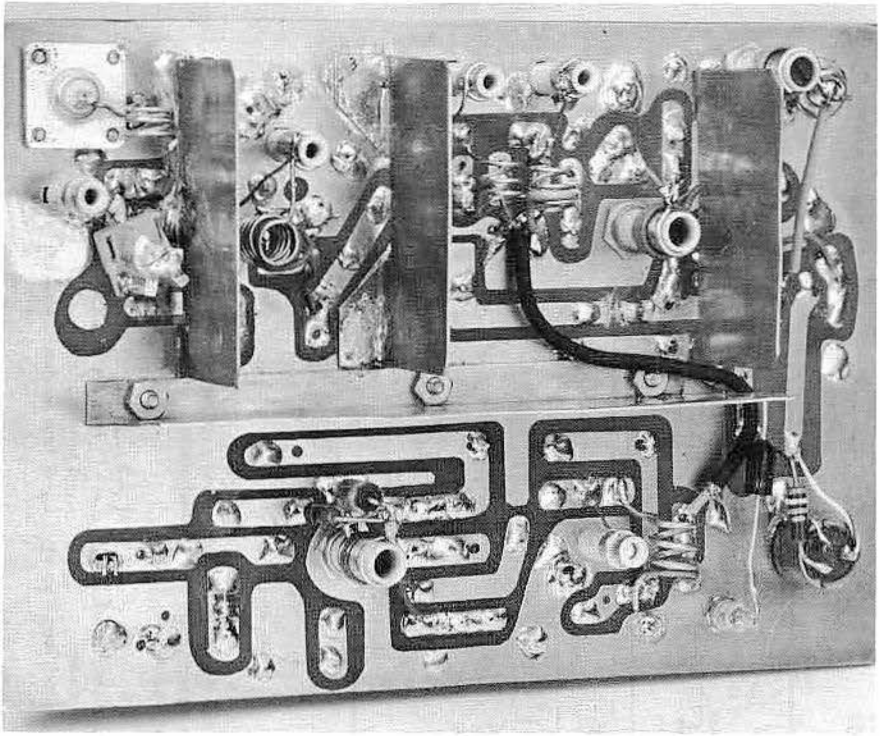


Fig. 4-30—Interior of the 220-Mc. FET converter. Minor differences from the 144-Mc. model, Fig. 4-26, are discussed in the text. The r.f., mixer and i.f. amplifier circuits, left to right, occupy the upper half of the circuit board. Board layout is identical to that of Fig. 4-29.

neutralization. Injection at 192 Mc. (for 28-Mc. i.f.) is provided by a 48-Mc. crystal oscillator and a quadrupler. Oscillator voltage is zener-regulated at 9 volts.

Almost any silicon v.h.f. transistor will work in the oscillator and quadrupler stages. The r.f. and mixer are FETs. Judging from experience with the preamplifiers described elsewhere in this chapter, most v.h.f. junction FET's should work well here. A noise figure of 3 db. or better should be obtainable with several different types, in addition to the Motorola MPF series shown here. The 28-Mc. i.f. amplifier stage is not shown, as it is identical to that in the 144-Mc. converter. It is definitely recommended, not only to assure adequate gain for some of the less-effective communications receivers, but also to permit setting the desired converter output level to match the particular receiver in use.

One difference between this converter and the one for 144 Mc. might not be readily apparent, but it is important. Note the resistor, R_2, in the line to the mixer drain circuit. This is not in the 2-meter version. It was put into the 220-Mc. model when a signal-frequency resonance developed in the circuit board, causing an oscillation problem that took some chasing down! Looking at the layout drawing of the circuit board, Fig. 4-29, pick out the 12-volt bus that runs from near the middle of the board horizontally to the right, before dropping vertically into the lower half. This should be severed below the letter "A" on the sketch. The 100-ohm R_2 is bridged across the gap.

Other minor mechanical differences resulting from the slightly-modified circuitry in the r.f. portion are apparent from the photographs. The small shield between L_1 and L_{14} in the 2-meter model is not needed here. The neutralizing capacitor, C_4, appears about where L_{14} was. The cylindrical trimmers, C_3, C_5, C_6 and C_7, are mounted where the slug coils are seen in the 144-Mc. model. Note the mounting positions of the r.f. coils. L_1, L_3 and L_4 are similar: their axes parallel to the chassis. L_2 is perpendicular to it.

Adjustment

The first step should be to get the oscillator and multiplier running. It may be advisable to keep voltage off the stages other than the ones being checked, at this point. Make sure that the oscillator is on 48 Mc., and no other frequency. (In this type of circuit it is possible to get oscillation on the crystal fundamental, in this case 16 Mc., if the collector circuit does not resonate at 48 Mc.) Now fire up the quadrupler and peak C_2 for maximum energy at 192 Mc.

With the converter connected to the receiver, there should be a marked increase in noise when voltage is applied to the r.f., mixer and i.f. amplifier stages. The i.f. can be peaked for maximum noise at 28 Mc. It is helpful at this point to have

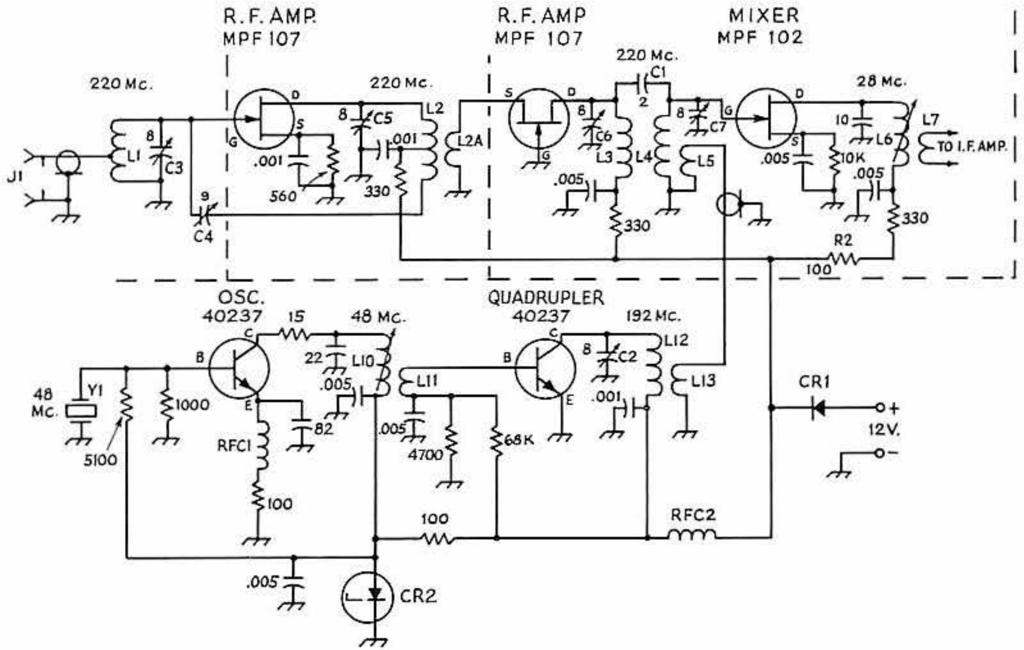


Fig. 4-31—Schematic diagram and parts information for the 220-Mc. converter. Only those portions wherein there are differences from the 144-Mc. circuit, Fig. 4-26, are shown. Parts are labelled similar to those of the 144-Mc. converter, wherever possible.

C_1—2 pf., or 2 1-inch lengths of insulated wire, twisted six times.

C_2, C_3, C_5, C_6, C_7—8-pf. cylindrical ceramic trimmer (Centralab 829-7).

C_4—9-pf. subminiature trimmer (Johnson 189-503-4).

CR_1—Silicon diode, 50 p.r.v. or greater, 200 ma. or more.

CR_2—9-volt Zener diode.

J_1—BNC coaxial fitting.

L_1, L_3, L_4, L_{12}—3 turns No. 22, ¼-inch diam., ¼ inch long. Tap L_1 at one turn from ground end.

L_2—6½ turns No. 22, ¼-inch diam., ½ inch long. Tap at 2 turns from top end.

L_{2A}—3 turns insulated wire between turns of L_2.

L_5—1 turn insulated wire between bottom turns of L_4.

L_6, L_7—Same as in 144-Mc. model.

L_{10}—7 turns No. 22 5/16 inch long, on ¼-inch iron-slug form (Miller 4500-4 form).

L_{11}—2 turns insulated wire over bottom turns of L_{10}.

L_{13}—1 turn insulated wire between first two turns of L_{12}.

R_2—See text.

RFC_1, RFC_2—25-uh. r.f. choke (Millen J-300-25).

Y_1—48-Mc. third-overtone crystal.

a signal on 220. A dipper signal will do. It is also desirable to have a properly-matched antenna connected to J_1, unless a good signal generator with 50-ohm termination is available for alignment purposes. If a random antenna must be used, put a 50-ohm resistor across J_1 to simulate the eventual load, for neutralization purposes.

There may be no oscillation in the r.f. stages, regardless of tuning, if the converter is operated with a proper load. If this is the case it is merely necessary to adjust the neutralizing capacitor, C_4, and the tuning of the input circuit, L_1C_3, for best signal-to-noise ratio on a weak signal. All other circuits affect only the gain and frequency response characteristics, so they can be adjusted for flat response across the desired frequency range, and there will be no sacrifice in the ability of the system to respond to weak signals.

Most realistic operation of the receiver's S-meter will be obtained if the meter adjustment is set so that there is an appreciable reading on noise only, with no signal. The converter i.f. gain control is then set so that the meter reads S-0 or S-1, with the antenna on. In this way the relative strength of signals will be indicated on the meter, within the usual variations encountered with these none-too-reliable devices. The receiver's antenna trimmer, if there is one, can also be used as an auxilliary gain control, and it will have no effect whatever on the sensitivity of the system.

FET PREAMPLIFIERS FOR 50, 144 AND 220 MC.

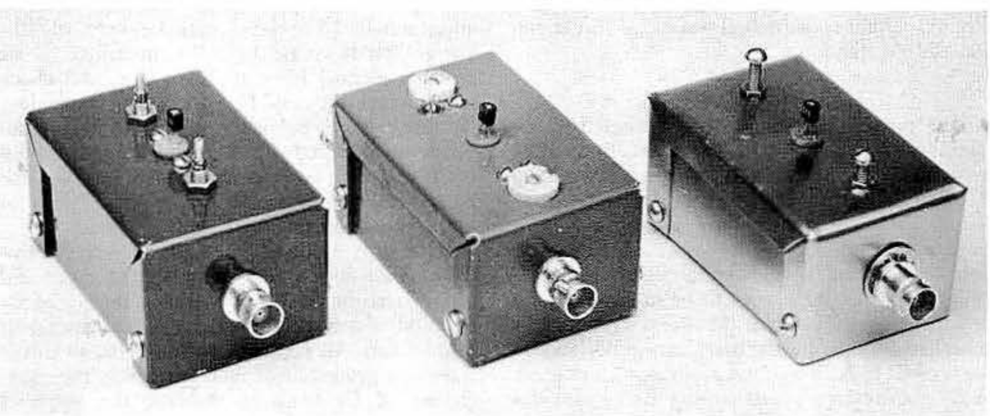


Fig. 4-32—Transistor preamplifiers for 50, 144 and 220 Mc., left to right. Appearance is similar, except for the type of tuned circuit used.

Where a v.h.f. receiver lacks gain, or has a poor noise figure, an external preamplifier can improve its ability to detect weak signals. Some multiband receivers that include the 50-Mc. band are not as good as they might be on 6. Converters for 144 Mc. having pentode r.f. stages, or using some of the earlier dual triodes, may also need some help. Most 220-Mc. converters are marginal performers, at best. The field-effect transistor preamplifiers of Fig. 4-32 should improve results with these, and with any other receivers for these bands that may not be in optimum working condition.

The circuits of the amplifiers are similar, though iron-core coils are used in the 50-Mc. model, and air-wound coils in the other two. The grounded-source circuit requires neutralization. This is done with a capacitive feedback adjustment, rather than with the inductive circuit commonly used. A tapped input circuit is used in the 50-Mc. amplifier, and capacitive input is shown for the other two, though this was done mainly to show alternative circuits. The output circuit is matched to the receiver input by means of C_2.

Many inexpensive transistors will work well in these amplifiers. Motorola MPF-102, 104 and 106, all low-priced molded-plastic units and the more expensive metal-case 2N4416 were tried, and all were more than adequate. The MPF-102 is the least expensive, and, surprisingly, it was as good as any, even on 220 Mc. Careful readjustment is required when changing transistors, so the builder should not jump to conclusions about the relative merit of different types.

Construction

The amplifiers were built in small handmade boxes, aluminum for the 50- and 144-Mc. models, and flashing copper for the 220-Mc. one, but any small metal box should do. Those shown are 1½ by 2 by 3 inches in size. The transistor socket is in the middle of the top surface, and the BNC input and output fittings are centered on the ends. The tuned circuits are roughly ¾ inch either side of the transistor socket, but this should be adjusted for good layout with the parts available. Flat ceramic trimmers are used for tuning the 144-Mc. amplifier, and the cylindrical

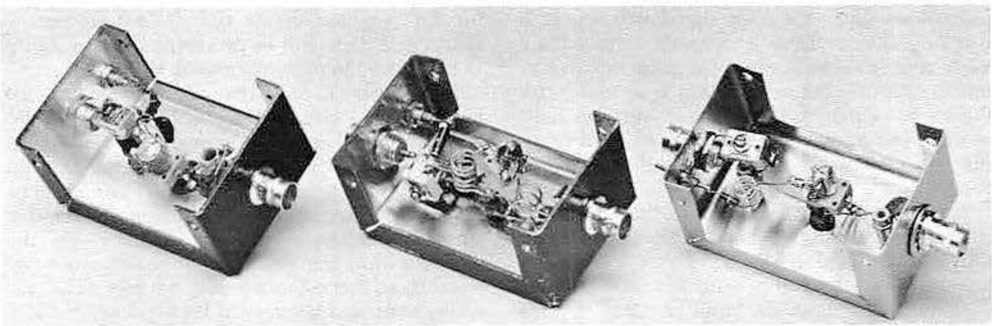


Fig. 4-33—Interiors of the FET preamplifiers, in the same order as in Fig. 4-32. The input end is toward the left in each unit.

type in the 220-Mc. one. Sockets were used mainly to permit trying various transistors; they could be wired directly in place equally well. Printed-circuit construction would be fine, if you like this method.

Adjustment

The preamplifier should be connected to the receiver or converter with which it is to be used, with any length of coaxial cable, or by hooking J_2 directly to the converter input jack with a suitable adaptor. If you have a noise generator or signal generator, connect it to J_1. If not, use a test signal from a grid-dip oscillator, or some other signal source known to be in the band for which the amplifier was designed. Preferably a matched antenna for the band in question should be hooked to J_1, if a signal generator is not used. A 50-ohm resistor across J_1 may be helpful if a random antenna is used for the adjustment work.

Set the neutralizing capacitor near half capacitance; then, with no voltage yet applied, tune the input and output circuits roughly for maximum signal. (The level may be only slightly lower than it would be with the converter or receiver alone.) Now apply voltage, and check current drain. It should be 4 to 7 ma., depending on the voltage. Probably there will be an increase in noise and signal when voltage is turned on. If not, the stage may be oscillating. This will be evident from erratic tuning and bursts of noise when adjustments are attempted.

If there is oscillation (and it is likely) move C_1 in small increments, retuning the input and output circuits each time, until a setting of C_1 is found where oscillation ceases, and the signal is amplified. All adjustments interlock, so this is a see-saw procedure at first. Increasing the capacitance of C_2 tends to stabilize the amplifier through increased loading, but if carried too far

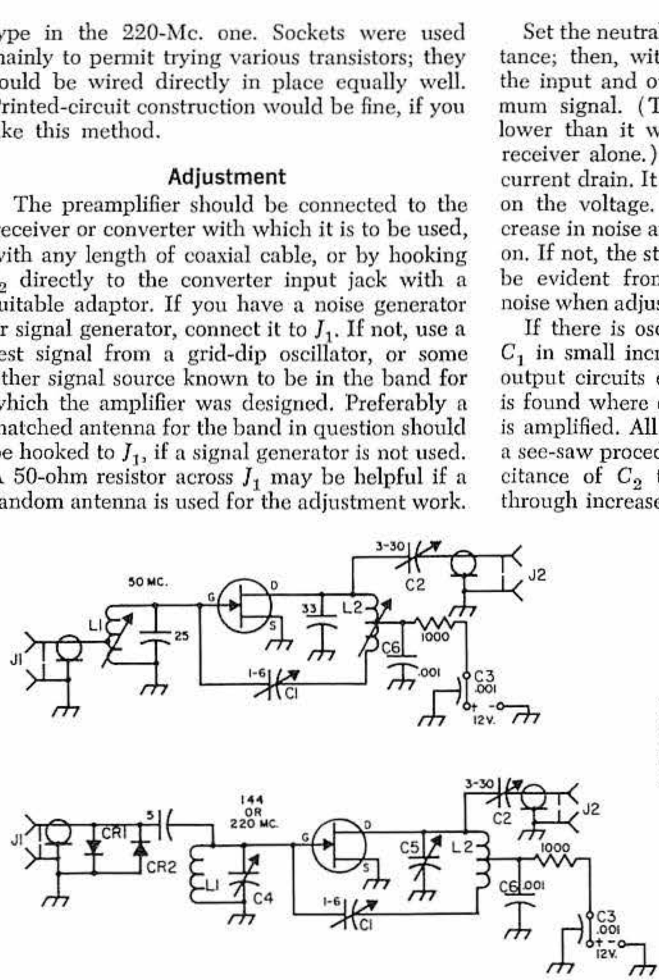


Fig. 4-34—Circuit diagrams and parts information for the FET preamplifiers. Values of capacitors not described are in picofarads (pf. or $\mu\mu$f.).

C_1—1.3 to 6.7-pf. subminiature variable (Johnson 189-502-5).

C_2—3 to 30-pf. miniature mica trimmer.

C_3—0.001-μf. feedthrough (Centralab MFT-1000; FT-1000 in 220-Mc. amplifier).

C_4, C_5—3 to 12-pf. ceramic trimmer in 144-Mc. amplifier; 1 to 6-pf. cylindrical ceramic in 220.

C_6—0.001-μf. 50-volt mylar. Omitted in 220-Mc. model.

CR_1, CR_2—1N34A or similar germanium diode.

J_1 J_2—Coaxial fitting. BNC type shown.

L_1—50 Mc.; 7 turns No. 24 enamel on ¼-inch iron-slug ceramic form, tapped at 3 turns from ground end (Form is Miller 4500). 144 Mc.: 3 turns No. 22, ¼-inch diam., ⅜ inch long. 220 Mc.: same, but with 2 turns ⅛ inch long.

L_2—50 Mc.: 10 turns like L_1, but center-tapped. 144 Mc.: 5 turns No. 22, ¼-inch diam., ½ inch long, center-tapped. 220 Mc.: Same but 4 turns.

will have an adverse effect on gain. The best setting is one where the input and output circuits do not tune too critically, but the gain is adequate.

The input circuit is first peaked for maximum signal, but final adjustment should be for best signal-to-noise ratio. This process is very similar to that with tube amplifiers, and the best point will probably be found with the input circuit detuned on the low-frequency side of the gain peak. In listening to a weak modulated signal, the fact that the noise drops off faster than the signal with a slight detuning is quite obvious. Typically the meter reading may drop about one full S-unit, while the noise level drops two S-units. The exact setting depends on the neutralization, and on the loading, both input and output, and can only be determined by experiment, with a noise generator or a weak signal.

Results

Because external noise is more of a limiting factor in 50-Mc. reception than on the higher bands, tuning for best reception is not critical on this band. Very likely you can set the neutralization to prevent oscillation, peak the input and output circuits roughly, and you'll be all set. On 144 the job is fussier if the amplifier is to effect a real improvement, particularly if your receiver is a fairly good one. This preamplifier should get you down to the point where external noise limits your reception, for sure, if you were not there before. On 220 the preamp is almost certain to help, unless you already have an exceptional receiving setup, and optimum performance is worth the trouble you take to get it. With all three, you should be certain that, if a given signal can be heard in your location, on your antenna, you will now be able to hear it.

Warning: if the preamp is to be used with a transceiver, be sure to connect it in the line to the receiver only, not in the main line from the transceiver to the antenna. It is best to do this before any work is done on the amplifier; otherwise you're sure to throw the send-receive switch inadvertently, and finish off the transistor.

If you're in doubt about the possibility of r.f. coming down the antenna line, connect protective diodes across the input, as shown with CR_1 and CR_2 in one of the circuits. Install these after the preamplifier tuneup, and check weak-signal reception with and without them, to be sure that they are not causing signal loss. Junction-type field-effect transistors are capable of withstanding much more r.f. voltage than bipolar transistors, so this kind of protection may not be needed in situations where it would have been mandatory with earlier types of transistor front ends.

432-Mc. Version

Results with these preamplifiers were so gratifying that a 432-Mc. model was tried. This was quite similar in layout, except that the metal case 2N4416 was used, and it was wired directly in place instead of using a socket. The transistor was suspended in a small notch in the bottom edge of a shield, which was mounted across the middle of the assembly. The case and source leads were soldered to the shield, with the gate lead projecting into the front compartment and the drain lead into the rear. The trimmers, C_4 and C_5, were 0.5 to 3 pf., and the input and output loading capacitors were 6-pf. miniature variables like C_1 in the other units. The coils were No. 20 tinned, 3/16-inch diameter, 1¾ turns in L_1 and 3 turns, center-tapped for L_2.

After some considerable juggling of adjustments, this stage was stabilized, and then tested with a poor crystal-mixer converter that serves as a trial horse in such work. The preamplifier gave about 10 db. gain, and this was all improvement in the converter noise figure. But it was not enough; this test setup requires about 18 db. gain for optimum performance and complete over-riding of the mixer and i.f. noise. For setups needing only a few db. gain, such a preamp should do very well.

50-MC. CONVERTER WITH 12-VOLT NUVISTORS

With the advent of 12-volt ignition systems and all-transistor broadcast receivers for car use, amateurs wishing to use the car receiver as an i.f. system for v.h.f. converters have been faced with something of a dilemma. Their broadcast receivers now have no power supply from which to draw plate voltage for the converter, and 12-volt hybrid tubes available have not been satisfactory for v.h.f. use.

The Nuvistor is now available in a model that works very well at low plate voltages. The 8056 is similar to the 6CW4, except that it has a transconductance almost as high when working at only 12 to 24 volts on the plate. It can be used in similar circuits, and it makes for very

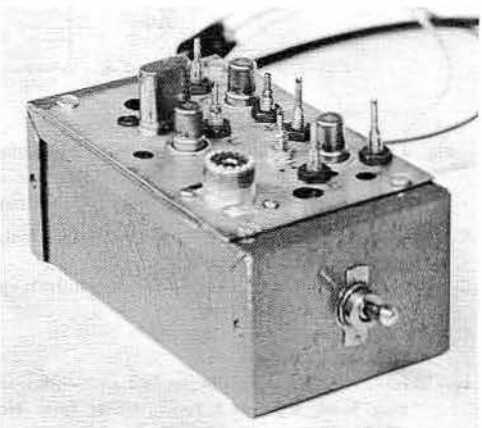


Fig. 4-35—50-Mc. mobile converter, using the 12-volt Nuvistors. I.f. output is 600 kc.

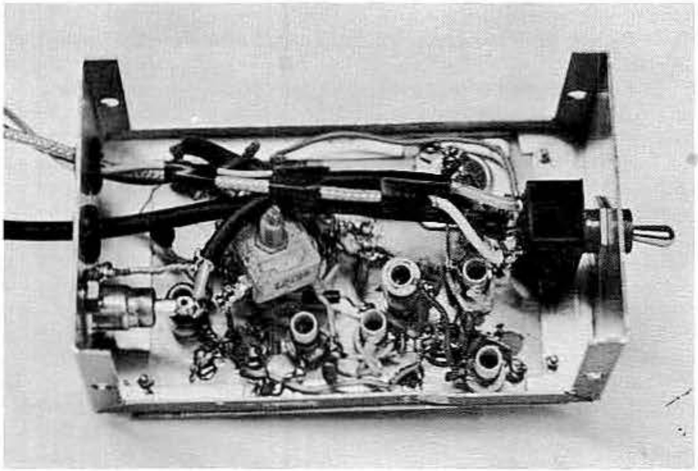


Fig. 4-36—Interior view of the W2UTH mobile converter.

good v.h.f. converter performance with no plate supply other than the 12-volt car battery.

The 50-Mc. converter of Fig. 4-35 was built by W2UTH, Victor, N.Y., for mobile use. Its circuit is only a slight modification of that given for 50-Mc. work with the 6CW4, in the converter series described earlier in this chapter. The output frequency is 600 kc., permitting coverage of the first megacycle of the band by tuning the car receiver from 600 to 1600 kc. The series-parallel hookup of the heaters, lower left portion of Fig. 4-37, takes care of using three 8056s from a 12-volt source.

The layout was modified slightly to permit construction of the converter in a 2¼ by 2¼ by 4-inch Minibox. The 58-Mc. trap in the origi-

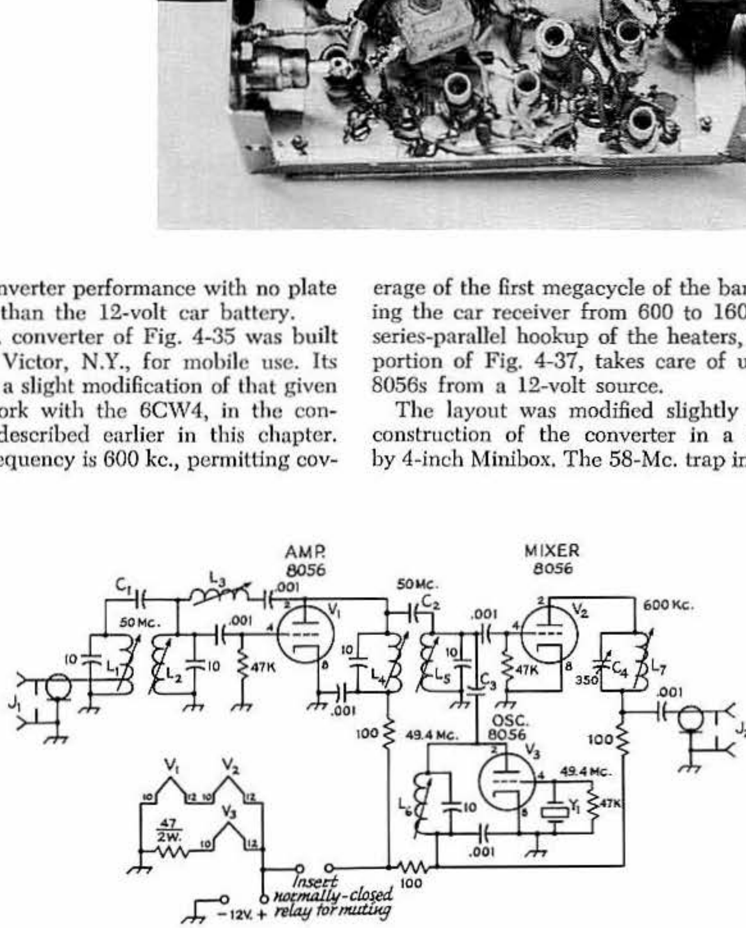


Fig. 4-37—Schematic diagram and parts information for the hybrid-Nuvistor converter. Decimal values of capacitance are in µf.; others in pf. Capacitors are ceramic unless specified. Resistors are composition, ½ watt unless specified.

C_1, C_2—No. 22 insulated hookup wires 2 inches long, twisted together 1½ inches.

C_3—No. 22 insulated hookup wires one-inch long, twisted together ½ inch.

C_4—Mica padder, approx. 350 pf.

J_1, J_2—Coaxial receptacle.

L_1—10 turns No. 28 enam., close-wound on ¼-inch iron-slug form; tap at 3 turns (Miller form No. 20A000RBI).

L_2, L_4, L_5—8 turns No. 28 enam., close-wound on ¼-inch iron-slug form (Miller 20A687RBI). L_1 and L_2 are ⅞ inch apart, c. to c.; L_4 and L_5, ¾ inch apart; L_5 *and* L_6, ⅞ inch apart.

L_3—No. 32 enam., close-wound ⅛ inch on ¼-inch iron-slug form (Miller 20A686RBI).

L_6—6 turns No. 32 enam., close-wound on ¼-inch iron-slug form (Miller form No. 20A000RBI).

L_7—Ferrite antenna coil for broadcast band (Miller 6300). Tune to 600 kc. with C_4.

Y_1—49.4-Mc. crystal (International Crystal Mfg. Co. F-605).

nal 6CW4 converter can be eliminated, as the problem it was designed to correct does not exist with an i.f. in the broadcast band. Adjustment is similar to the procedure given for the earlier model. W2UTH has found the performance of this converter-receiver combination to be considerably better than that of a number of commercially-available 50-Mc. transceivers.

SUPERREGENERATIVE RECEIVERS FOR 50 AND 144 MC.

The strong and weak points of superregenerative detectors are discussed at some length in Chapter 3. Receivers of this type for the most popular v.h.f. bands are shown in Fig. 4-38. Each has features that make the most of this simplest of all practical receiving systems, while playing down its principal limitations, insofar as possible. An r.f. amplifier ahead of the detector provides some isolation, reducing radiation and making adjustment less critical than would be the case with the antenna coupled directly to the detector.

The receivers will not make anyone the weak-signal DX champion of his call area, but they are capable of usable reception on any a.m. or wideband f.m. signal that would be solidly readable on more sophisticated gear. They won't separate signals a few kilocycles apart, but they do better than most simple receivers, thanks to their high-Q detector circuits and r.f. amplifiers. Either will radiate some interference, and reradiate any signal being received, so they should be used with discretion where other stations are operating on the same band in the same area, but they are not the neighborhood curse that one-tube rushboxes are bound to be. They are not the simplest you could build and still hear signals, but they are cheaper and easier to make than any other complete v.h.f. receiving system of comparable performance.

The two receivers are similar, to simplify the project for anyone interested in making both. The chassis are alike, as are the audio stages. The same power supply will do for either. The audio portion uses a 6AV6 triode, set up primarily for headphone use. Any audio pentode could be substituted, if a suitable output transformer were connected in place of the plate-dropping resistor and coupling capacitor. The audio level then would be adequate for speaker operation.

The 50-Mc. Model

Two tubes are used in the 50-Mc. receiver. The pentode portion of a 6U8A is a broadband r.f. amplifier, and the triode is the superregenerative detector. As set up it tunes the entire 50-Mc. band, with some leeway on either side. More or less tuning range, to suit the builder's preference, can be had by using a different number of plates in the detector tuning capacitor, C_1, in Fig. 4-39. The r.f. amplifier input circuit, L_1, is peaked for maximum signal when the receiver is placed in service, and requires no retuning thereafter.

The principal circuit novelty is the use of a section of coaxial line, L_2, for the detector tuned circuit. This is self-shielded, and it gives somewhat better selectivity and smoother control of regeneration than would be possible with a coil-and-capacitor circuit. Regeneration is

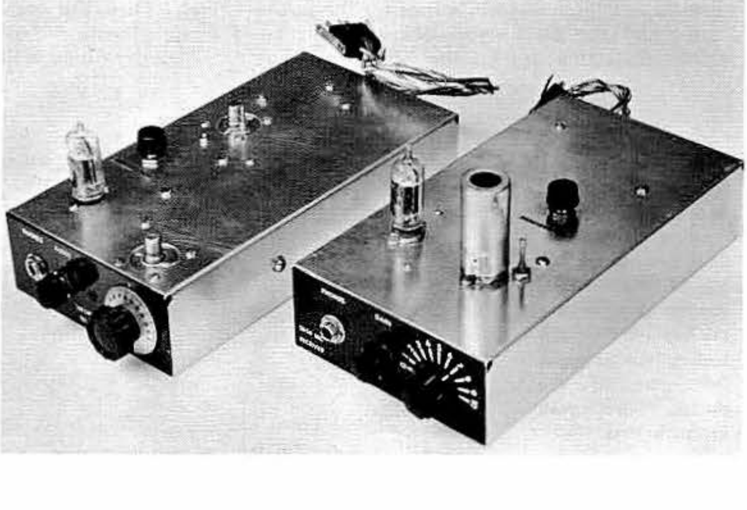


Fig. 4-38—Superregenerative receivers for 50 and 144 Mc. The 50-Mc. model, right, uses a 6U8A triode-pentode as a combined r.f. amplifier and detector. The 144-Mc. version has 6CW4 Nuvistors for these functions. The audio amplifier is a 6AV6 triode in each receiver.

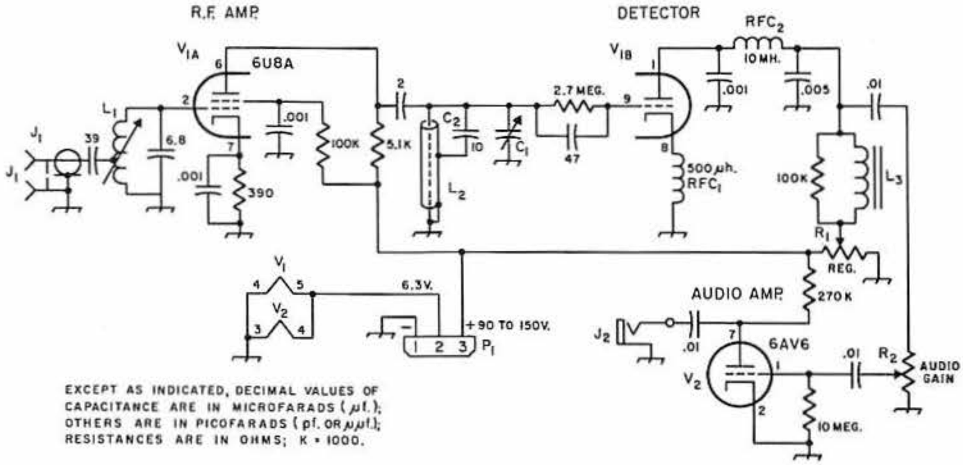


Fig. 4-39—Schematic diagram and parts information for the 50-Mc. receiver.

C_1—Double-spaced variable, originally 30 pf. (Hammarlund HF-30X reduced to 5 stator and 6 rotor plates—see text).

C_2—10-pf. dipped mica.

J_1—Phono jack or coaxial connector.

J_2—Single-circuit jack.

L_1—11 turns No. 24 enamel on ¼-inch slug-tuned form, tap at 3 turns from ground end.

L_2—Coaxial-line tank circuit—see text.

L_3—20-hy. 15-ma. filter choke. Output transformer primary also suitable.

P_1—3-pin male cable connector.

R_1—0.25-megohm control, linear taper.

R_2—1-megohm control, audio taper.

RFC_1—500-µh. r.f. choke.

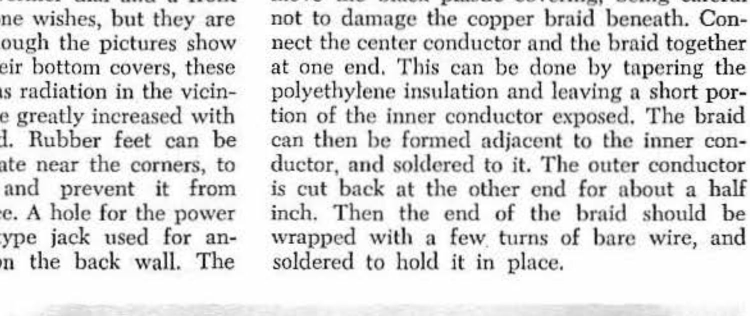
RFC_2—10-mh. r.f. choke.

controlled by varying the detector plate voltage by means of R_1.

The chassis is aluminum, 5 by 9½ by 2 inches in size, with the front end serving as the panel. Refinements such as a vernier dial and a front panel can be added if one wishes, but they are not really necessary. Though the pictures show the receivers without their bottom covers, these should not be omitted, as radiation in the vicinity of the receiver will be greatly increased with the r.f. circuits exposed. Rubber feet can be added to the bottom plate near the corners, to cushion the receiver and prevent it from scratching a table surface. A hole for the power leads, and the phono-type jack used for antenna connection are on the back wall. The connector can be replaced with the standard SO-239 coaxial fitting, if desired.

The detector grid circuit, L_2, is made from a 27-inch length of 72-ohm RG-11/U coax. Remove the black plastic covering, being careful not to damage the copper braid beneath. Connect the center conductor and the braid together at one end. This can be done by tapering the polyethylene insulation and leaving a short portion of the inner conductor exposed. The braid can then be formed adjacent to the inner conductor, and soldered to it. The outer conductor is cut back at the other end for about a half inch. Then the end of the braid should be wrapped with a few turns of bare wire, and soldered to hold it in place.

Fig. 4-40—Bottom view of the 50-Mc. receiver, showing the coaxial-line detector circuit, top. R.f. amplifier components are between the line and the 6U8A socket, near the upper left part of the chassis. Audio circuits at the bottom.



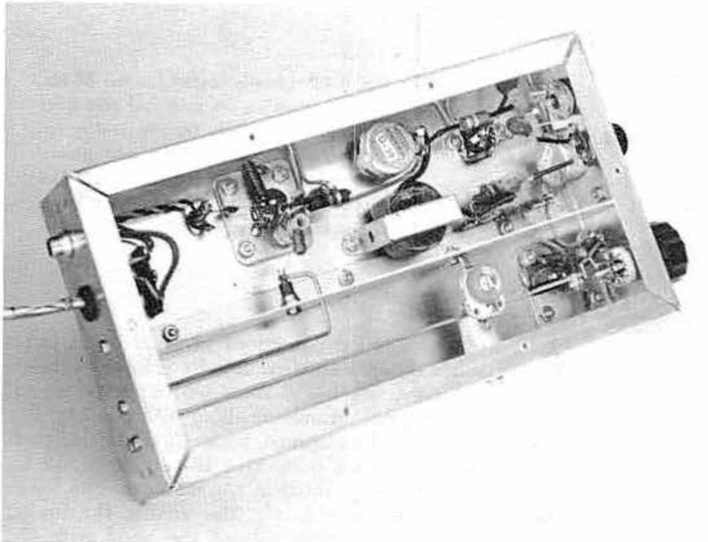


Fig. 4-41—Underside of the 144-Mc. receiver. The detector tuned circuit in this model is a trough line, at the bottom. The detector socket and small components associated with this stage are mounted inside the trough, at the right end. The r.f. amplifier portion is at the upper left, with audio components at the right front.

Now fold the coax in thirds, resulting in a length that will just fit inside the chassis. Wrap the three portions tightly together with bare wire at three points, and solder. Use as little heat as possible to do this to prevent melting of the insulation. The completed tank circuit is then put into the inside edge of the chassis and clamped in place with two brackets cut from brass or aluminum. The position should be such that the open end of the coax is adjacent to the stator bar of the detector tuning capacitor, C_1.

The tuning capacitor is a double-spaced 30-pf. model, with some of its plates removed. The number to be removed can be determined after the builder gets the receiver working, and decides how much tuning range (or bandspread) he wants. The fixed padder capacitor C_2 is connected from the cut-off left stator post, as viewed from the bottom, to the wire wrapping at the open end of L_2. Other connections to C_1 are made to the stator bar on the right side of the capacitor. If the layout drawing, Fig. 4-43, is followed, placement of parts should present no problems. Wire the r.f. circuits with as short

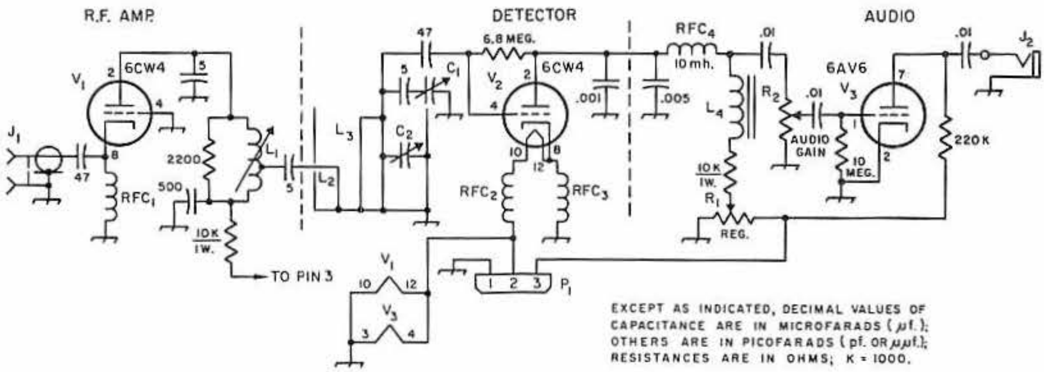


Fig. 4-42—Schematic diagram and parts information for the 144-Mc. superregenerative receiver. All components shown between the broken lines are inside the detector tuned-circuit assembly.

C_1—1.5 to 5-pf. miniature trimmer (Johnson 160-102).
C_2—1.5 to 13-pf. ceramic trimmer.
J_1—Phono jack or coaxial connector.
J_2—Single-circuit jack.
L_1—4 turns No. 24, 3/8 inch long, on 1/4-inch slug-tuned form. Tap 1 turn from B-plus end.
L_2—Coupling loop, 5 inches No. 14, main portion 4 3/4 inches long, spaced 3/8 inch from center conductor of L_3.
L_3—Trough line; see Fig. 4-48.
L_4—20-hy. 15-ma. audio choke. Primary of output transformer also usable.
P_1—3-pin male power cable connector.
R_1—0.1-megohm control, 2-watt, linear taper.
R_2—0.5-megohm control, audio taper.
RFC_1, RFC_2, RFC_3—1.8-µh. r.f. choke (Ohmite Z-144).
RFC_4—10-mh. r.f. choke.

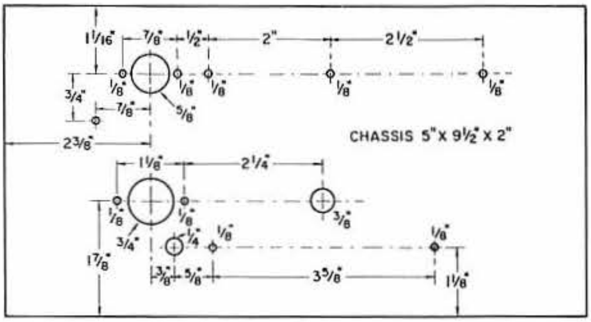


Fig. 4-43—Chassis layout for the 50-Mc. receiver, showing the principal hole sizes and locations. Holes for the tuning capacitor, audio gain control and phone jack are centered on the front wall. The antenna connector and cable-exit holes are on the back wall.

leads as possible.

Mount the 6U8 socket so that Pins 1 and 9 are toward the front of the receiver. A 4-lug tie-point strip (one grounded) is mounted in back of the 6U8 socket, and another about midway between the sockets and an inch in front of them. A 3-lug strip near the back is used for a cable termination. Connection between the first tiepoint strip and the antenna jack can be made with a small coax if the builder wishes, though twisted hookup wire works equally well. Placement of audio parts and power leads is not critical.

Adjustment

The receiver can be placed in operation with C_1 set up for a wide tuning range, with only 3 rotor and 3 stator plates removed. It will then cover from about 45 to 55 Mc. It may be of interest to have this much range, as it permits listening to many of the commercial users of frequencies below the band, to get a good idea of propagation conditions in areas or at times when 6-meter amateur activity is low. As shown the capacitor has 6 rotor and 5 stator plates, which gives a range of about 49 to 54 Mc. For adjustable bandspread, make C_2 a variable padder, and reset it to give the coverage on C_1 that may be wanted at the moment. The lower the capacitance in C_2, the greater the tuning range on the high side of the band with C_1. Increase C_2 to extend the tuning range on the low side.

Control of regeneration should be quite smooth with the 50-Mc. receiver. With R_1 turned up to the point where the rushing noise is smooth the tuning will be the least critical. Turning down regeneration will result in a beat note being heard as a signal is tuned in. C.w. signals can be copied in this position, and if great care is used it may be possible to get readable reception of s.s.b. signals.

Adjustment of the slug in L_1 is not fussy. Tune in a signal in the middle of the range you

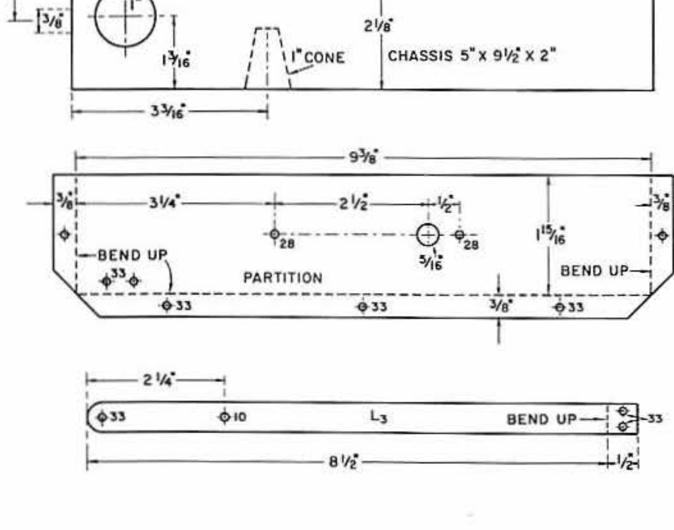


Fig. 4-44—Metalwork layout for the 144-Mc. receiver. Dimensions at the left edge of the chassis, top, are for hole centers on the front wall, for the phone jack, audio gain control and main tuning control, respectively, reading down from the top. The trough wall, center, and stripline inner conductor, bottom, are shown before bending. Numbers adjacent to small holes are approximate drill sizes. Mounting holes in the partition can be located best by marking to match chassis holes, after bending the partition.

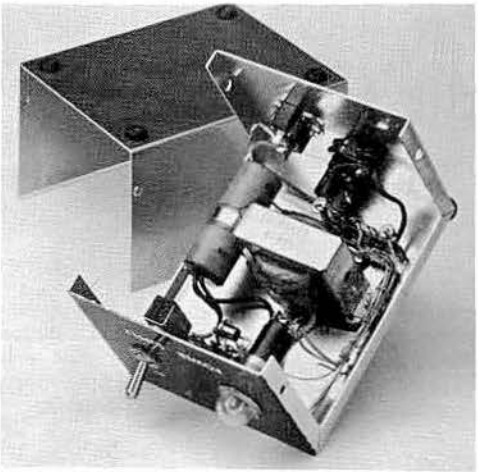


Fig. 4-45—Power supply for use with the superregenerative receivers or other equipment items requiring similar current and voltage.

want to use most often, and then adjust the slug for best reception.

The 144-Mc. Version

Two 6CW4 or 6DS4 Nuvistors are used as r.f. amplifier and detector in the 144-Mc. model. The first is a grounded-grid stage, in the interest of circuit simplification. It is lightly coupled to a strip-line tank circuit in the grid of the superregenerative detector. (See Fig. 4-42.) The audio stage is similar to the 50-Mc. receiver.

The 6CW4 sockets and associated components are mounted on rectangular brass or copper plates about 1¼ by 1½ inches in size. This permits good grounding, and also allows the builder to assemble the critical parts of the circuit before the plate is mounted in the chassis. When the principal connections have been made the subassemblies can be mounted in the 1-inch chassis holes shown in the layout drawing, Fig. 4-44.

The resonant element of L_3 can be copper, brass or aluminum, preferably at least 1/16 inch thick. Dimensions of this and the partition that comprises the inner wall of the trough line are shown, before bending, in Fig. 4-44. The line is held in place by two 4-40 screws and nuts at the bent end, and by a 1-inch ceramic cone insulator 2½ inches from its high-impedance end. The cone also serves as an anchor point for the tuned-circuit padder, C_2.

Coupling into the tuned circuit is by means of a wire loop, L_2, which is tapped down on the r.f. amplifier plate coil, L_1, with a small capacitance in series to prevent shorting the plate voltage. The position of L_2 with respect to L_3 is adjusted for the tightest coupling that can be had without introducing dead spots in the tuning range due to "suck-out" effect of L_1. This turns out to be a spacing of about ⅜ inch. The top end of L_2 is mounted on a small standoff or tiepoint, adjacent to L_1.

Heater and plate power leads for the detector are brought out through the partition through two small holes near the socket. R.f. leads in the detector and amplifier circuits should be as short as possible. Grounds may be made to the plates on which the sockets are mounted. Arrangement of audio components is not critical. The layout drawing can be followed, for convenience, but other arrangements should work equally well.

Adjustment and Use

With power on, turn up the regeneration control, R_1, until the smooth hiss of superregeneration is heard throughout the tuning range, but no further than necessary. Tuning range can be checked roughly by means of a grid-dip oscillator, using it as a signal generator. Adjust the slug position in L_1 for maximum signal. This will interact with the detector tuning, making it necessary to "follow" the signal with the detector tuning, until maximum noise suppression is achieved. A modulated signal received on the air, or from a signal generator, makes adjustment easier, as suppression of the hiss by the signal is not an entirely reliable way to judge proper tuning. Some practice in adjusting the regeneration and coupling will enable the operator to get optimum results in reception of weak signals.

With either receiver an outboard amplifier may be driven nicely from the 6AV6. The audio level from the amplifier, as shown, is adequate for speaker use under quiet room conditions. The audio stage may be converted to any audio pentode, for better speaker operation, by substitution of an output transformer in the plate circuit of the amplifier and rewiring the tube socket for a 6AQ5 or 6AK6.

The tuning range of the detector can be adjusted to suit the builder. Using the tuning capacitor, C_1, in its original state the detector tunes the band with about two megacycles leeway on each end, when C_2 is properly set. Increasing the setting of C_2 lowers the actual frequency range covered, and slightly increases the spread of a given number of megacycles on C_1. More tuning range can be obtained by using the next size larger tuning capacitor for C_1. As presently used, the capacitor has 2 stator and 2 rotor plates, which gives a tuning range of about six megacycles.

Power Supply

The receivers require 6.3 volts a.c. or d.c. at about ¾ ampere, and 90 to 150 volts d.c. at 20 ma. The latter can be from 90 volts of B battery, if this is desirable. In this case, a switch should be inserted in the power lead from Pin 3, to open the circuit when the receiver is not in use. Otherwise there will be a small drain

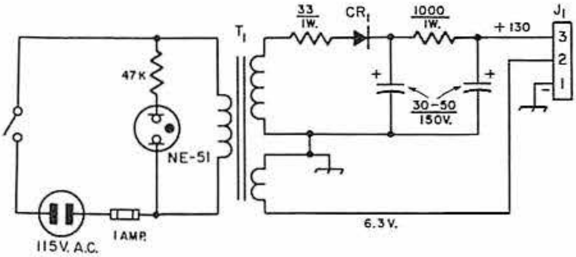


Fig. 4-46—Schematic diagram of the receiver power supply. Filter capacitors are electrolytic.
J_1—3-pin female power connector.
T_1—Small replacement transformer, 125 volts at 30 ma., 6.3 volts at 1 amp.
CR_1—400-volt p.i.v. silicon diode, 500 ma.

on the batteries due to the regeneration control. This switch could be an integral part of either the regeneration or audio volume control, if desired.

Details of a very simple power supply are given in Figs. 4-45 and 4-46. This type of supply is useful for many purposes around a station where the owner likes to build small units such as preamplifiers, test equipment and other items requiring only a small amount of current.

A BLANKER FOR PULSE-TYPE INTERFERENCE

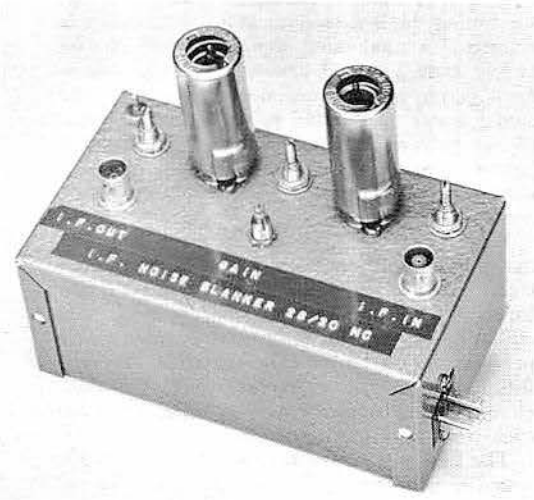


Fig. 4-47—A noise blanker for insertion between a v.h.f. or u.h.f. converter and the following communications receiver. Model shown is set up for 28 to 30 Mc., but other converter output frequencies could be used by altering circuit constants.

Our frequencies from 220 Mc. up are blessed with a minimum of static, but cursed with a maximum of pulse-type noise, thanks to the shared nature of our assignments above 200 Mc. The "Government Radio Positioning Service" is in there, too. This means radar interference, and radar looks like about the worst interference you can have. The situation could be worse, however, for elimination of static has yet to be accomplished, while short pulses of the radar variety can be blanked out very effectively with suitable equipment.

Radar pulses, ignition and most other noise encountered in the world above 200 Mc. have fast rise time and thus can be dealt with if eliminated from the receiver before they get to circuits that are selective enough to lengthen the pulse. Ideally a noise blanker should be applied at the antenna terminals of the receiver. This is not necessarily impossible, but it is difficult, so customary practice is to employ a noise blanker in the circuits after the first mixer, before the selectivity-determining stages. This type of circuit as described in all modern editions of the ARRL *Handbook* for use at 455 kc. is adequate for eliminating auto ignition and the like, but radar interference is several orders of magnitude more severe. A noise blanker used in this manner is too late in the circuit of a v.h.f. converter-receiver combination to prevent overloading of earlier stages.

The blanker of Fig. 4-47 is installed between the v.h.f. or u.h.f. converter and the receiver used as the i.f. system, allowing the noise pulses to be eliminated before they reach the high-gain stages. One advantage of this approach is that no modification of existing equipment is required. The noise blanker is connected in the cable between the converter and receiver, and requires no other connections to either.

Circuit Description

The blanker uses two 6AG5s as amplifiers at the converter output frequency, in this instance 28 to 30 Mc. or 14 to 18 Mc. It could be any other converter i.f., with suitable modification of the tuned circuits. The 6AG5s were used mainly because of their ready availability; other pentodes such as the 6BH6, 6BA6, 6AK5 and the like should work equally well. The input circuit of the first stage is tapped to match the low-impedance line from the converter. Single-tuned circuits are used. A gain control is connected in the cathode leads of both tubes, and maximum gain is limited by 150- and 100-ohm series resistors.

The output of the second stage is coupled into a pair of back-to-back diodes, CR_1 and CR_2. The first is an ordinary second-detector diode such as the 1N64 or 1N60, the other a selected computer diode such as the 1N920 or 1N3730. Output brought out through C_1 can be used to monitor visually the noise pulses which are being blanked. With a high-gain audio system connected to C_1 all the signals in the bandwidth of the converter can be monitored. This is not practical for hearing weak signals, but it does provide continuous monitoring on a lightly occupied band.

The i.f. output coupling for the following receiver is merely the very small capacitance of a wire connected to the center conductor of J_2, and placed near the output coil, L_3. No direct connection is made at this point. The stray capacitive coupling should be adjusted so that the gain through the noise blanker is the same as it was before the blanker was connected.

A word of warning: this is a high-gain lightly loaded amplifier. The r.f. chokes decoupling the heaters and plate leads are essential if stable operation is to be obtained. Physical layout like the original is also necessary, unless the builder is experienced in such matters.

Layout

The noise blanker is built in a standard 2 by 3 by 5-inch Minibox, with all r.f. components in a line down one side. The three coils are 2 inches apart, and the sockets are centered between them. The sockets are mounted so that the grid and plate pins, 1 and 5 respectively are in a straight line, and nearest to their respective coil terminals.

A grounding lug is placed under each socket mounting nut. A wire is run from the one adjacent to Pin 7 across the socket to the center shield ring and then to Pin. 3. Disk ceramic capacitors are connected with the shortest possible leads in the following manner: Pin 2 to Pin 3, Pin 7 to ground, Pin 4 to adjacent ground lug, and Pin 6 to this same lug. The bypasses at the bottom of the coils L_2 and L_3 are also returned to these lugs, which are adjacent to the side of the unit. Spare lugs on the coil form for L_3 are used for tiepoints, for the junction of the three diodes and for the common point of CR_3 and RFC_4.

Adjustment and Use

After the wiring has been completed and checked for errors the tubes should be installed and the coils L_1, L_2 and L_3 adjusted to the desired intermediate frequency with a grid-dip meter.

The unit should now be connected between an operating converter and its i.f. receiver. The noise level in the receiver should be adjusted by means of the stray coupling between L_3 and J_2, to be the same as before the blanker was installed.

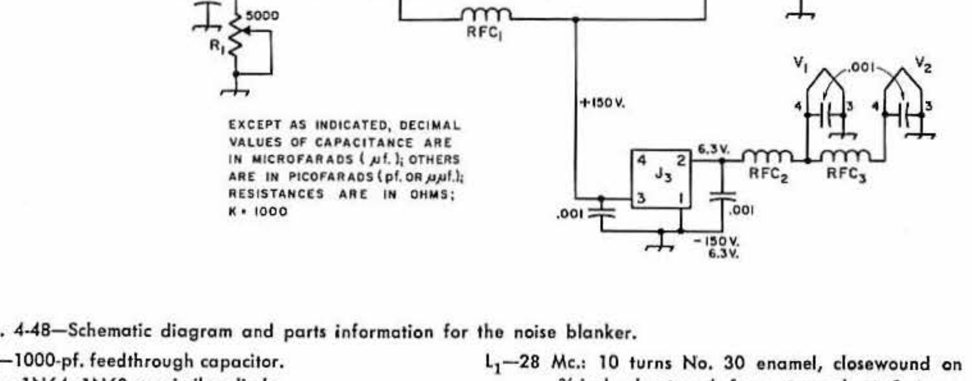


Fig. 4-48—Schematic diagram and parts information for the noise blanker.

C_1—1000-pf. feedthrough capacitor.

CR_1—1N64, 1N60, or similar diode.

CR_2—1N920 or 1N3730 switching diode.

J_1, J_2—BNC connector. J_2 has extension wire to be trimmed in length and adjusted in position with respect to L_3. See text.

J_3—4-pin male power fitting.

L_1—28 Mc.: 10 turns No. 30 enamel, closewound on ⅜-inch slug-tuned form, tapped at 3 turns. 14 Mc.: 30 turns No. 30 enamel, tapped at 10.

L_2—Like L_1, but no tap.

L_3—13 turns like L_2.

R_1—5000-ohm miniature control

RFC_1 through RFC_3—27-µh. r.f. choke.

RFC_4—500-µh. r.f. choke.

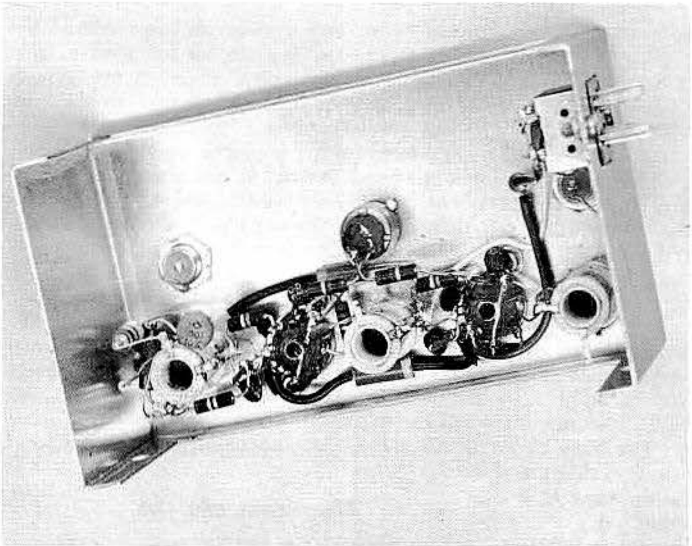


Fig. 4-49—Interior of the noise blanker. The input and power connector fittings are at the right end in this view.

The gain control in the blanker is *not* for purposes of adjusting output level. Its function is to adjust the input to the blanking diodes for optimum efficiency. Normally it will be at or near full gain. While listening to the noise it should be possible to remove the input cable from the blanker and observe that the noise drops to approximately the same level that it would if the blanker was not turned on.

It should be possible with a converter connected and the blanker gain control turned full on, to peak up L_1, L_2 and L_3 for maximum noise without any evidence of regeneration. The 3-db. bandwidth of the blanking amplifier with the coils all peaked to the same frequency is approximately ½ Mc. Operation over 1½ Mc. or more can be obtained with the coils adjusted in this manner. In order to make the final adjustments on the blanker it is necessary to have a source of noise and a method of switching a blanker in and out of the circuit. First, tune in a strong pulse-type signal such as radar or very strong automobile-ignition noise. With the blanker in the circuit, the power to the blanker should be turned off and the r.f. gain on the receiver opened wide. At this time there should be no noise coming from the converter. Turning the converter power on and off should make no change in the output noise of the receiver.

Now the blanker should be turned back on and after warm-up the amount of noise reduction should be observed. When the blanker is properly operating there should be no noise *pulses*, regardless of their amplitude when the blanker is out. If the blanker provides a substantial reduction in noise pulses, but does not completely eliminate them, it is not working properly. There are only three reasons for this. The first is feedthrough around the blanker. The test for this was performed in a previous step and it is assumed that there was no feedthrough. There may be insufficient gain in the two amplifying stages. If you are using a normal v.h.f. or u.h.f. converter having 25 to 30 db. or more of gain, and tubes similar in characteristics to a 6AC5, the likelihood of low gain is quite small. The third and most likely reason for poor operation lies in the selection of the proper diodes for CR_1 and CR_2. Some experimentation with the polarity of the diodes and the size of the diode load is usually required to obtain optimum performance. The particular constants given in the circuit were successful in three different models tested and no difficulty should be encountered in obtaining optimum operation.

Results

The performance of the noise blanker in on-the-air operation leaves little to be desired from the standpoint of external noise elimination in u.h.f. work. With or without noise, the insertion of the noise blanker in the circuit has no discernible effect on the readability of a weak signal. In the presence of pulse-type noise the signal continues to be perfectly readable and the noise is not evident at all in the output of the receiver. Radar pulses strong enough to draw grid current in the r.f. stage of the i.f. receiver are completely eliminated by the use of this blanker. However, like all good things, there are some drawbacks to the use of a noise blanker. The worst of these is that a very strong local signal will overload the noise blanker and cross-modulate other signals on the band. This is an inherent trait of noise-blanker circuits for which no solution has been found. In order to obtain sufficiently strong blanking pulses, high-gain amplifiers are required and high-gain amplifiers necessarily overload. This disadvantage is far outweighed by the ability to copy weak signals in the presence of strong pulse-type interference. Even when the blanker is overloaded, signals which could not be heard through radar interference without it are readable.

V.H.F. Transmitter Design

Before we discuss transmitting techniques for the amateur bands above 50 Mc. in detail it will be well for us to see what standards are set for us in the U.S. Regulations. The two numbered paragraphs below are not exact quotes, but they summarize pertinent regulations.

97.71—Amateur stations operating below 144 Mc. must employ adequately filtered d.c. plate power for transmitting equipment, to minimize modulation from this source.

97.73—Spurious radiations from an amateur station below 144 Mc. shall be reduced or eliminated in accordance with good engineering practice. . . . In case of A3 emission, the transmitter shall not be modulated to the extent that spurious radiation occurs, and in no case shall the carrier wave be modulated in excess of 100 percent. . . . Simultaneous amplitude and frequency modulation is not permitted. . . . The frequency of the emitted carrier wave shall be as constant as the state of the art permits.

It will be seen that stability and quality requirements imposed on all lower amateur frequencies apply equally on the 50-Mc. band, but not to 144 Mc. and higher. Over-modulation, sideband splatter, unstable carrier frequency, a.c. hum, and the like are illegal on 50 Mc., but not on 144, 220 or 420 Mc. From the standpoint of the law, there is a vast relaxation in the technical standards we must meet after we pass the 50-Mc. band on the way up through the spectrum. Looking at the question from the amateur point of view, however, there is little or none. The desirability of radiating the best signal that is technically feasible is the same throughout the v.h.f. region, if the user is interested in worthwhile results, and in causing a minimum of trouble for his fellow users of the v.h.f. bands.

It should be remembered that the use of unstable equipment is *legal* above 144 Mc., so long as the radiation from the transmitter remains entirely within the assigned frequency band. There are some circumstances where very simple and therefore unstable gear may serve useful ends, and these will be touched on later in this book, but our main emphasis will be on transmitters that employ crystal control or its equivalent in stability and freedom from spurious emissions. That such concern for clean signals is required by law only on the 50-Mc. band will be largely ignored.

TRANSMITTER DESIGN CONSIDERATIONS

Our v.h.f. and u.h.f. assignments are not in exact harmonic relationship, but they are close enough to lining up harmonically so that this factor influences the design of stabilized equipment for the bands from 50 to at least 1300 Mc. It is possible, for example, to build one good frequency-control unit that will serve for any combination of bands in this range. Whether this is a desirable approach for your set of circumstances is a decision that only you can make, but the possibility should not be ruled out in station planning.

This basic unit could be an oscillator with a wide selection of crystals, or a well-designed and accurately calibrated v.f.o. or VXO. This would be followed by such frequency-multiplier stages as may be required to give output on the desired band or bands. A similar frequency-control device can also be used with heterodyning equipment to produce replicas of the original signal on the desired frequencies. We will examine both approaches in some detail.

The Oscillator-Multiplier

Most v.h.f. transmitters employ an oscillator on some lower frequency than that at which the station is to operate, followed by one or more frequency multipliers. The starting frequency is usually in the 8-Mc. range, though 6 Mc., 12 Mc., and others may be used. Typically, a 50-Mc. transmitter is controlled at 8.334 to 9 Mc., the frequency then being tripled to 25 to 27 Mc., and doubled to 50 to 54 Mc. before amplifier stages build up the power level. A 144-Mc. transmitter may start at 8 to 8.222 Mc., which is then tripled, doubled and tripled to 144 to 148 Mc. A 220-Mc. lineup would use 8.149 to 8.333 Mc., tripled three times to 220 to 225 Mc. The portions of the 420- and 1215-Mc. bands where stabilized equipment is usually employed are in third and ninth harmonic relation to the low end of the 144-Mc. band. From these figures and from the chart, Fig. 5-1, it may be seen that coverage of 8 to 9 Mc. in the oscillator stage takes care of most of our requirements through 1300 Mc.

Many other starting frequencies and orders of multiplication are usable. The oscillator-multiplier approach is convenient in multiband designs, and it is used almost entirely in equipment other than the single-sideband variety. One weak point is that any instability in the controlling oscillator is multiplied: 6 times in the 50-Mc. example of the previous paragraph, 18 times for 144-Mc. operation, 27 times for 220

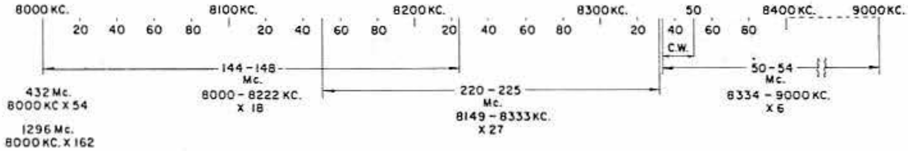


Fig. 5-1—Our bands above 50 Mc. are nearly harmonically related. The possibility of using a single frequency-control system for all bands from 50 through 1300 Mc. is illustrated in this chart. The example is for oscillators in the 8-Mc. region, but other frequency ranges such as 6 or 12 Mc. may be used.

Mc., and so on. An oscillator that seems quite stable at 8 Mc. may suffer from drift, hum modulation, mechanical instability or frequency modulation to the extent that the signal at 432 Mc. may be unacceptable to the critical worker. This is particularly true of the conventional v.f.o. The crystal oscillator is much superior for frequency control on 144 Mc. and higher bands.

The possibility that harmonics other than the desired ones will appear in the output should be considered in designing a v.h.f. exciter. Such unwanted frequencies may be a source of interference to TV, f.m. and other v.h.f. reception in the vicinity of the amateur station. They can be reduced or eliminated by taking suitable circuit precautions, but their interference potential should not be ignored.

The number of frequencies that could cause trouble can be reduced by using a high starting frequency in the v.h.f. exciter. A 24-Mc. oscillator instead of one at 6 or 8 Mc. eliminates most of the harmonics that are potential sources of TVI in the low TV channels. Starting at 48 or 50 Mc. is still better. There are good reasons for using 6 or 8 Mc., however. A stable oscillator is much more readily built in this frequency range than for 24 Mc. or higher. Crystals for 6 to 9 Mc. are inexpensive, reliable and easy to use, while those for 12 Mc. and higher cost more initially and require more care in application. V.h.f. crystals are used mainly where economy in number of stages and over-all current drain is an important consideration.

Heterodyning

Any two frequencies may be fed into a mixer stage to produce signals at the sum and difference of the two. This process is inherent in the superheterodyne receiver, but it was not widely used in transmitters until the advent of single sideband. If heterodyning is done properly, the product is an exact replica of the original signal, with no more frequency instability than was present in the two components mixed. Heterodyning is thus a good way of obtaining variable frequency control, since it is relatively easy to build a variable oscillator for the h.f. range that is adequately stable at its fundamental frequency.

The process is shown in block diagram form in Fig. 5-2. The control signal is generated at some frequency below about 10 Mc., 8 to 9 Mc. in this example. A 42-Mc. signal from a stable source beats with the control signal in the mixer. Two main products, one at 34 to 33 Mc., and the other at 50 to 51 Mc., result. The unwanted difference product is rejected by the filter, while the desired sum at 50 to 51 Mc. is passed on to succeeding amplifier stages.

The v.h.f. man who works 50 Mc. and higher bands may employ heterodyning again to reproduce the 50-Mc. signal on another band. In the example the 50-Mc. signal is mixed with one at 94 or 95 Mc., to give coverage of the lower half of the 144-Mc. band. Other crystals can be used in either crystal oscillator to extend the coverage to any one-megacycle segment desired.

The chief problem in heterodyning is to prevent unwanted products from being radiated. In our example we use the sum of 42 and 8 to 9 Mc., but the difference is also produced. The selectivity of the tuned circuits may be sufficient to reject the unwanted product, but this should not be assumed. The output of any transmitter employing heterodyning should be checked carefully to be sure that frequencies other than the intended one are not being radiated. A mixer stage requires only a very small amount of energy on the mixing frequencies to produce output, so harmonics and other components of the signals being mixed may beat with each other and produce all manner of unwanted frequencies. Mixing at low level, careful exam-

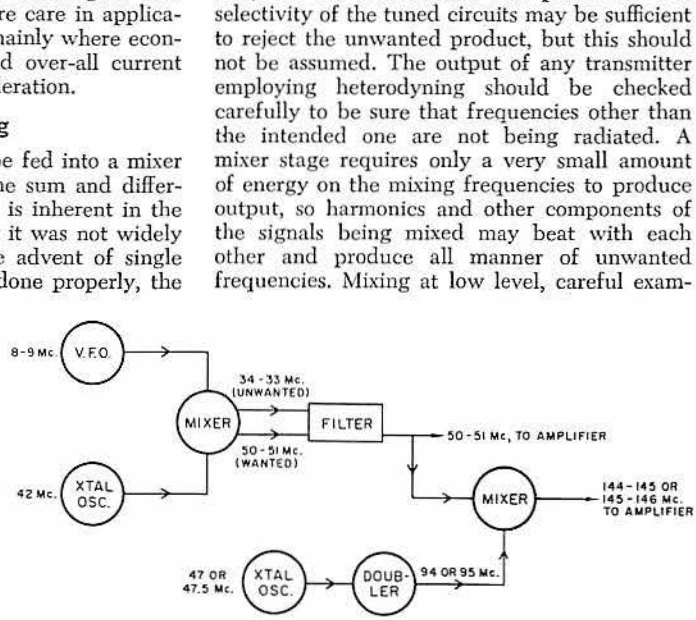


Fig. 5-2—Block diagram of a typical heterodyning process for producing stable v.f.o.-controlled signals at 50 and 144 Mc.

ination of the spectrum for spurious products, and use of highly selective circuits for passing on the desired product and rejecting others are musts for the builder of a heterodyne exciter.

Crystal Oscillators

Quartz crystals of many kinds and cuts are used for frequency control but all have one characteristic in common: when a voltage is applied across the crystal it is distorted mechanically. The converse is also true: mechanical distortion of the crystal develops a voltage across it. This is the basic piezoelectric effect, discovered many years ago and applied to crystal control of oscillators as far back as the 1920s.

The greatly magnified edge views of crystal plates in Fig. 5-3 show, in simplified form, what happens to an oscillating crystal. The quiescent state is at the far left. The next two sketches show the distortion at the positive and negative peaks of the oscillation cycle. The crystal is a very high-Q device. It will oscillate on one frequency only, determined principally by the thickness of the crystal. (The thinner the plate, the higher the frequency.) Connected properly in an oscillatory circuit, the crystal will control its frequency within very narrow limits.

Crystals and circuits for their use in v.h.f. transmitters are of two principal types: fundamental and overtone. It is important for the v.h.f. worker to understand the basic differences between them. The fundamental crystal, whose mode of operation has just been described, is usually supplied for frequencies up to about 12 Mc. Though fundamental crystals can be made for frequencies up to about 30 Mc., they are very thin and difficult to handle and process above the normal commercial limit of 12 Mc.

At higher frequencies it is customary to go to overtone oscillators. Almost any crystal can be made to oscillate on its third overtone, which is roughly three times the frequency for which the crystal was ground. In overtone operation the crystal in effect breaks up into an *odd* number of layers, as shown in the right half of Fig. 5-3. The oscillation cycle is given in the two sketches at the right. Because of mechanical considerations, the overtone may not be an exact multiple of the fundamental frequency, though it is always close to it. Only the odd multiples are available as overtones; there is no such thing as a second, fourth or sixth overtone.

Overtone operation with crystals processed for fundamental service depends on several factors, principally the flatness of the crystal and the method of mounting in the holder. Because the layers for third-overtone oscillation in an 8-Mc. crystal are less than 0.004 inch thick, and for higher-order overtones progressively thinner, it can be seen that minor variations in flatness or surface imperfections quickly inhibit overtone oscillation. Crystals clamped between metal plates, as in the common FT-243 holder, seldom work well above the third overtone.

Crystals processed for overtone operation usually can be made to oscillate on higher-order overtones than the intended frequency.[1] A crystal marked for 24 Mc., normally an 8-Mc. fundamental, will often work well on 40 Mc., 56 Mc., or even 72 Mc. in suitable circuits. Unless the purchaser specifies otherwise, crystal companies customarily supply third-overtone crystals for frequencies from about 12 to 54 Mc., fifth-overtone for 54 to 70 Mc., and seventh-overtone for frequencies up to around 100 Mc. Overtone crystals for frequencies as high as 150 Mc. can be made, but in amateur service frequencies above about 72 Mc. are seldom used for direct control.

For best stability any crystal oscillator should be run at low power input, and this is increasingly important as one goes to higher frequencies. The crystal oscillator should always be regarded as a device for controlling frequency, not as a source of r.f. power. Control of feedback is also important. However control is achieved, feedback should be at a level that will allow the oscillator to start readily, but not enough to cause heating or frequency jumping.

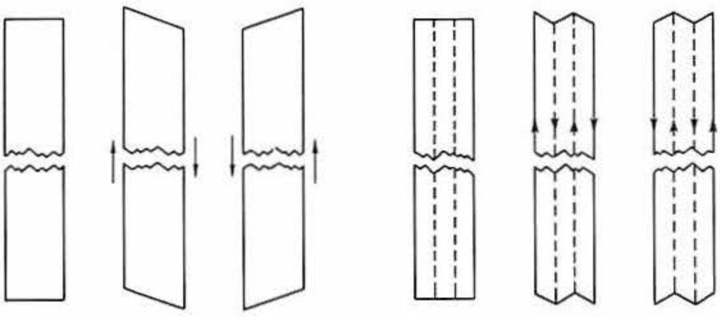


Fig. 5-3—Greatly magnified edge views of quartz crystals, showing the mechanical distortion effect when voltage is applied across the crystal. At the left is a fundamental crystal, and at the right is one oscillating on its third overtone. Frequency of oscillation depends on crystal thickness—the thinner the crystal the higher the frequency.

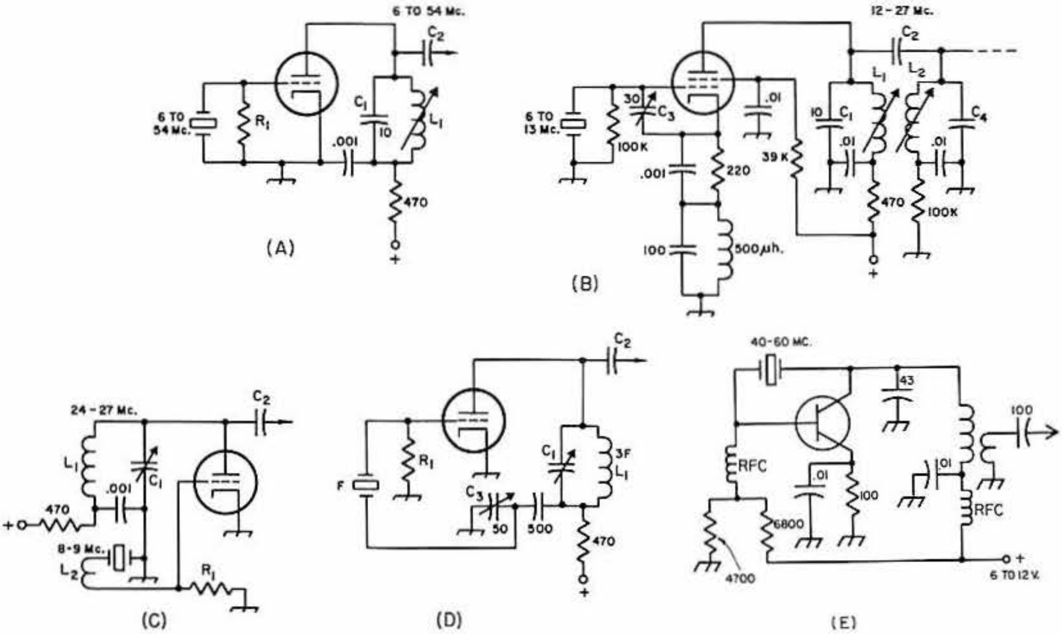


Fig. 5-4—Typical crystal oscillator circuits. The triode circuit, A, works with either fundamental or overtone crystals, up to at least 54 Mc., though circuit Q is more critical for overtone operation. A comparable transistor circuit is shown at E. A popular pentode oscillator-multiplier is shown at B. Double-tuned coupling circuit helps to select the desired harmonic and reject others. Circuits C and D can be used to make fundamental crystals oscillate on their third overtone, through control of feedback. Value of R_1 depends on the tube type; 47,000 ohms to 100,000 ohms is common. C_2 should be the lowest usable value; about 2 pf. in B, 10 pf. in others.

Crystal Oscillator Circuits

An almost infinite variety of crystal oscillator circuits may be employed in v.h.f. transmitters. Only a few will be described here, to demonstrate basic principles. These will satisfy most requirements, and though the literature contains special claims for innumerable variations, proper adjustment and operating conditions are the principal factors in achieving the desired results.

A simple triode oscillator useful for both fundamental and overtone crystals is shown in Fig. 5-4A. Feedback in this circuit is through the tube capacitance. If a broadly resonant coil were substituted for the crystal, the circuit would oscillate on a frequency determined principally by the tuned circuit, L_1C_1. With a fundamental crystal in place, the circuit may oscillate on the crystal frequency whether the tuned circuit is peaked or not, with output rising sharply as the circuit hits resonance. With an overtone crystal it may be that oscillation will occur only when the plate circuit is resonant at the desired overtone frequency.

Occasionally with overtone crystals, oscillation occurs at the fundamental frequency of the crystal at any value of C_1 except that which tunes L_1 to the crystal overtone frequency. With a 24-Mc. crystal it would then be possible to hear an 8-Mc. signal until the circuit is resonated. Then the fundamental signal disappears, and the oscillator can be heard only on the overtone frequency, 24 Mc., or multiple thereof. If the overtone oscillator is working correctly there will be no fundamental signal. If there is one, the plate circuit is not resonating at the overtone frequency, or its Q is too low to sustain overtone oscillation. A typical transistor overtone oscillator is shown at E.

Circuit B is commonly used with fundamental crystals in v.h.f. service, where *harmonic* output is desired with a minimum of stages. The control grid, cathode and screen of a pentode or tetrode tube comprise the oscillatory circuit. The cathode is raised above ground for r.f. by the choke and capacitor combination, and feedback is controlled by the variable grid-cathode capacitor, C_3. The plate circuit, L_1C_1, is tuned to the second, third or fourth harmonic of the crystal frequency; usually the third. The circuit will oscillate at all times, regardless of the tuning of the plate circuit, always on the fundamental frequency of the crystal. Output on the various harmonics (*exact* multiples of the crystal frequency) develops as the plate circuit is tuned. If C_1 is a wide-range variable it will be possible to tune adjacent harmonics. Examples would be 18 and 24 Mc. with a 6-Mc. crystal, or 16 and 24 Mc. with an 8-Mc. one. The most common use of this circuit in v.h.f. work is with 8-Mc. crystals, with the plate circuit tuning the 24- to 27-Mc. range.

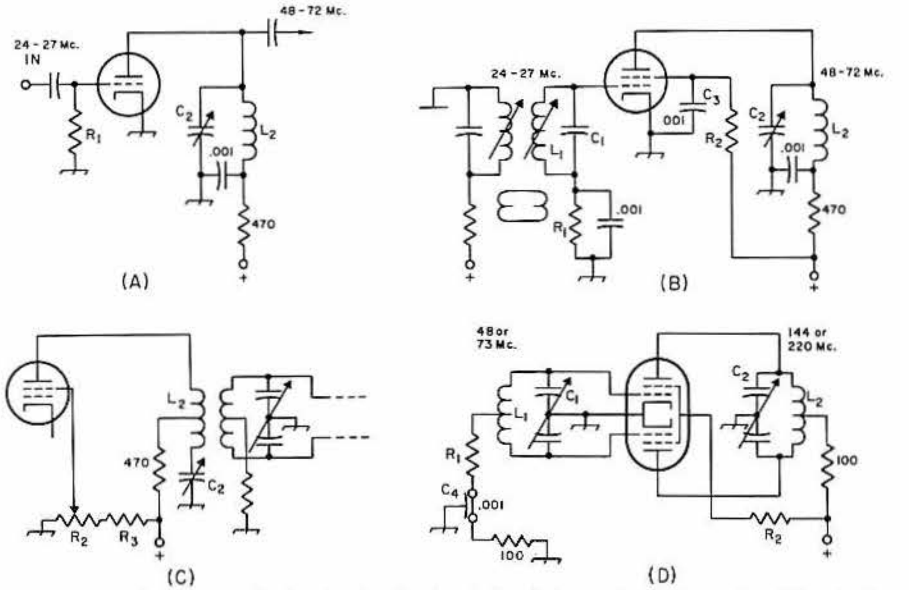


Fig. 5-5—Common frequency multiplier circuits. Simple triode, A, is popular, but provides little rejection of unwanted harmonics. Double-tuned link-coupled circuit, as in grid of B, is much better in this respect. Plate circuit may be made to cover both second and third harmonics by having wide range in the tuning capacitor. Adjustable screen-voltage control, R_2 in C, permits adjustment of drive to following stage. Series-tuned plate circuit, C, extends useful range. Push-pull tripler, D, is commonly used with dual tetrodes.

Since there is always oscillation on the fundamental frequency with this circuit, all harmonics appear in the output to some extent, and precautions must thus be taken to prevent unwanted harmonics from being passed on to succeeding amplifiers. Simple capacitive coupling, as shown in the circuits other than B, is bad in this respect, as only the selectivity of the tuned plate circuit is present to attenuate undesired frequency components. Considerably more selectivity and good coupling efficiency are available with the double-tuned circuit in B. The coupling capacitor, C_2, is very small; as little as 2 pf. Link coupling, usually with one-turn loops around the bottom ends of L_1 and L_2, may be used in place of C_2.

To make a fundamental crystal oscillate on an overtone, more feedback may be needed than is available in circuit A. The circuit of 4C is one of the more popular. Here the plate circuit, L_1C_1, tunes to the desired overtone frequency, and the crystal is connected in a feedback loop that is inductively coupled to the plate circuit. Feedback can be controlled by the number of turns in L_2, or by its position with respect to L_1, or both. Any crystal that will oscillate on its fundamental should take off on at least the third overtone when used in this circuit. A lower value of L_1 and careful adjustment of feedback, and a suitable modification of L_1C_1 usually permit fifth-overtone oscillation as well.

A capacitive feedback adjustment is shown in circuit D. The plate circuit tunes to the desired overtone frequency, and feedback is controlled by C_3. Usually this can be made a fixed capacitor, 50 pf. being customary for crystals in the 8-Mc. range. The *lower* the capacitance the greater the feedback. The circuits of Fig. 5-4 are often used with dual triodes, or triode-pentode tubes, with the second tube section operating as a frequency multiplier. Thus 8-Mc. crystals can be used with one dual tube to obtain output in the 48- to 54-Mc. range. Where stability and flexibility are important, circuit B is preferred. The overtone circuits give better protection against interference arising from unwanted harmonics, and are somewhat more economical in number of parts required.

The transistor oscillator, E, may be used for fundamental or overtone service. Note that higher capacitance is used for a given frequency, in the tuned circuit of a transistor oscillator. Value shown is for 40 to 60 Mc., usually with third-overtone crystals.

Frequency Multipliers

Frequency-multiplier stages in v.h.f. transmitters are similar to those in lower-frequency setups, except that more emphasis should be placed on keeping r.f. leads as short as possible. This is particularly important at 144 Mc. and higher. Tubes should preferably have low input and output capacitance, and high transconductance. Stages should operate at the lowest usable power level, to minimize the interference potential of any unwanted harmonics they may generate or amplify.

Typical circuits are shown in Fig. 5-5. In all of these the value of grid resistor, R_1, depends on the type of tube and the driving power available. In general, bias and drive should be as high as the grid dissipation of the tube will allow, if high efficiency is desired. The grid resistor should be such that the tube will run no more than the normal grid current, regardless of drive level.

Circuits A and B can be set up to double or triple. If C_2 is a wide-range variable (50 pf. or more) the value of L_2 can be such that the stage will triple from 24 to 72 Mc. at near minimum setting, and double to 48 to 54 Mc. with more capacitance in the circuit. This may be convenient in a combined 50- and 144-Mc. exciter, or in one where the stage will drive triplers to 144 or 220 Mc.

Circuit A uses capacitive coupling in and out. This is simple, but it does not offer much protection against unwanted harmonics from preceding stages. Double-tuned inductively-coupled circuits, as shown in the input to circuit B, are much superior in this respect. A similar output coupling arrangement may, of course, be used for better selectivity.

The plate circuit of C is helpful where it is difficult to resonate L_2 with the single-ended circuits A and B, or where inductive coupling is used between single-ended and push-pull stages. With no tuning capacitance in parallel with the output capacitance of the tube, a larger coil is usable for a given frequency than is possible with parallel-tuned circuits. This circuit works to best advantage when the value of L_2 is adjusted so that C_2 tunes at a setting roughly equal to the tube's output capacitance. This same approach is useful in tuning a grid circuit, and may be a necessity to resonate a circuit on the higher bands, with a tube having input capacitance.

The push-pull tripler circuit, D, is often used with dual tubes, both triodes and tetrodes. It is an effective tripler and has the advantage of tending to cancel out even harmonics of the driving frequency. When used with loose inductive coupling between stages, a push-pull tripler from 48 to 144 Mc. is good insurance against TVI in Channel 10, that can result from the fourth harmonics of 48 Mc. being passed on to the following amplifier, along with the desired third.

Where tetrodes or pentodes are used for frequency multipliers, the screen resistor, R_2, may be used to control the power level. This can be done by selecting a suitable value for R_2, as in B and D, or by making R_2 variable, as in C. Such a variable screen-voltage control is handy in adjusting the drive level to a succeeding stage. A limiting resistor, R_3, should be used in series with R_2 to prevent the screen voltage from exceeding the maximum rating for the tube concerned.

The screen bypass capacitor, C_3 in circuit B, can often be dispensed with in frequency multipliers. If the stage is stable without it, and it makes no difference in the output obtainable, there is no point in putting it in. A screen bypass is seldom needed in push-pull circuits in the v.h.f. range, as in circuit D.

A means of measuring grid current is usually helpful in adjusting frequency multipliers. If a grid milliammeter is not to be connected in the circuit permanently, a low-value resistor may be inserted in series with the main grid resistor, and the meter temporarily connected across it. The feedthrough capacitor, C_4 in circuit D, is often a convenient way of bringing a terminal to serve as a test point for reading grid current.

In transistor multipliers it is very important to match impedances and to use as much selectivity as possible. Otherwise there may be trouble with unwanted harmonics, to a greater extent than is normal with vacuum tube circuits.

POWER AMPLIFIERS

Principles of transmitter operation are essentially the same regardless of frequency. Since basic amplifier theory is covered adequately in the ARRL *Handbook,* modes of operation and circuit design of power amplifiers will be discussed here only insofar as the special aspects of the v.h.f. field are concerned.

Most transmitting tubes presently used by amateurs will work on 50 Mc., and inclusion of this band in a transmitter for lower frequencies is often practical. For 144 Mc. and higher, the tubes and other amplifier components must be selected for their v.h.f. qualities. Suitable tubes have low input and output capacitance, and are constructed physically so as to have the lowest possible lead inductance, both within the tube and in connection to the circuit elements. Most of the higher-powered types are designed for forced-air cooling. Some are designed to be integral parts of coaxial tank circuits. Where conventional coil-and-capacitor circuits are used, they must include the absolute minimum of stray inductance and circuit capacitance. This rules out bandswitching in the usual sense, and even plug-in coils are seldom practical above the 50-Mc. band.

Multiband operation is possible, even up through the 420-Mc. band, but special techniques are required to achieve it. Thus, with certain exceptions to be described, v.h.f. power amplifiers are usually designed to do the best possible job on a single band.

Linear or Class C?

Most v.h.f. operators start with a small transmitter, homebuilt or manufactured, usually complete with power supply and modulator. When they want more power they look for another single package that can be connected to the first unit. Such a device is called a *linear amplifier.* Adjusted properly, it will reproduce the original signal faithfully at a higher power level. When used with an a.m. phone transmitter, the linear amplifier is inherently a low-efficiency device, but it has its virtues, particularly when other modes of operation are planned.

Various routes to increased power are shown in Fig. 5-6. Our basic transmitter, A, requires only the linear amplifier, B, to run up to full legal power of 1 kilowatt. This can be a high-efficiency system on c.w., delivering up to 750 watts output. As an a.m. phone linear, its maximum power output is 350 watts. If the original transmitter includes provision for s.s.b., as well as c.w. and a.m., the linear becomes more attractive. Setup C can give full power on c.w. and s.s.b., and medium power on a.m., without auxiliary audio equipment.

Maximum power output on a.m. phone requires a high-level modulator, as in D. A 500-watt audio system, needed to modulate a kilowatt amplifier, is an expensive and bulky proposition, however, and with the current trend to s.s.b. in amateur voice communication, more and more v.h.f. men are thinking twice before making the considerable investment in terms of money, space and weight that a kilowatt a.m. phone station entails. There will undoubtedly be considerable use of a.m. in v.h.f. work for many years to come, despite the inroads of s.s.b., so the relative merits of linear and high-level modulated amplifiers deserve careful thought.

If one is to concentrate on a.m., to the exclusion of other modes, a plate-modulated power amplifier of no more than 200 to 500 watts may be desirable. The cost of both r.f. and audio components rises very rapidly above the 500-watt level, and it may well be that the extra cost could be better spent in other ways. A final amplifier of under 500 watts input, properly modulated, may be a better choice than an a.m. linear that is capable of running a full kilowatt. The former will be much easier to adjust and more flexible in operation, and it will deliver about the same power to the antenna. For the all-mode operator, the linear approach is more attractive, since only minor modification of the operating conditions will permit high-efficiency operation on c.w. and s.s.b., while retaining a.m. capability at a somewhat lower power level—all with a station that can be built compactly and at a moderate cost for the high-power portions.

A distinct advantage of the linear approach is the matter of driving power. With a Class AB_1 linear (most commonly used with v.h.f. tetrodes), no driving power is required; only voltage. Kilowatt amplifiers like the ones shown in Chapter 6 can deliver over 300 watts output with nothing more than a 1-watt a.m. rig as a driver. A little more drive will push the c.w. output to as much as 600 watts. With a driver output of 7 to 10 watts, the amplifiers will give up to 750 watts Class C output on c.w., s.s.b. or f.m.

Operating conditions for linear service are critical. The amplifier must be heavily loaded. If it uses tetrodes, the screen voltage, and preferably the bias as well, should be regulated. The drive level must be watched closely, to be certain that the amplifier is never driven into the grid-current region, if it is operated Class AB_1. An oscilloscope is practically a necessity, if true linear conditions are to be achieved and maintained. In all these respects the linear is more demanding than Class-C c.w. or plate-modulated a.m. service would be. For more on linear-amplifier operation, see Chapter 6.

Single-Ended, Parallel or Push-Pull?

On lower bands, use of two or more tubes in parallel has become almost standard practice. Often it is less expensive to use several small

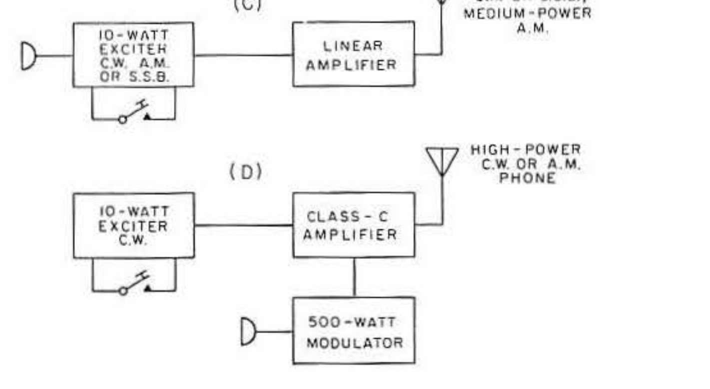


Fig. 5-6—Some ways to increase power in a v.h.f. station. Transmitter A is a typical packaged unit, complete with modulator and power supply. Adding a linear amplifier, B, can give up to 300-watts output on a.m. phone, or 750-watts output on c.w. The sideband exciter, C, usually also makes provision for c.w. and a.m., so it combines well with a linear amplifier for high power on s.s.b. or c.w., and medium power on a.m. phone. A small r.f. unit, D, is used to drive a Class-C amplifier for high-efficiency c.w. Addition of a modulator is required for high-efficiency a.m. phone.

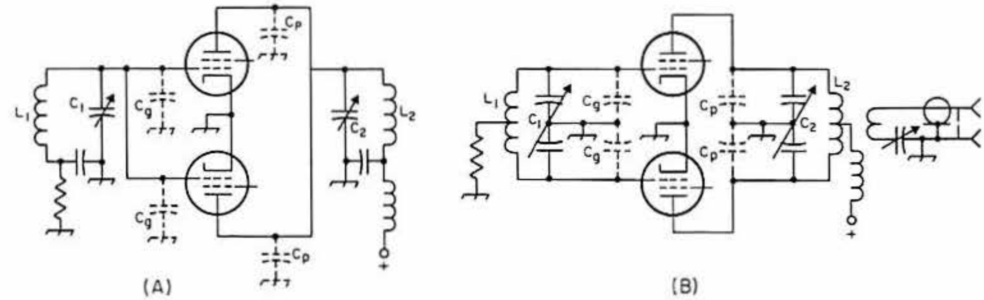


Fig. 5-7—Loading effect of input and output capacitances in single-ended circuits limits their use at higher frequencies in the v.h.f. range. In the push-pull circuit, right, these strays and also the tuning capacitances are in series across the tuned circuits, permitting use of a given tube type at much higher frequencies.

tubes in parallel than one larger one of the same power capability. Parallel is preferred to push-pull in h.f. transmitters mainly because of its ready adaptability to bandswitching. Where tube and stray circuit capacitances do not represent a large percentage of the total, parallel connection of tubes is entirely satisfactory.

Looking at Fig. 5-7A, we can see readily why parallel operation is not practical for the higher v.h.f. bands. The tubes' input and output capacitances, C_g and C_p, shown in broken lines, are in parallel with the tuning capacitors, C_1 and C_2, across the tuned circuits. Suppose we select a pair of good v.h.f. tubes like the 7034/4X150A. This tube's input capacitance is 16 pf. Thus in circuit A we have 32 pf., plus the minimum of C_1, plus unavoidable circuit capacitances, all in parallel across L_1. Output capacitance is 4.4 pf., so the plate circuit has 8.8 pf. in C_p, plus the minimum of C_2 and circuit stray capacitance, across L_2. Obviously it will not be possible to resonate conventional tuned grid and plate circuits at 144 Mc. and higher, with tubes connected in parallel, even when they are types designed for v.h.f. service.

In the push-pull circuit B, the input capacitances are in series across the tuned circuit. So are the two halves of the split-stator tuning capacitor, C_1. The effective total capacitance across the tuned circuit will be about one fourth that of the parallel connection. The same is true in the plate circuit. It can be seen that our chances for reasonably good v.h.f. circuit efficiency are vastly better with push-pull than with parallel.

With single-tube amplifiers the parallel effect of the tube and circuit capacitance still prevails, but it is not nearly so bad as with two or more tubes in parallel. Most single-ended amplifiers for the higher bands employ tank circuits which permit direct connection to the tube element or socket tab, with no leads in the usual sense. Coaxial lines or flat-strip tank circuits are preferred, especially for higher-power amplifiers. Even with the lowest possible capacitance, r.f. circulating current will run very high in a v.h.f. amplifier, so low d.c. and r.f. resistance is of utmost importance. Large conductors have the added advantage of helping to dissipate heat developed in the tube elements.

Because of their compact construction and short leads, power transistors work well in parallel up through 150 Mc., at least.

Pi-Network or Inductive Coupling?

In recent years the pi-network tank circuit, Fig. 5-8A, has become popular for transmitter use, largely because of its adaptability to band-switching amplifiers. In single-band v.h.f. designs there may be little choice between it and the inductively coupled circuit, 5-8B. The output circuit of an amplifier has two basic functions: to tune the stage to the desired frequency, and to act as a matching device between the stage's high output impedance and the low-impedance load. In the pi-network the tuning and loading capacitors, C_1 and C_2, serve these purposes. With inductively coupled circuits, either the single-ended, 5-8B, or push-pull, 5-7B, the coil and the output coupling loop comprise the matching transformer. The two circuits work equally well, and choice between them can be dictated by adaptability to the particular amplifier being built.

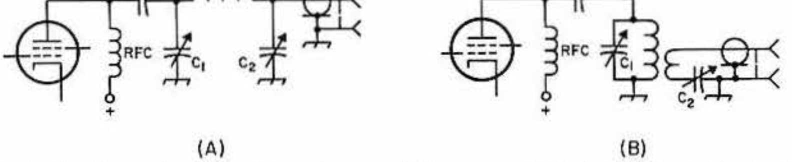


Fig. 5-8—Basic functions of tuning and impedance matching are performed equally well in the pi-network, A, and inductively coupled output circuit, B. Choice in single-band v.h.f. amplifiers is mainly a matter of convenience in a particular design.

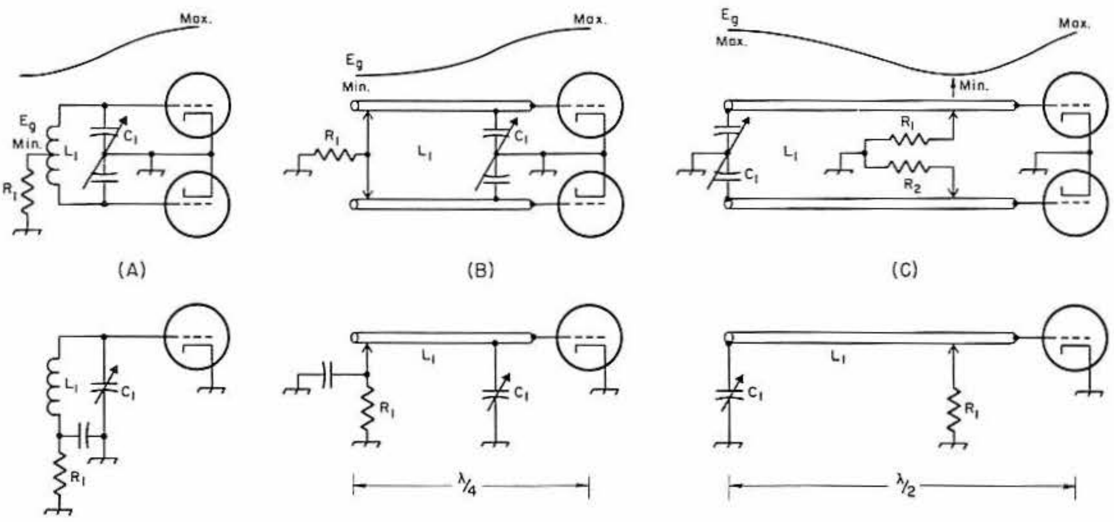


Fig. 5-9—Evolution of the tuned circuit in v.h.f. amplifiers. Conventional coil-and-capacitor tuning, A, becomes a quarter-wave line circuit in B. A half-wave circuit is shown at C. Each has a progressively higher upper useful frequency limit for a given type of tube, whether single-ended or push-pull design, is used.

Coils or Lines?

On lower frequencies the fact that any capacitor has some inductance and any coil some capacitance can be neglected in most circuit-design work, for these "strays" are too small to have any significant effect. At frequencies in the upper v.h.f. range they become all-important. Connecting leads, which at lower frequencies merely join coils and capacitors, may, in a v.h.f. circuit, have more inductance than the "coil" itself. Similarly, leads within tubes and sockets may become appreciable portions of a wavelength. Unavoidable capacitance in r.f. circuits also severely restricts the upper limit of frequency for satisfactory v.h.f. amplifier performance.

At 50 Mc. these factors are not insurmountable, if care is used in laying out amplifier stages. Single-ended or push-pull circuits such as Fig. 5-9A still work well if tubes and components designed for v.h.f. service are used. Conventional circuitry may serve at 144 and even 220 Mc. with suitable tubes, but in general the usefulness of coil-and-capacitor circuits is limited above 100 Mc.

Transmission-line adaptations of conventional tuned circuits, 5-9B, extend the range and improve performance as we reach frequencies where we "run out of coil" with the circuits at the left. In the push-pull version, the inductance L_1 may take the form of a U-shaped loop, or it can be a pair of copper pipes, ⅛ to 1 inch or more in diameter, with an adjustable shorting device at the end away from the tubes to adjust the total inductance in the circuit. The single-ended version below it can be grounded at the left end, if a blocking capacitor is used at the grid, and the resistor R_1 is connected from grid to ground. The effective electrical length of L_1 can be made variable by use of a sliding contact.

With either circuit of B the upper limit of frequency is reached when C_1 is removed and the ground point is moved up to the grid terminal. In practice, the limit is reached when there is no longer enough exposed circuit to permit effective coupling. We then can go to the circuits of C. The r.f. grid voltage E_g, is shown by the curve above each set of circuits. In A and B the zero-voltage point is at the center tap or by passed end of L_1, or at the left end of the line. If minimum r.f. voltage occurs close to the tube, the line can be extended a quarter wavelength to the left, and the tuning capacitance connected across the left end. The whole circuit, including tube and tuning capacitance, now becomes an electrical *half* wavelength of line, loaded capacitively by the tube at one end and C_1 at the other.

The bias resistors R_1 and R_2 should be connected at the point of lowest r.f. voltage in C. This can be determined by feeding r.f. power into the circuit and touching L_1 with a lead pencil or insulated metal object, until the point is found where there is no reaction on the circuit.

The half-wave line circuit will extend the useful frequency range well above the maximum obtainable with quarter-wave or coil-and-capacitor circuits. Though a grid circuit is shown in the examples of Fig. 5-9, the principle is equally applicable to plate circuits. The next steps after this, coaxial and cavity circuits, will be discussed in our u.h.f. chapter.

Multiband Amplifier Circuits

Though conventional bandswitching and plug-in coil arrangements are ineffective at 144 Mc. and higher frequencies, it is possible to build multiband tank circuits for v.h.f. transmitters. Simple adaptations of the plug-in coil idea are shown in Fig. 5-10A. Here a 144-Mc. circuit, L_1, is completed by plugging in a shorting bar at the end of the line. To use the circuit on 50 Mc. or even lower frequencies, we plug in a suitable coil, L_2. This general idea was used effectively in the plate circuit of a 4-65A amplifier for 28, 50 and 144 Mc. described some years ago in *QST*[2] and the *Handbook*.[3]

A similar principle is applied to half-wave lines in B. Again the 144-Mc. half-wave line, L_1, becomes merely the "leads" between the 50-Mc. coil, L_2, and the tuning capacitor. Remembering that the connection on a half-wave line is made at the point of lowest r.f. voltage, where it has no effect on the operation of the line, we realize that the 50-Mc. circuit of B can even be permanently connected. This has been done many times in both manufactured and home-built gear for 50 and 144 Mc., and it can be adapted to line tank circuits for still higher frequencies.

Some critical problems are involved in turning this trick for 50 and 144 Mc., especially when both the grid and plate circuits of an amplifier are operated in this way. Because of the nearly third-harmonic relationship, considerable care must be exercised in proportioning the tank circuits to prevent radiation of energy on unwanted frequencies, or oscillation troubles due to unwanted resonances in the grid and plate circuits. An example of a design in which these potential troubles were avoided was shown in an amplifier by WØIC in *QST*.[4]

By thinking in terms of the job to be done, rather than of the way such tasks have been handled in the past, it is often possible to come up with solutions that are unique to the v.h.f. field. A grid circuit tuning both 144 and 220 Mc. made possible an efficient transmitter for these bands, in which only the plate circuit was changed. A completely removable plate circuit also brought 432 Mc. into the picture, permitting use of the stage as a doubler or tripler. This design by W1VLH appears in Edition 1 of this Manual.

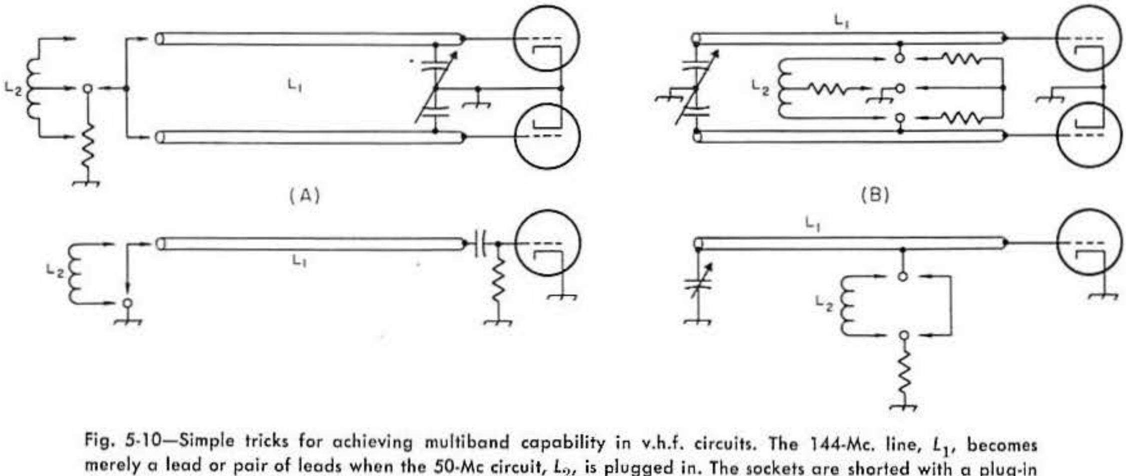


Fig. 5-10—Simple tricks for achieving multiband capability in v.h.f. circuits. The 144-Mc. line, L_1, becomes merely a lead or pair of leads when the 50-Mc circuit, L_2, is plugged in. The sockets are shorted with a plug-in device in three of the circuits. Those of B are half-wave line circuits for 144, with the 50-Mc. coil plugged in at the point of lowest r.f. voltage.

USING FREQUENCY MODULATION

Though it is almost standard for commercial mobile services in the v.h.f. range, frequency modulation has not been widely used by amateurs, mainly because of receiver problems discussed in Chapter 3. In the transmitting side of the picture the advantages of f.m. are many and impressive:

1. Elimination of high-level amplitude modulation effects great savings in transmitter cost, size and weight. The modulation equipment for f.m. is the same regardless of transmitter power level. Frequency modulation can be added to a v.h.f. transmitter with very little added expense or complication.

2. TVI is greatly reduced with f.m. Audio-caused interference, which accounts for a high percentage of all our TVI and BCI trouble is practically eliminated. Particularly in densely-populated urban areas, f.m. often permits v.h.f. work during peak viewing hours, when operating with a.m. would result in intolerable neighbor problems.

3. Transmitter adjustment is greatly simplified. Grid drive level is important only as it

affects output. The f.m. transmitter can be run Class AB_1 or AB_2 with a very small exciter with only a slight reduction in efficiency from that obtainable with Class-C conditions requiring much more driver power.

4. Peak voltage does not change with modulation. Smaller components can be used than with a.m. of comparable power level.

5. Reducing power is a simple matter with f.m. Much v.h.f. communication can be done at low power levels, and FCC regulations require use of only enough power to carry on satisfactory communication, but significant power reduction is not easy with plate-modulated a.m. transmitters.

6. The varactor multiplier, now coming into more general use at 420 Mc. and higher frequencies, is ideal for use with f.m. transmitters.

How Much Deviation?

Choice of frequency deviation depends mainly on the receivers in use at stations we wish to work. In parts of bands where a.m. and s.s.b. are the principal voice modes we almost have to use narrow-band f.m., as most receivers will be incapable of handling anything wider. Below 52.5 Mc., f.m. is restricted to the narrow-band variety by FCC Regulations, as well.

Above 52.5 Mc. wide-band f.m. is legal, but practical considerations will determine whether it is advisable or useful. In general, wide-band f.m. (any deviation beyond that capable of being copied on a selective communications receiver) should be used only in lesser-occupied segments of any band: above 52.5 Mc., 146 or 221 Mc., and between 420 and 432 or 436 and 450 Mc. Narrow-band f.m. is acceptable wherever voice is used, since its bandwidth, by definition, is no more than required for a properly-modulated a.m. signal.

Beyond these considerations, deviation may have to be adjusted for the receiver at the other end, almost on an individual basis, because of the considerable variation in receiver selectivity characteristics. For more on this, see Chapter 3. To get the most out of f.m., regardless of the bandwidth, it is important to maintain a high *average* deviation. This has the same effect as high average modulation percentage with a.m.; the net result is a high audio level at the receiver, and best signal-to-noise ratio. Some form of audio level control is thus a must for optimum results in transmitting with f.m.

Restricted frequency response in the speech equipment is also very helpful, since frequencies above 2500 or 3000 cycles contribute little or nothing to the intelligibility of voice modulation. Any unnecessary audio bandwidth results only in unnecessary signal bandwidth, and sideband interference. Elimination of unnecessary highs from the speech equipment response is particularly important with narrow-band f.m. in crowded voice bands.

Generating Frequency or Phase Modulation

Narrow-band f.m. (or p.m.) is very easily produced when the order of frequency multiplication is high. It is customary to operate the oscillator on 8 or 6 Mc., so the oscillator frequency and deviation are multiplied 18 or 24 times, respectively, to reach 144 Mc., 27 or 36 times for 220 Mc., and 36 or 54 times for 432 Mc. For satisfactory f.m. it is necessary, therefore, to swing the oscillator frequency only about 150, 100 or 50 cycles, respectively, for n.f.m. at 144, 220 or 432 Mc. This is readily done, even with crystal oscillators.[5] At 50 Mc., with a multiplication of only 6 or 8 times, adequate deviation is somewhat more of a problem, except with self-controlled oscillators.

There are many ways to frequency- or phase-modulate a v.h.f. transmitter. With v.f.o. control the problem may be to avoid modulating it too much! Probably the simplest f.m. system involves only swinging the plate voltage of a triode or the screen voltage of a pentode a small amount. An example is shown in the VXO described early in Chapter 6. Enough deviation for good-quality n.f.m. is obtainable in this way for 144 Mc. and higher bands, but it is marginal at 50 Mc. The same procedure with any self-controlled oscillator will give plenty of deviation for 50-Mc. work. It should be used only with crystal control for any higher band, unless exceptional stability precautions are taken.

The variable-capacitance diode (varactor) makes possible a very simple and effective frequency modulator, as shown in Fig. 5-11. When back-biased, the varactor has the property of changing capacitance in relation to the voltage applied across it. The varactor may thus be connected across the tuned circuit of a variable oscillator, and audio voltage applied to it will result in proportional changes in capacitance across the circuit, and linear frequency modulation. There is more on varactors in Chapter 10.

The circuit of Fig. 5-11 also works with crystal controlled oscillators, though shift obtainable is mainly phase- rather than frequency-modulation, because of the very high Q of the crystal. This necessitates some cutting of the speech highs, to give a flat response at

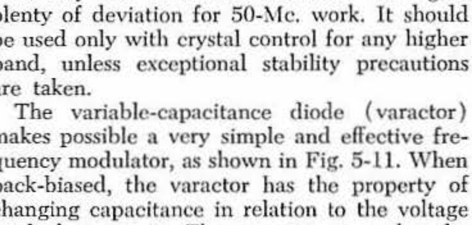


Fig. 5-11—The "simplest modulator." A single tube connected as a cathode follower applies varying audio voltage to a varactor diode, CR_1, causing its capacitance to vary proportionately. This varying capacitance is across the tuned circuit or crystal in an oscillator, resulting in frequency modulation.

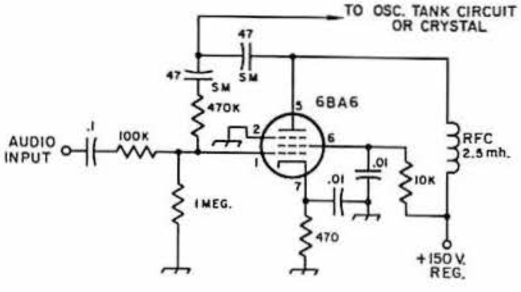


Fig. 5-12—Reactance modulator for f.m. Operation is similar, in effect, to the varactor device of Fig. 5-11.

the receiving end. Enough frequency shift is possible with 8-Mc. crystals to provide narrow-band f.m. at 144 Mc. and higher, but at 50 Mc. a self-controlled oscillator must be used for true f.m. with this method.

The reactance modulator, Fig. 5-12, has been used for many years in f.m. work. Its net effect is similar to that of the varactor modulator, and the same limitations apply. With either of these circuits, any voltage change causes a frequency change, so the supply voltage should be regulated. It should also be very well-filtered, as any a.c. voltage will cause frequency modulation. Particularly when the signal is received by slope detection, a.c. hum stands out to a very marked degree. The same hum level might be almost inaudible when true f.m. detection is used.

Where these methods are used for f.m. it is well to leave the modulator connected at all times, with its heater and plate supply left on, whether f.m. is in use at the moment or not. Merely turn the audio gain down, to eliminate f.m. Removing plate voltage, or turning off the heater of the reactance modulator, will result in some frequency change. Where a deviation control is used in the reactance modulator grid circuit it should be isolated for d.c. by a coupling capacitor, as shown in Fig. 5-12; otherwise changes in gain setting will cause frequency shifts. The two coupling capacitors (47 pf. in Fig. 5-12) should be silver-mica or other good quality, as any capacitance change there will affect frequency.

Checking Deviation

The usual methods for checking modulation do not apply with f.m. One advantage with f.m., however, is that the transmitter power level has no bearing on the modulation quality, or on the audio level required. Thus it is possible to set up an f.m. system for the desired deviation without ever putting it on the air. You merely listen to the desired harmonic of the oscillator, and make the necessary adjustments as to audio level and deviation for the intended bandwidth. So long as the harmonic can be heard at the final operating frequency, the audio will sound exactly the same no matter how many stages are turned off. Only a selective receiver, and preferably an audio oscillator, are needed.

Where the oscillator plate or screen is to be modulated, as in the VXO of Chapter 6, temporary provision can be made to vary the d.c. voltage a small amount each side of the center value. Listen to the signal on the band where operation is intended, and note the frequency with normal screen voltage. Now vary the screen voltage in equal amounts above and below the center value. Check the frequency changes, either by direct dial reading, or by comparison of the beat note with the tone of a calibrated audio oscillator. If 3000 cycles above and below the center frequency can be reached with equal voltage change each way the modulation should be linear with voice. This assumes that the speech amplifier response is restricted to about 3000 cycles maximum, of course.

With the reactance-tube modulator a similar check can be made by applying an adjustable d.c. voltage to the modulator grid and noting oscillator frequency change. With the varactor modulator the d.c. on the diode can be varied in the same way.

Deviation can be checked with an audio tone and a selective receiver. With the receiver a.v.c. off and the beat oscillator on, tune in the signal and adjust the b.f.o. for a convenient pitch. Now apply a 2500-cycle tone through the speech amplifier, turning up the gain until a small amount of modulation is observed. Using high selectivity, tune the receiver both sides of the signal frequency. There will be sidebands 2500 cycles away from the carrier on each side. At low audio input these will be the only ones audible.

Now increase the gain or audio input, and tune about 10 kc. either side of the carrier. A second pair of sidebands at 5 kc. either side will eventually appear. The audio level at which these become detectable is the maximum that should be used. If the signal is frequency-modulated only (no p.m.) the linearity can be checked by listening to the carrier beat note as the audio level is raised. If there is a shift in average beat-note frequency before the second set of sidebands appears, they may be the result of modulator nonlinearity, rather than excessive modulation. The modulator is incapable of shifting the frequency over a wide enough range if the center-frequency shift appears before the second set of sidebands.

The transmitter output should be checked for evidence of amplitude modulation, especially in the 50-Mc. band, for simultaneous a.m. and f.m. is illegal in this band. It is permissible on higher bands, but that does not make it right to have it. Any movement of a plate, screen or grid meter in the r.f. stages of the transmitter as the result of modulation is a sure sign of trouble, as is any vestige of variation in transmitter output under modulation. Both are indications of modulator nonlinearity, and resultant widening of the channel occupied by the signal that serves no useful purpose.

About Wide-Band F.M.

Though we have been concerned mainly with narrow-band f.m. thus far, most of what has been said applies to wide-band as well. For amateur purposes, we can define the latter as any f.m. with deviation wider than can be received intelligibly on a receiver adjusted for the highest selectivity that will accept properly-modulated a.m. signals. All wide-band f.m., whether the deviation is 15 kc., as in the outmoded commercial equipment now being put to use in amateur v.h.f. communication, 25 kc., as in TV sound channels, or 75 kc., as in f.m. broadcasting, will require special receiving equipment. Wide-band f.m. of whatever deviation has real merit, and is deserving of more attention from v.h.f. men than it has received in the past.

The six numbered advantages listed earlier for f.m. apply equally to wide or narrow deviation. In addition, wide-band f.m. is adaptable to quite simple transmitters, and it is capable of signal quality and freedom from noise of all kinds that is unequalled by other voice modes. Receiver considerations are covered in Chapter 3. Let's look at the transmitter angles.

First, there are the transmitters made obsolete for their intended purpose by the narrower channels now imposed on commercial f.m. users. There is nothing wrong with this gear for amateur communication, and its availability in large quantities at low cost has boosted amateur interest in wide-band f.m. greatly in recent years.[6] Little work is required to adapt this gear for amateur purposes, so it need not be discussed in detail here.

The equipment is for three principal bands: 30 to 50 Mc., 150 to 160 Mc. and 460 Mc. and higher. Both transmitter and receiver are crystal-controlled, so operation is necessarily on fixed channels. Conversion to amateur frequencies 52.5 to 54 Mc., the upper portion of the 144-Mc. band, or the 420-Mc. band, consists mainly of substitution of suitable crystals, and retuning to the amateur band. The deviation is plus or minus 15 kc.

Wider deviations can be used to advantage in amateur work, particularly on the higher bands where occupancy is relatively low and suitable receiving equipment is available. Most wide-band f.m. other than that with the fixed-frequency gear mentioned above has used deviations roughly the same as in f.m. broadcasting: plus or minus 75 kc. The ease of adapting TV receiver sound i.f. systems to amateur needs also raises the possibility of using 25-kc. deviation. Either 15- or 25-kc. deviation is practical with crystal control in the 220- or 420-Mc. bands, but for 52.5 or 144 Mc. it is more readily achieved with self-controlled oscillators. For wider deviation, as in f.m. broadcasting, the self-controlled oscillator is almost a necessity below 420 Mc., except with fairly complex equipment.

Wide-band f.m. is very easily achieved with any v.h.f. transmitter that is v.f.o. controlled, using the methods already described. Stability requirements are not so severe as for n.f.m. Only reasonable precautions as to mechanical and electrical stability need be taken. Even very simple oscillator or oscillator-amplifier transmitters can be pressed into service where the deviation is to be 75 kc. or more. The principal considerations in transmitter design are adequate power supply filtering, and some provision for keeping the average deviation close to the maximum permissible for the receivers in use at the stations to be worked.

Wide-Band F.M. with Simple Gear

It was mentioned earlier that modulating the oscillator of a v.h.f. transmitter is a simple way of obtaining f.m. This principle can be applied to simple oscillator-type transmitters in which there is no frequency multiplication. The stability obtainable with such simple equipment is not sufficient for use in heavily-occupied bands, but it could very well serve for good work on 220 or 420 Mc.

Receivers should have bandwidth like that of f.m. broadcast receivers and i.f. systems: 150 kc. Transmitter deviation of plus or minus 75 kc. at 220 Mc. is obtained with something on the order of 8 to 10 percent modulation, applied to the oscillator. The oscillator can feed the antenna directly, or drive a following amplifier. The latter is preferred from the standpoint of stability, but a simple oscillator rig can be made to sound better than you might think. Examples of the oscillator-amplifier approach, built and extensively demonstrated by W1CTW, were described by him in *QST* [7].

The author of this book demonstrated an even simpler arrangement for 420-Mc. work at many radio clubs and conventions years ago. A little 6J6 oscillator was modulated by a 6AQ5 audio stage.[8] With speech input held so low that the modulation percentage was only about 5 per cent, the signal could be received with quite satisfactory quality with a simple tunable converter and an f.m. broadcast receiver. In bands where activity is presently low, we could very well make use of such elementary gear for local communication today. The signals sound more like a buzz-saw than speech, when picked up on a selective communications receiver, but with wideband f.m. detection they can be above reproach. We have plenty of room for them above 220.5 Mc., and almost anywhere in our 420-Mc. band.

For those who want something better, wideband f.m. can readily be applied to any v.h.f. or u.h.f. transmitter. It is so easy that there is almost no reason for not giving it a try. You don't have to do it the simple way, but simplicity *does* work. With more advanced gear, where the oscillator is on a lower frequency, the results are about the closest thing to "broadcast quality" that can be found in amateur radio today.

USING TRANSISTORS IN V.H.F. TRANSMITTERS

Many inexpensive transistors work very well in low-powered v.h.f. transmitters. Overall efficiency and battery economy have made solid-state devices supreme in the field of lightweight portable transmitters. At power levels up to about one watt output, operation from small dry batteries is entirely practical. A 100-milliwatt station for 50 or 144 Mc. may weigh as little as two pounds, ready to go on the air, and yet be capable of surprizing coverage if used with a good antenna.

Transistors are taking over in mobile v.h.f. transmitters also, effecting great battery economy over tube-equipped gear. Maximum power presently obtainable from a 12-volt system with low-cost transistors is about two watts, but more is in prospect. At higher power levels the mobile v.h.f. station may employ a solid-state exciter, with perhaps a single-tube amplifier. If the latter is the quick-heating type it may be practical to run considerable power in the mobile station when it is needed, and revert to the transistorized portion when conditions make its low power usable.

Where amplitude modulation is employed the transistors in the modulated stage should be capable of withstanding at least four times the supply voltage. With f.m. there is no comparable peak-voltage problem. In small hand-carried portables it may be desirable to keep the battery voltage to a maximum of 9, in order to use more of the popular low-cost transistors.

An extraordinary "plus" with transistors is their ability to perform well with declining battery voltage. There being no cathode to be kept hot, the transistor continues to work normally after a battery voltage drop that would put a vacuum tube transmitter out of business. If it is well-designed the transistor transmitter remains usable after as much as a 35 percent drop in voltage. Its output drops off, but signal quality remains essentially unimpaired. In the intermittent service that is characteristic of amateur v.h.f. operation the recuperative powers of the common dry batteries keep them usable for long periods.[9]

Amplitude modulation of transistor amplifiers involves some special problems. It is often necessary to modulate one or more stages preceding the final amplifier, to develop satisfactory modulation percentage and quality. In other conditions the output stage may have to be detuned slightly from the point at which maximum output is obtained. Satisfactory amplitude modulation involves many factors, and any a.m. transmitter should be checked out carefully in its design and adjustment phases.

The low impedances encountered with bipolar transistors call for much lower L/C ratios than are normal for tubes. The base and emitter usually operate at very low impedance, and customarily are tapped down on their associated tuned circuits. This is important in the design of interstage coupling circuits, if selectivity is to be retained and the amplification of unwanted frequencies is to be prevented.

The new amateur who is "brought up on" transistors will not think anything of the differences between solid-state and vacuum tube technology, but the tube-oriented old-timer may experience some difficulty in making the conversion. The author, who had to make this transition, found it helpful to keep in mind the similarities and differences between the tube and the transistor, when designing, building and trouble-shooting transistor equipment.

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4—Maer, "The Perseids Powerhouse," October, 1959, *QST*.

5—Southworth, "Phase-Modulation Exciter for the V.h.f. Man," August, 1954, *QST*.

6—Aagaard, "Noise-Free Communication with 2-Meter F.M.," July, 1960, *QST*. Same author has made available conversion information for Motorola equipment in *Wide-Band F.M. for the Amateur,* price $1.75. James S. Aagaard, K9OJV, Dept. of Electrical Engineering, Northwestern University, Evanstown, Ill.

Much helpful information on Motorola f.m. gear may be found in "F.M. Schematic Digest," put out by Two-Way Radio Engineers, Inc., 1100 Tremont St., Boston, Mass. Price $3.95.

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Chapter 6

V.H.F. Transmitters and Exciters

As was done with the subject of receiving in two previous chapters, we are covering transmitter design, adjustment and operation in detail in one chapter, and practical examples in another. In the descriptive items to follow, explanatory material and adjustment procedure will be held to the minimum necessary for adequate coverage of each unit described. The reader is urged to examine Chapter 5 thoroughly before embarking on the construction of equipment that follows.

This section will deal mainly with the r.f. portions of transmitters. Some items are coordinated in design with units that appear elsewhere in the book, and where this is the case the companion items will be pointed out. Power supplies and modulation equipment are seldom included, as these usually follow practice that changes only slightly over periods of many years. The reader is referred to appropriate sections of the ARRL *Handbook* for details of accessories that are needed in these fields.

Some equipment for the 420-Mc. band will be found herein, despite the designation of 30 to 300 Mc. semantically as "v.h.f." Where the design techniques involved are truly u.h.f. in nature, items for the 420-Mc. band will be found in the chapter dealing with u.h.f. and microwaves, later in the book.

VARIABLE FREQUENCY CONTROL

Most of the transmitters in this book are shown with crystal control. There is good reason for this. Though being able to move around at will is becoming almost as important in v.h.f. work as on lower bands, the fact remains that many variable frequency-control systems presently in use above 50 Mc. are far from satisfactory. In the case of 50-Mc. operation, some are downright illegal. From 144 Mc. up we are not required by law to transmit stable signals, but self respect and consideration for others dictate that we keep our signals above reproach, regardless of frequency.

This is not easy when continuously variable frequency control is used, especially at 144 Mc. and higher. A v.f.o. that sounds good enough in the 8-Mc. region may be only fair at its 6th harmonic in the 50-Mc. band. At the 18th harmonic, 144 Mc., it very likely will be unacceptable to the critical ear. By the time we multiply 54 times, to 432 Mc., even average crystal control is not good enough for narrow-band work.

There are two solutions: heterodyning, which duplicates the fundamental-frequency stability on a higher frequency, and very special attention to the stability problem in oscillators that are to be followed by one or more frequency multipliers. An example of the latter approach is detailed below.

A VXO FOR 50 THROUGH 450 MC.

Crystal control has many advantages. By the very nature of the quartz crystal, the frequency of a crystal oscillator is maintained very close to the desired spot. The effects of heating (expansion and contraction of tube and circuit elements), mechanical vibration and variations in heater and plate supply voltages are greatly reduced, in comparison with these effects in any self-controlled oscillator. But even with crystal control, the fundamental requirements must be met if we are to have highly stable control of frequency. These become more stringent as the order of frequency multiplication is increased.

It is possible to "pull" the frequency of a crystal oscillator a small amount in several ways. A mechanical method is described in Chapter 13, but it is adapted to use only with pressure-mounted crystals. Controlled voltage variation causes some shift, but is usually associated with large changes in output. Adding capacitance across the crystal works well with some crystals, and the swing with a given amount of capacity change can be increased by adding inductance in series with the crystal. The frequency change with these methods (as with any other) is limited by the amount of instability you are willing to accept.

The variable crystal oscillator (VXO) shown in Figs. 6-1 through 6-4 allows the operator a choice of variable capacitance alone, or in conjunction with a series coil. Furthermore, the amount of inductance in series with the crystal, and consequently the frequency shift obtained by rotating the variable capacitor, can be adjusted to suit the builder's desires. Since temperature variation is the principal cause of drift in crystal oscillators, this one is run at low input, and drift is held to a very small amount, even from a cold start. The oscillator runs continuously, so there is no heating and cooling cycle effect in transmitting.

With just variable capacitance (no series coil) the maximum usable swing is roughly 750 cycles for each megacycle of crystal fundamental frequency. A 6-Mc. crystal moves about 4 kc.; an 8-Mc. one about 6 kc., for crystals in the small

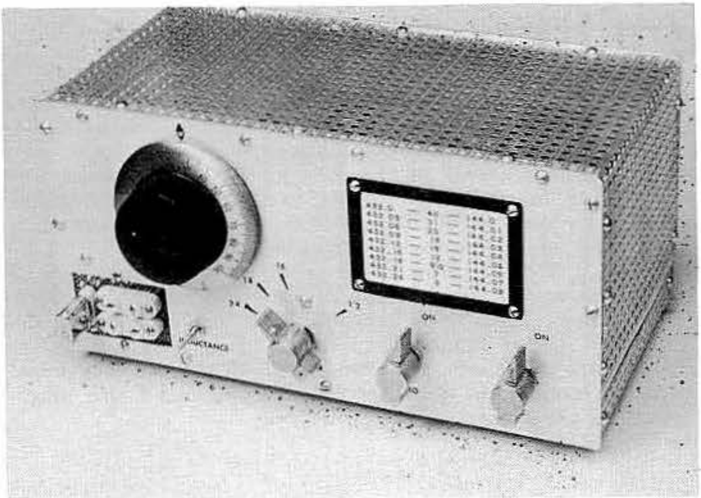


Fig. 6-1—A VXO especially for v.h.f. men. Calibration on the front panel is for a favorite crystal used for c.w. work on 144 and 432 Mc. Crystal sockets at the lower left are mounted on insulating material, to reduce circuit capacitance to the lowest possible value. Frequency variation per crystal depends on which socket is used. Pointer knobs are for the output plate circuit and the spotting and power switches.

The vernier dial is a National type AM.

metal hermetically-sealed holders. FT-243s and other pressure-mounted crystals having high holder capacitance may swing quite a bit less. There is a certain *total* capacitance at which each crystal goes out of oscillation, and it varies markedly from one to another, depending on crystal activity and mounting methods.

The 6000-kc. crystal shown plugged into the VXO in Fig. 6-1 covers 432.24 to 432.0 Mc. and 144.08 to 144.0, without use of the series coil. This gives all the coverage usually needed for weak-signal c.w. work in these two bands, but it goes lower than 6000 kc., and if it had been ordered for a shade higher frequency it would have been more useful for 144-Mc. service. About 6001 or 6002 would have been ideal. Available swing is partly on the low side of the marked frequency.

With the series coil, L_1 in Fig. 6-2, about three times as much variation is possible without serious degrading of the stability. This means 100 kc. per crystal at 50 Mc. and 300 kc. at 144, with average 6-Mc. crystals. 8-Mc. crystals do about as well. Rubberiness varies considerably when the series coil is used, as without it. One ordinary FT-243 8.3-Mc. crystal tested will swing from 50.34 down to almost the low end of the band before it becomes too unstable or drops out of oscillation, but this is exceptional. The need for variable control is confined mainly to the low part of the band, and three or four crystals will do the job for most 6-meter men. Random spots higher in the band need not be swingers, ordinarily.

On two meters, precise control at the low end, and in the region just above 145 Mc., is nice, but other segments are well served with fixed or small-swing frequencies. Operation on 220 and 420 is almost always channeled to one narrow segment of the band, easily handled with one crystal for each, even when the high-stability circuit is used, as it should be, for best results on the two higher bands.

Circuit Details

The oscillator, V_1, is a 6AK5, but almost any small receiving r.f. pentode will do. The frequency is pulled by the split-stator capacitor, C_1, connected between plate and screen. The oscillator plate voltage is regulated 150. Input is held to about 3 ma., combined plate and screen, so this oscillator is not going to move much unless you move it with C_1. An r.f. choke is used in the plate circuit, instead of a resonant coil, as tuning here would tend to pull the frequency.

To build up the very-low oscillator output to a usable level, and to provide isolation, a buffer amplifier follows, using the pentode section of a 6U8, V_{2A}. This tube was selected because it has the lowest grid-plate capacitance of any dual tube of the pentode-triode class. The triode portion V_{2B}, is a multiplier, the output frequency depending on the crystals used. Provision is made for covering 12 to 26 Mc. with C_2L_3. The plate circuit of the pentode amplifier is broadly tuned, and an intermediate setting of the slug in L_2 can be found that will permit use of either 6- or 8-Mc. crystals in the oscillator. The plate circuit of the multiplier may then be tuned to the second, third or fourth harmonic of 6 Mc., or to the second or third of 8 Mc. Which output frequency you use may depend on the type of circuit into which the VXO works. More on this later.

Construction

Mechanical layout of the oscillator portion was dictated by the need to keep circuit capacitance to a minimum. The lower the total capacitance in the circuit, the higher the frequency will go with C_1 at minimum, and the wider swing you'll achieve per crystal. This rules out crystal switching, though if convenience outranks crystal economy in your objectives, switching can be used. Crystal sockets are mounted on a Plexiglas

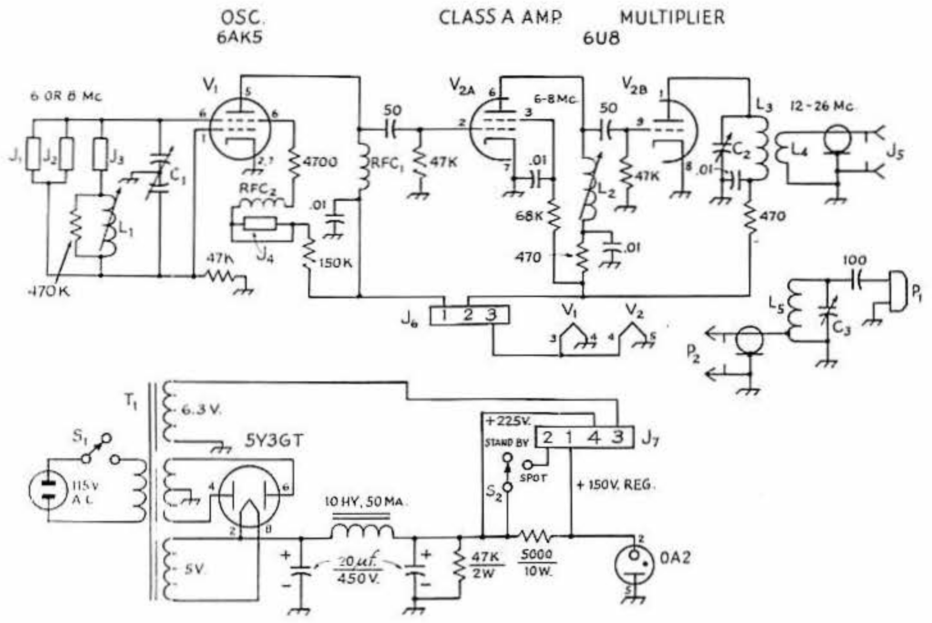


Fig. 6-2—Schematic diagram and parts information for the VXO and power supply. Unless specified, resistors are ½ watt. Decimal values of capacitance are in μf.; others in pf. Capacitors with polarity marked are electrolytic. Terminal strips J_6 and J_7 may be omitted and connections made directly where the power supply is built in. Pin 4 of J_7 permits use of the supply for other purposes.

C_1—100-pf. per-section split-stator variable (Hammarlund HF100-D).
C_2, C_3—50-pf. miniature variable (Hammarlund HF-50). Higher maximum capacitance (HF-100) may be used. Grounded-rotor type preferred.
J_1—Crystal socket for 0.05-inch pins, spaced 0.487 inch.
J_2, J_3—Crystal socket for 0.095-inch pins, spaced 0.487 inch.
J_4—2-terminal barrier strip. Omit if f.m. is not to be used. Remove jumper when f.m. is connected.
J_5—Coaxial receptacle.
J_6—3-terminal barrier strip.
J_7—4-terminal barrier strip.
L_1—16–24-μh., iron slug, ceramic form (Miller 4507).
L_2—24-35-μh., iron slug, ceramic form (Miller 4508).
L_3—3.5 μh., 21 turns No. 24 tinned, ½-inch diam., 32 t.p.i.
L_4—3 turns like L_3, spaced 1 turn from it. Make both from single piece of B&W Miniductor No. 3004.
L_5—Same as L_3, but tapped at 3 turns. Coax from L_5 to P_2 may be any convenient length.
P_1—300-ohm line plug.
P_2—Coaxial cable fitting.
RFC_1—750-μh. r.f. choke.
RFC_2—1.0 mh. r.f. choke.
S_1—S.p.s.t. switch.
S_2—S.p.s.t. switch. (See text).
T_1—Power transformer capable of delivering 200 to 250 volts d.c. at 50 ma. through filter, 6.3 volts a.c. at 1 amp. and 5 volts a.c. at 3 amp.

insert in the front panel, instead of directly on the metal. The tuning capacitor is shimmed up an extra quarter inch above the chassis, to hold down its minimum capacitance, and r.f. leads through the chassis have half-inch clearance holes. Any one of these steps yields little, but combined they net quite a few more kilocycles coverage at 432 Mc. This dividend is at the low-C end of the range of C_1, where oscillator stability is at its best.

Three crystal sockets, J_1, J_2 and J_3, are wired so that a crystal may be plugged into the circuit either with or without the series coil. Two different types of sockets in parallel, J_1 and J_2, permit small-pin or large-pin crystals to be plugged into the high-stability low-swing portion of the circuit. Use a wider variety of sockets if your crystal stock requires it, though each one adds a little capacitance.

Any crystal you plug into this circuit will oscillate on its fundamental, including those intended for overtone operation. Most crystals above 12 Mc. are overtone types, the third overtone being used up to about 54 Mc. A crystal marked for 24 Mc. will oscillate near 8 Mc., but not necessarily at exactly one-third the marked frequency. If you're ordering crystals especially for this purpose, we recommend 6 to 6.5 Mc., which will cover 50 to 52 Mc., 144 to 148, 220 to 225 and 432 to 450 Mc. The output frequency would then be, preferably, 12 to 13 Mc., as this will allow the crystal oscillator stage of most v.h.f. transmitters to work as a frequency multiplier when driven by the VXO. Use of 8-Mc. crystals and 24-Mc. VXO output is usually satisfactory where the first stage in the transmitter proper is a pentode, but triodes may self-oscillate, unless operated as multipliers.

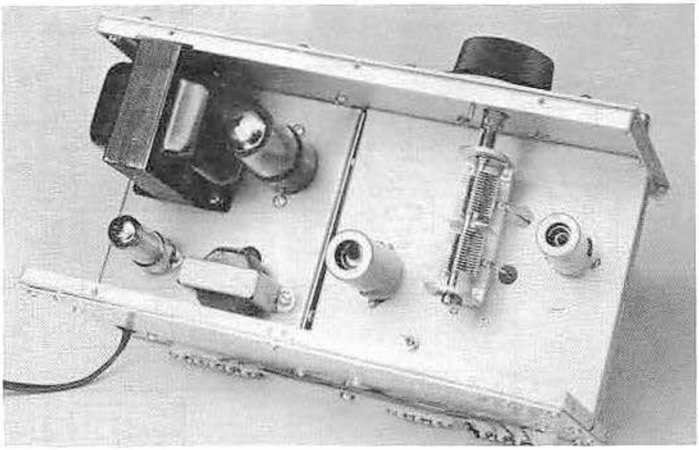


Fig. 6-3—Interior view of the VXO. The oscillator tube is at the right. The power supply, shown here as a separate assembly, could be built on the same chassis with the r.f. circuits, if the constructor wishes.

Coupling to the Transmitter

The coupling system shown in Figs. 6-5 and 6-6 is not the simplest way of hooking the VXO to a transmitter, but it has certain advantages. Low-impedance coupling terminated in the tuned circuit, L_5C_3, permits use of any convenient separation between the VXO and the transmitter. R.f. from the plug, P_1, can be fed into the transmitter in several ways. Some experimenting may be needed with your setup, but typical circuits are shown in Fig. 6-6.

Triode Overtone Oscillators. Don't try to plug directly into the crystal socket without modifying the circuit. Mounting an extra socket, J_2 in Fig. 6-6A, allows you to return to direct crystal control at will, yet gives optimum transfer of power from the VXO. Remove the regular crystal from J_1 when the VXO is used, of course. With the capacitive feedback circuit, Fig. 6-6B, the 50-pf. capacitor should be shorted out, and the VXO output fed to J_1.

Pentode Oscillators. The pentode crystal oscillator circuit used in many v.h.f. transmitters should have its cathode r.f. choke shorted by means of a switch. So should the 6CX8 pentode oscillator used in the transmitters that follow, and the similar circuits in the ARRL *Handbook.* Plugging into the crystal socket may work with such circuits, but more reliable operation is likely when the cathode is grounded for r.f., as with S_1 in Fig. 6-6C.

Another possibility in working into an existing transmitter is to disable the transmitter crystal oscillator, and couple into the grid of the second stage from the VXO. Opening the screen or B-plus lead of the first stage, as is done with S_2 in Fig. 6-6C, is handy for this, and a crystal socket may be connected to the grid of the second stage, as shown by J_2 in Fig. 6-6C. Here again, reversion to standard crystal control is easy.

Simplification is possible when the VXO is built directly into a transmitter designed for it. Here, the output of the isolation amplifier will be

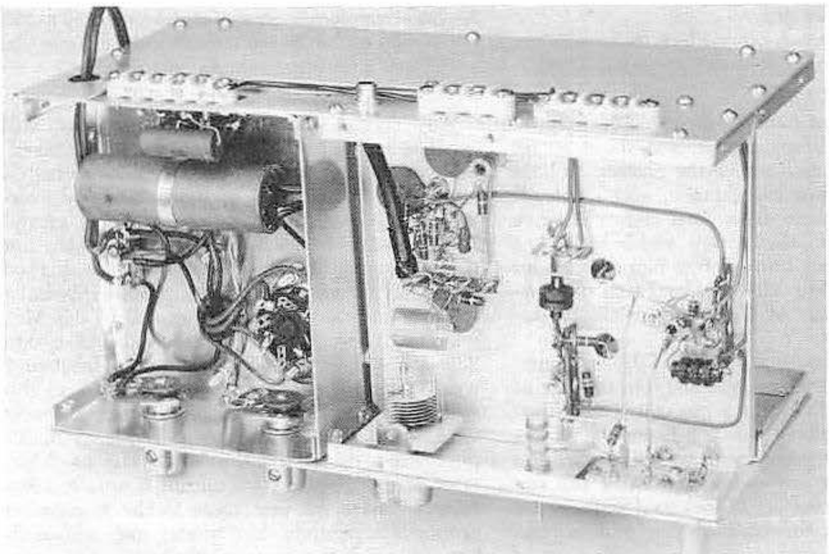


Fig. 6-4—Bottom of the VXO. The oscillator components are at the right, the amplifier and multiplier stages near the center, and power supply at the left.

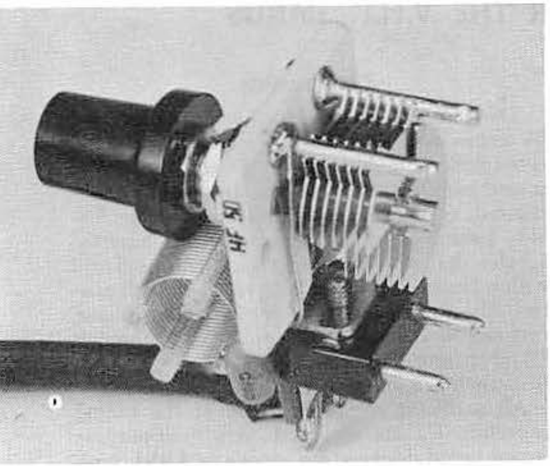


Fig. 6-5—Coupling assembly to be used for plugging into the exciter driven by the VXO. Component are L_5, C_3, and P_1, of Fig. 6-2. The tuned circuit covers 12 to 26 Mc. A larger variable capacitor may be used to make the value of L_5 less critical, if desired.

sufficient to drive a frequency multiplier to 24 Mc., so one stage is saved compared with the system wherein the VXO is used to work into the crystal oscillator stage of a transmitter designed for 6-, 8- or 24-Mc. crystals. But don't skip the buffer amplifier; its functions are vital.

Operation

For maximum stability, particularly in 432-Mc c.w. work, it is well to leave the VXO on continuously during an operating period, and preferably warm it up a few minutes before going on the air. This way there is almost no frequency change, except those deliberately made by moving C_1.

Refinements in the spotting technique can be made to suit the operator's preference, though the circuit is useful as shown. With power applied to the amplifier and multiplier through S_2, the signal is just plainly audible on 432 Mc., when the heaters are on in the rest of the transmitter. It is stronger progressively on each lower band, but the signal from the oscillator alone is inaudible, even on 50 Mc. If you make a practice of zeroing the other fellow's frequency most of the time you may want to install a small relay, actuated by your main transmitter control, in parallel with S_2. Then leave the switch in the open position normally, closing it only for spotting purposes. A spring-return substitute for S_2 may be desirable in this case.

The series coil, L_1, is adjusted by the core stud seen on the front panel, just to the right of the crystal sockets. Moving the core into the coil raises its inductance and increases the swing per crystal. Some practice with various crystals will be needed before you know just what to expect from each one. The coil comes into play only when the crystal is plugged into J_3. Instability increases with inductance, and also with *increasing* capacitance in C_1. Listen to the note critically, and check for mechanical effects when the unit is jarred. Don't push your luck, or expect to swish all over the band with one crystal, even though you'll find one now and then that will make this possible. A movement of 100 kc. at 50 Mc. or 300 kc. at 144 should be possible with most crystals before the instability approaches that of even a good v.f.o.

Generating F.M.

Frequency modulation of the variable oscillator is easily done. A small audio voltage inserted at J_4 will give good-quality f.m. on 144 Mc. and higher, even with the high-stability oscillator arrangement. For 50-Mc. work it may be necessary to use the series-coil circuit to get enough deviation for good audio recovery at the receiving end. All that is needed is a microphone transformer, a flashlight cell and a carbon microphone. Remove the jumper shown across J_1 in Fig. 6-2, of course.

A topnotch f.m. signal can be generated with a very simple audio amplifier having a limiter, and a good microphone. Swinging the frequency with a varactor diode across C_1 offers interesting possibilities. If f.m. is not to be used, omit J_4, RFC_2, and the 4700-ohm resistor, and connect Pin 6 to the 150K dropping resistor.

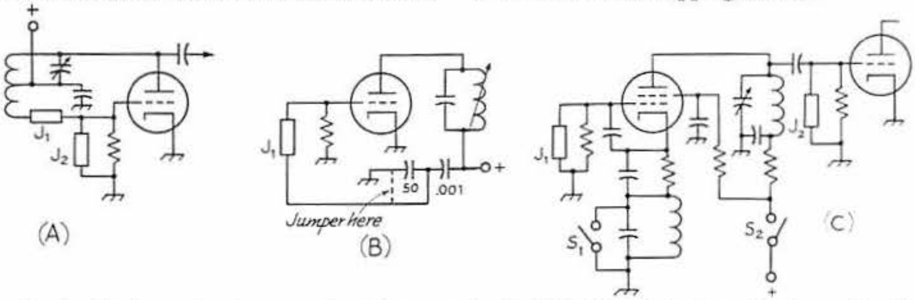


Fig. 6-6—Modifications of various crystal oscillator circuits for VXO drive. J_1 is the original crystal socket. J_2, where required, is an additional socket, for VXO input. A and B are typical triode overtone oscillators. C is a popular pentode oscillator. Two options are shown for C. To convert the oscillator to a multiplier stage, close S_1 and S_2 and feed drive into J_1. The oscillator may be disabled by opening S_2, in which case drive is fed to J_2.

TRANSISTOR V.F.O. FOR THE V.H.F. BANDS

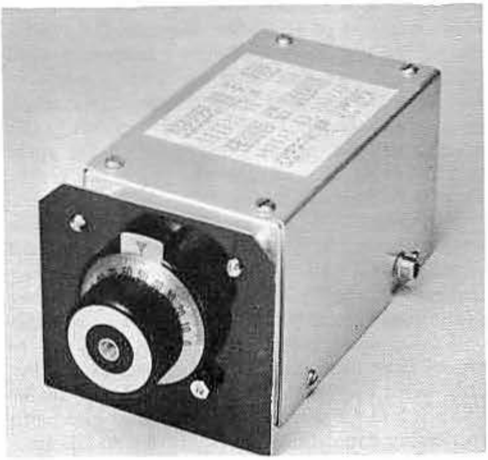


Fig. 6-7—Transistor v.f.o. for v.h.f. use. Tuning range is about 8 to 8.42 Mc., as described.

One source of hum in variable-frequency oscillators using vacuum tubes is the a.c. voltage used on the heater of the oscillator tube. Another is inadequate filtering of the oscillator plate supply. By using a transistor oscillator, these two factors can be eliminated, as the transistor runs entirely on d.c. Some other instability problems are more readily solved with transistors as well. If a battery is used for a power source, changes in frequency due to voltage fluctuation are nil. Drift cycles resulting from alternate heating and cooling of the oscillator can be virtually eliminated in a good transistor v.f.o. design.

This is not to say that the transistor is a cure-all for v.f.o. troubles, but with proper attention to basic principles of oscillator design, it is possible to make a reasonably good v.f.o. with transistors, at moderate cost. By using an etched circuit board to mount and wire the oscillator components, most of the sources of mechanical instability are taken care of. The oscillator should operate at the lowest practical power input, to prevent heat cycling, and the oscillator should be isolated from the rest of the transmitter with at least one buffer stage.

These points were considered in the design of the v.f.o. unit shown in Fig. 6-7. Its transistor oscillator tunes 8.0 to about 8.42 Mc., or enough to cover the entire 144- and 220-Mc. bands, and the 50-Mc. band from the low end to about 50.5 Mc. More coverage can be included with a larger capacitor for C_8. The oscillator input is less than 40 milliwatts, so there is little heating and warm-up drift. What there is can be corrected by adjusting the temperature-compensating padder across the tuned circuit. A buffer-amplifier provides some isolation, and builds up the output up to permit operation with most v.h.f. transmitters having crystal-controlled oscillators that multiply into the 24-to-25-Mc. range.

Construction

The v.f.o. is built in a box 3 inches square and 5¼ inches long. This was made to fit the job, and is not difficult to duplicate, but standard cases of something like this shape and size could be substituted, as there is nothing sacred about the layout. Tuning is by means of a small imported vernier dial, the knob of which was replaced with a larger one, in the interest of smooth control. The front panel is 3⅛ by 3¼ inches in size, fastened to the case with bolts passing through 11/16-inch metal sleeves. These were filed down from their original ¾-inch length.

All oscillator components except the tuning capacitor, C_8, are mounted on a multipurpose circuit board designed for service in ARRL projects where tunable oscillators are required. A layout is given in Fig. 6-11, if you want to make your own. If not, duplicates can be purchased.* The drawing and photographs should make clear where the various parts are mounted.

The amplifier stage is at the back of the chassis. The transistor is mounted on three small feed-through bushings (Johnson Rib-Loc). These stay in place when pressed into a 0.136-inch hole. Two other Rib-Loc terminals provide for the connection of the power source. These and the output jack are on the back wall of the chassis. On a side wall, near the back, is another phono jack for external connection of a spot-

*ARRL Type CO-1, Harris Co., 56 East Main St., Torrington, Conn. 06790

Fig. 6-8—Interior of the transistor v.f.o. The oscillator portion is built on a circuit board. The amplifier stage is close to the rear wall.

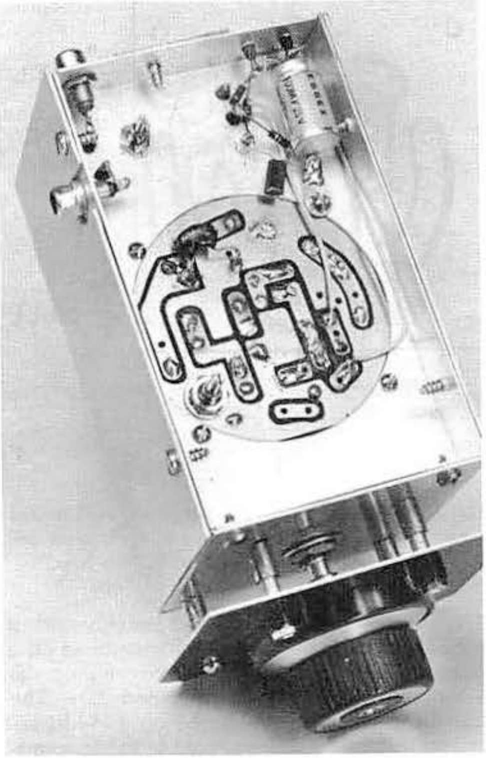


Fig. 6-10—Bottom of the transistor v.f.o. assembly. The oscillator circuit board assembly mounts over a 2¾-inch round hole. Power feed-through bushings are in upper right corner.

ting switch or relay, to turn the amplifier stage on or off.

Small front and back plates have folded-over edges. The cover is a U-shaped piece, fastening to these, and to the sides of the chassis. The tuning capacitor mounts on the front panel with a nut on the rotor shaft. The type that grounds the rotor in this way is preferred to those having small mounting studs, as the electrical and mechanical grounding is much better. If more frequency coverage is desired, the tuning capacitor should be the next larger size than the 15-pf. type shown. Plates can then be removed if the coverage is too great.

Almost any h.f. or v.h.f. transistors may be used. The circuit board is designed so that either p-n-p or n-p-n can be accommodated, and there is no advantage either way. Motorola 2N4125 plastic-case p-n-p types are shown. RCA 2N1177s were also tried. A zener diode, CR_1, shown in the circuit diagram, does not appear in the photographs. This is used to regulate the supply voltage if other than a battery source is used. It would be important if the v.f.o. were to be used for mobile work, on the car battery. In operation from a small separate battery the Zener can be omitted, and the positive voltage applied directly to the junction of R_2, R_3 and R_4.

Adjustment and Use

Make certain that all circuits are wired correctly, then apply voltage. It would be well to check the current drawn, which will be about 10 ma. at 12 volts and 7 to 8 ma. with 9 volts. Of this, about two thirds is drawn by the amplifier. If the oscillator current is measured separately on a low-range meter, a slight flicker in current will be seen when the oscillator coil or tuning capacitor is touched, if the circuit is oscillating.

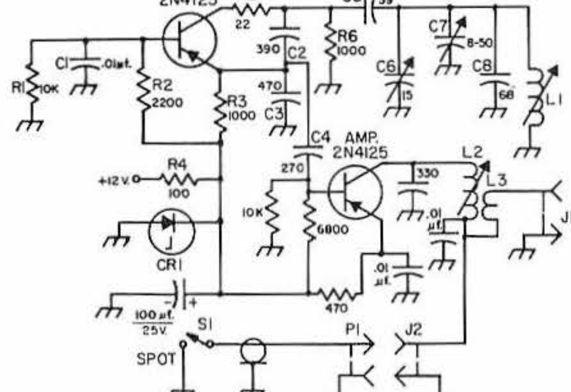


Fig. 6-9—Schematic diagram and parts information for the 8-Mc. v.f.o. unit. Parts not described below are numbered for identification in text. Where not otherwise indicated, capacitor values are in picofarads (pf. or μμf.). Those across tuned circuits are dipped-mica.

C_6—15-pf. variable (Millen 22015). For more tuning range use larger value.

C_7—8 to 50-pf. neg. temp. coef. ceramic trimmer (Centralab N-650).

J_1, J_2—phono jack.

L_1—3-uh. slug-tuned coil: 13 turns No. 22 on ⅜-inch iron-slug form (Miller 42A336CBI, with 2 turns removed).

L_2—1.3-uh. slug-tuned coil: 14 turns No. 26 on ¼-inch iron-slug form (Miller 4502).

L_3—2 turns No. 26, wound over low end of L_2.

CR_1—9-volt zener diode.

P_1—Phono plug. Short out if no spotting switch is used.

S_1—Remote spotting switch, any type.

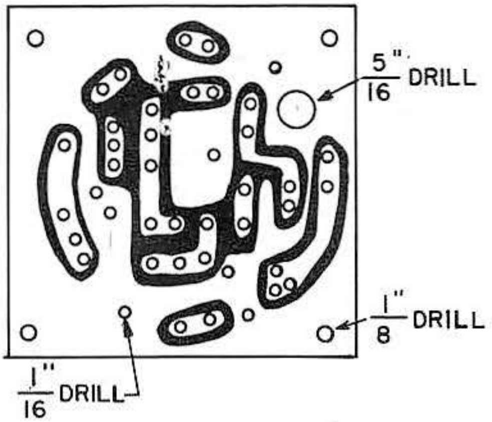


Fig. 6-11—General purpose circuit board used for the transistor v.f.o. is 2⅞ inches square. Dark areas are etched. Board mounts over a 2¾-inch hole.

The frequency can be checked roughly with a wavemeter, or the signal can be monitored on a receiver. A well-calibrated receiver tuning the 8-Mc. range is handy, but not necessary. The signal can be heard in your 50- or 144-Mc. receiver, if the v.f.o. is connected to the transmitter with which it is to be used. Methods of connection to various types of crystal oscillator circuits are discussed in connection with the VXO. See Fig. 6-6 and associated text. A shielded 8-Mc. coupling unit is shown in Fig. 6-12.

To calibrate the oscillator, turn capacitor C_6 to all-in, and adjust the core slug in L_1 so that the frequency is approximately 8000 kc. This will be heard at the low end of the 144-Mc. band, if you are listening there. It may be necessary to run a link from J_1 to the 144-Mc. receiver input, if you are not driving the transmitter at this time.

A drift check should next be made. Be sure that the receiver is not drifting, then turn the v.f.o. on and note the drift for the next two minutes or so. If it is appreciable, it can be reduced by proper setting of the compensating trimmer, C_7. Move the trimmer a few degrees either way, and reset the core slug so that the signal is again heard at 8 Mc. Check the drift cycle again. If there is less drift this time, from a cold start, you moved the trimmer in the right direction. If there is more, move the trimmer about as far in the opposite direction, reset the slug, and try again. A combination will be found eventually where the temperature-compensating qualities of the trimmer will almost exactly nullify the effect of the slight transistor warm-up drift. The temperature compensation is really just a refinement; the stability of the oscillator is quite good without compensation, if you don't mind letting it run for 10 minutes or so before using it. It's not bad, even from a cold start.

A dummy load for the amplifier can be made with a 2-volt 60-ma. pilot lamp, wired to a phono plug and inserted in J_1. It will just glow when the amplifier output circuit is peaked, indicating an output of about 50 milliwatts. This is enough to drive any of the common crystal oscillator circuits, if the coupling system of Fig. 6-2 is used. For an 8-Mc. circuit, L_5 in Fig. 6-2 should have 24 turns.

The simple bipolar transistor buffer stage does not afford a high degree of isolation. There will be an appreciable frequency change as the coupling circuit is adjusted, but the stability in actual use is fairly good. This is not a device for working close to the band edge, however. Play safe, and be sure that you know where you are.

This v.f.o. was tested on 50 and 144 Mc. with the "Two-Band Station" transmitters described in the following pages. These were modified by installing a switch that shorts out RFC_1 (Fig. 6-14) when the v.f.o is used. The v.f.o. works well with the 220-Mc. transmitter described later in this chapter, the oscillator circuit of this unit having been modified in a similar manner. It drove a Clegg 22-er nicely when the coupling circuit, Fig. 6-12, was plugged into the crystal socket. In each instance the note quality was acceptable, and there is substantially no frequency modulation. Drift is no more than some crystal-controlled transmitters show at comparable frequencies. Stability does not approach that of the VXO just described, but it is better than many v.f.o. units currently heard on the v.h.f. bands, both commercial and home-built.

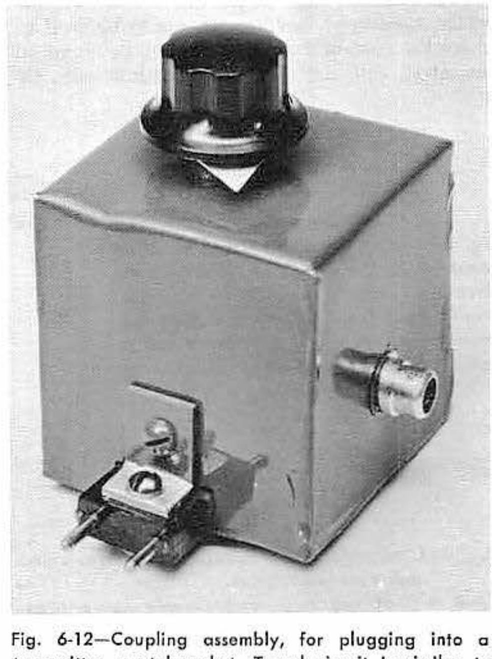


Fig. 6-12—Coupling assembly, for plugging into a transmitter crystal socket. Tuned circuit is similar to that shown in Fig. 6-2, except L_5 has 24 turns. Interior appearance is similar to Fig. 6-5.

V.H.F. TRANSMITTERS

The amateur about to begin construction of a v.h.f. station is frequently bewildered by the choice of tubes, circuits and equipment layouts available to him. The newcomer often feels that he wants a single unit, complete with r.f., audio and power supply, all in one table-top package. This is convenient in some respects, but it is by no means the most versatile approach. There are other ways to do the job, particularly if the amateur is the "building" type.

Most of us don't build transmitters for the sole purpose of being able to talk to people. Rather, we like to try different circuits and combinations. We look forward to a gradual building-up and refinement of our station, as experience and finances permit. With some planning, today's low-powered transmitter can become tomorrow's exciter, and if it is a separate unit it can be used for occasional portable work.

There is much to be said for subassembly design, and most of the equipment in this book is worked out along these lines. Even the one complete medium-powered r.f. section for 144 Mc., Fig. 6-18, is laid out so that the exciter and amplifier can be built and used separately, if desired, though the two go together in one standard 17-inch-wide package.

Such r.f. subassemblies are usually much easier to build, adjust and trouble-shoot than are complex all-on-one-chassis designs, and for this reason they are especially desirable for the newcomer. Each represents a relatively small investment in labor and parts; when the time comes for something better, or you want to add an accessory or make a modification, you can go about it without the fear of spoiling something that has all your financial and technical resources tied up in it.

EFFICIENT LOW-POWER TRANSMITTERS FOR 50 AND 144 MC.

The r.f. units of Fig. 6-13 were designed to be easy to build and adjust. They are stable in operation, and are relatively free of unwanted frequencies that could cause TVI. They scrimp on no essentials, and they have features that may save you money in the long run. Both employ crystals that are inexpensive and reliable. By shopping for surplus crystals you can afford enough of them to operate close to any desired frequency. Shifting from one spot to another is done with a minimum of retuning, thanks to a reserve of driving power all along the line. The oscillator circuit is readily adapted to v.f.o. control, should you want to go to it eventually. See Fig. 6-6C for method of connection. The transmitters can be keyed for c.w. communication, and the signal will sound like a c.w. signal should, without the annoying yoops so often heard in v.h.f. c.w. work. With this equipment your signal will require no apologies, and you will have a fine base on which to expand to higher power later on.

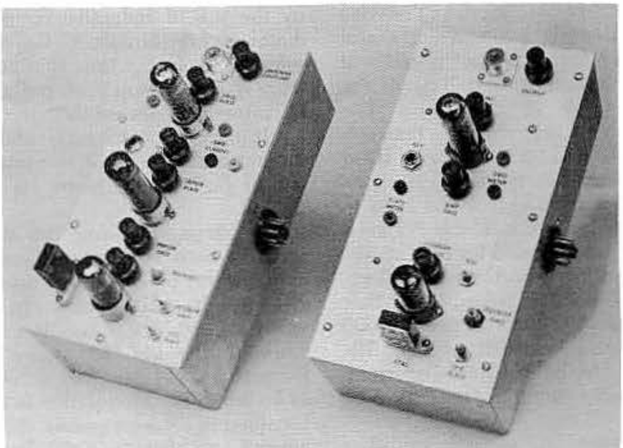


Fig. 6-13—The r.f. units for 144 Mc. (left) and 50 Mc. are as much alike mechanically and electrically as possible. Shown here side by side, they have their crystal oscillators at the low end of the picture. Provision is made for measuring grid and plate current by plugging a meter into insulated tip jacks. The transmitters plug into the side of the modulator and power-supply chassis described in Chapter 7, or they may be connected to it through 4-wire cables of suitable length.

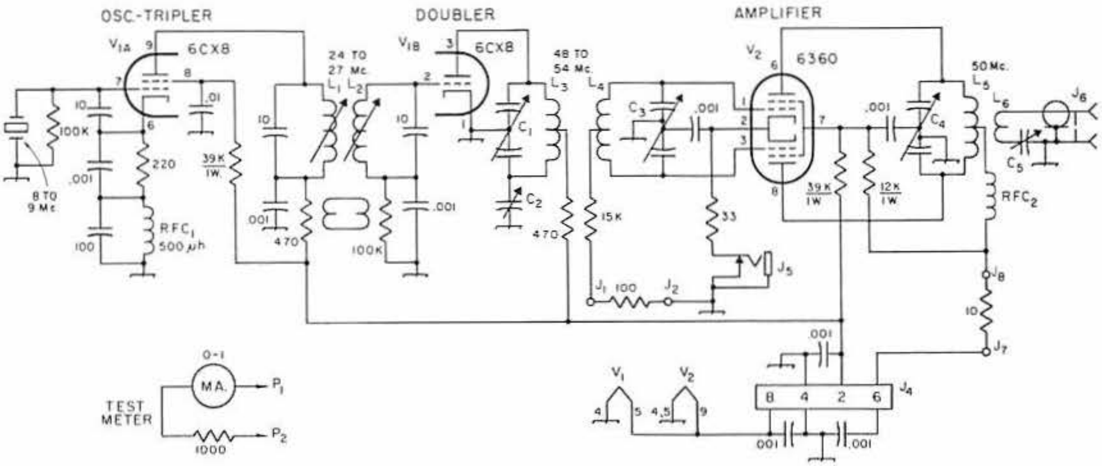


Fig. 6-14—Schematic diagram and parts information for the 50-Mc. transmitter. Fixed 10- and 100-pf. capacitors are mica; 0.01- and 0.001-μf. are ceramic disk. Decimal values are in μf. Resistors are ½ watt unless specified; values in ohms. The oscillator and doubler stages of the 144-Mc. transmitter are similar.

C_1, C_3, C_4—8-pf. miniature butterfly variable (Johnson 160-208 or 9MB11).
C_2—8.7-pf. miniature variable (Johnson 160-104 or 9M11).
C_5—50-pf. miniature variable (Hammarlund MAPC-50-B).
J_1, J_2, J_7, J_8—Insulated tip jack.
J_4—8-pin male chassis fitting (Amphenol 86-CP8).
J_5—Closed-circuit phone jack.
J_6—Coaxial output receptacle, SO-239.
L_1, L_2—3-μh. (approx.) iron-slug coil (Miller 4404). Link L_1 and L_2 with 1-turn loops of insulated hookup wire.
L_3—10 turns No. 20 tinned, ¾-inch diam., 16 t.p.i., c.t. (B & W No. 3011).
L_4—8 turns like L_3. L_3 and L_4 are side by side, 1 inch apart center to center.
L_5—11 turns like L_3.
L_6—1¾ turns insulated hookup wire around center of L_5.
P_1, P_2—Insulated tip plug.
RFC_1—500-μh. r.f. choke.
RFC_2—Single-layer v.h.f. r.f. choke, 4 to 7 μh. 42 turns No. 26 enamel, close-wound on 9/32-inch dowel.

The Circuits

It will be seen that the transmitters are very similar. They are so much alike, in fact, that we did not repeat duplicate parts of the circuit in the diagram of the 144-Mc. model. The two transmitters will be described concurrently, and unless the text states otherwise, what is said will apply to both units. The crystal oscillator is the pentode section of a 6CX8 dual tube. The 6CX8 triode is a doubler stage. Crystals between 8000 and 8222 kc. are used for the 2-meter band (8056 to 8166 kc. for the Novice-Technician portion between 145 and 147 Mc.) and 8334 to 9000 kc. for the 50-Mc. band. Those between 8334 and 8350 kc. should be used for c.w. operation only, as they multiply into the first 100 kc. of the 50-Mc. band, which is set aside for that mode only. Appropriate crystals between 6000 and 6750 kc. may also be used, as may 12- and 24-Mc. crystals. The latter two are overtone types and will not be as stable as those for 8 or 6 Mc.

The oscillator requires no adjustment other than moving the core in the plate coil, L_1. This is tuned between 24 and 27 Mc., depending on the crystal frequency. The 6CX8 pentode triples the frequency for 8-Mc. crystals and quadruples it for 6-Mc. ones. Loosely coupled tuned circuits, L_1 and L_2, in the oscillator plate and doubler grid emphasize the desired harmonic and help to reject unwanted frequencies that are developed in the oscillator.

Attenuation of unwanted frequencies is aided by the use of inductive coupling between the doubler plate circuit, C_1L_3, and the following grid circuit, C_3L_4. Note that here a single-ended stage is coupled to a push-pull one. The capacitor C_2 is used to balance this circuit for coupling to L_4. It places a capacitance similar to the plate-to-ground capacitance of the tube at the opposite end of L_3 from the plate. Its adjustment is not critical.

The stage following the doubler looks the same in both schematic diagrams, but it is an amplifier for 50 Mc. and a frequency tripler for 144 Mc. Its plate-screen circuit is modulated when the 50-Mc. transmitter is used for voice work. In the 144-Mc. model this stage triples from 48 to 144 Mc. and drives a similar stage as an amplifier. Modulation is applied to the latter stage in 2-meter phone operation. Both tripler and amplifier are 6360 dual tetrodes. Power input to the amplifier runs about 15 watts on phone, but may be increased to 20 watts or more on c.w. The key is inserted in the amplifier cathode jack, J_5. Tuned antenna coupling conveys the transmitter output to a coaxial line to the antenna change-over relay, which is part of the modulator unit, described in Chapter 7. Tip jacks are provided for measuring tripler and amplifier grid current, and amplifier plate current.

Construction*

The transmitters are built on aluminum plates that are screwed onto aluminum chassis 5 by 10 by 3 inches in size. Leads are brought to a plug mounted in the right side of the transmitter chassis, for plugging into the power socket on the left side of the modulator chassis. The transmitter and modulator units may also be operated apart by making up a suitable cable for connecting the two.

It will be seen that various components come close to the edges of the plate on which the transmitter is built. To avoid possible damage when the units are mounted on or removed from the chassis, it is desirable to cut notches in the folded-over edges of the chassis to give plenty of clearance around these parts. This is particularly true of the output tuning capacitors, which are vulnerable in this respect.

The transmitters are shown together in the top view, with the 144-Mc. model at the left. The crystal oscillator and doubler are at the bottom of the picture for each, with the antenna jack and tuning capacitor at the opposite end.

* Templates for use in drilling the top surfaces of these transmitters are available from the ARRL Technical Department. Please send 25 cents and a stamped self-addressed envelope and give the publication, edition, and figure number for which the template is requested.

In the bottom views the oscillator tube and components are at the left end of the assembly. The oscillator plate and doubler grid coils appear in the upper left corner. These are ⅞ inch apart, center to center, in the 144-Mc. transmitter, and 1 inch in that for 50-Mc. Smaller-diameter coils were used in the former, though similar ones could have been used in each. The coupling link between these coils is made of a single piece of insulated wire looped around one coil, the leads crossed over and then looped around the other coil and then the ends soldered together. The figure-8 loop is visible in both pictures.

The spacing between the inductively coupled coils elsewhere in the transmitters is given below the schematic diagrams. It will be seen that the doubler plate and tripler-amplifier grid coils, L_3 and L_4, are side by side, whereas the tripler plate and amplifier grid coils in the 144-Mc. rig, L_5 and L_6, are mounted on the same axis.

Wiring of the transmitters is extremely simple. Use tie-point strips liberally for terminating power leads and mounting resistors and bypass capacitors. Shielded wire can be used for power leads, though it was not done in these units. Run power wiring flat against the plate. The ready-wound coil stock can be tapped most readily if the turn where the tap is to be made is pressed down toward the center of the coil with a small screwdriver. Connection to the tap is then made on the inside of the coil, using a small soldering iron. The coils are supported by their own leads, soldered to the tuning capacitors as directly as possible. Cutting of Miniductor stock is described in Chapter 13.

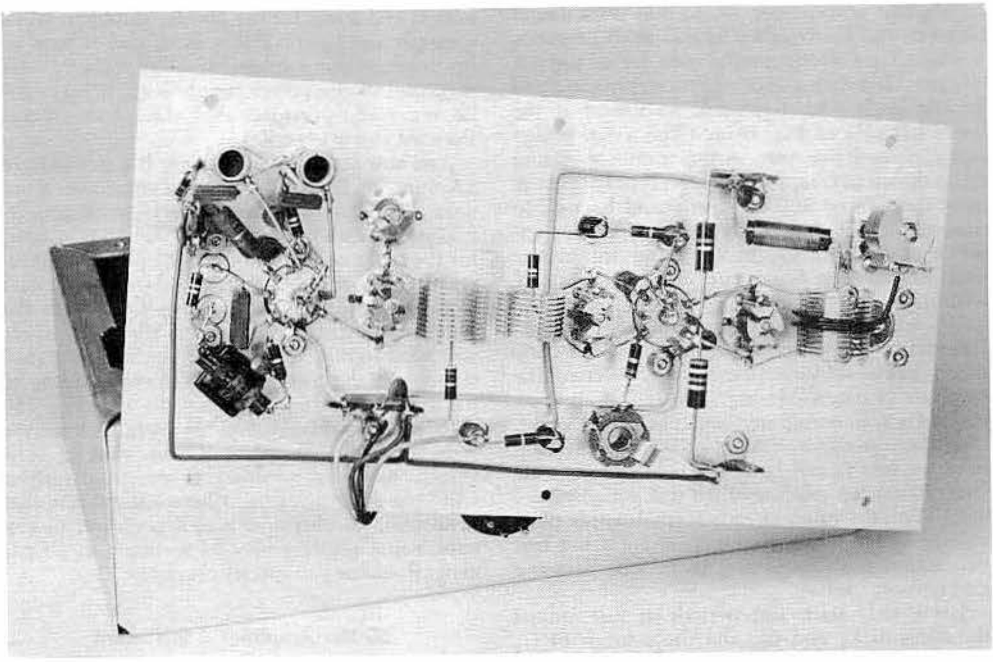


Fig. 6-15—Bottom view of the 50-Mc. transmitter. The crystal oscillator is at the left. The amplifier plate circuit and antenna loading control are at the right.

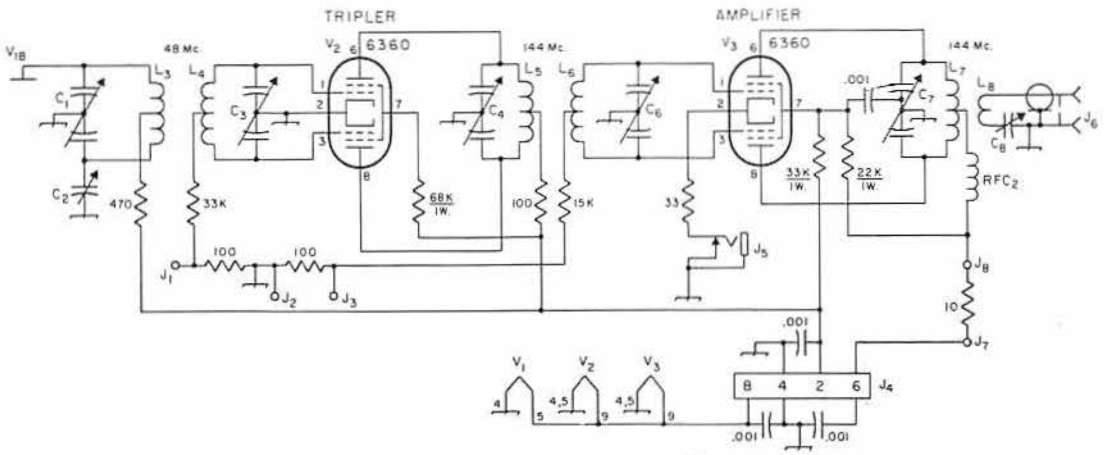


Fig. 6-16—Schematic diagram and parts information for the 144-Mc. transmitter. Only the tripler and amplifier portions are shown, as the oscillator and doubler stages are similar to the 50-Mc. unit. Components not listed below are identical to those of Fig. 4.

- C_3, C_6—5-pf. miniature butterfly variable (Johnson 160-205 or 5MB11).
- C_4, C_7—8-pf. miniature butterfly variable (Johnson 160-208 or 9MB11).
- C_8—30-pf. miniature variable (Johnson 160-130 or 30MB).
- J_8—Insulated tip jack.
- L_1, L_2 (in Fig. 6-14)—4-μh. (approx.) ¼-inch iron-slug coil (Miller 4504). Link with 1 turn insulated hookup wire; see photo and text.
- L_3—13¼ turns No. 24 tinned, ½-inch diam., 32 t.p.i., c.t. (B & W Miniductor 3004).
- L_4—Same as L_3, but 10¼ turns. Mount L_3 and L_4 ⅞ inch apart, center to center.
- L_5—3¾ turns No. 20 tinned, ½-inch diam., 16 t.p.i. (B & W No. 3003).
- L_6—2¼ turns like L_5.
- L_7—6 turns No. 18 tinned, ⅜-inch diam., 9/16 inch long, c.t.
- L_8—1 turn insulated hookup wire around center of L_7.

Adjustment and Operation

The transmitters can be tested with any power supply that will deliver 200 to 300 volts d.c. at 100 ma., and 6.3 volts a.c. or d.c. at 2½ amperes. A single 1-ma. meter can be used for all tests, if it is provided with a 1000-ohm series resistor and flexible leads with terminals, as shown at the lower left side of Fig. 6-14. (The s.w.r. bridge meter is used this way in the complete station setup shown in Chapter 7.) If the supply delivers more than about 200 volts, it would be well to connect a 5000-ohm 10-watt resistor in the supply lead temporarily, to keep the transmitter from drawing excessive current at the start of testing.

We will test the oscillator and doubler first. Disconnect the screen resistors from both 6360 stages in the 144-Mc. transmitter, or from the amplifier in the 50-Mc. rig. This will keep these stages from drawing anything but grid current. Plug the test meter, with the 1000-ohm resistor in series, into J_1 and J_2. It will read as if its scale were 10 ma. (A reading of 0.4 will actually be 4 ma.) With the tubes already heated, apply plate voltage briefly, through Pin 2 of J_4. If the first two stages are functioning there will be some grid current reading.

Using only short test periods at first, adjust the cores in L_1 and L_2, and the settings of C_1 and C_3, for maximum grid current. Now adjust C_2 for maximum grid current. It will be seen that C_1 and C_2 interlock. Move first one and then the other until the combination is found that gives the highest grid current. This should be at least 1 ma. in the tripler of the 144-Mc. transmitter, and 2 ma. for the 50-Mc. amplifier, when a supply voltage of 250 is used, and it may be up to twice these values. If a dropping resistor was used in the power supply lead, it may now be removed, provided that the plate voltage does not rise to over 300.

And how do you read voltage? It's nice to have a voltmeter, but you can make your own. Connect a 1-megohm resistor in series with your 1-ma. meter, with or without its 1000-ohm resistor, for the latter will make only a 0.1 per cent difference. Connect the negative side of the meter to the chassis, and the positive side (with the 1-megohm resistor in series) to the point where you want to measure voltage. You can now read voltage on the meter scale. A meter reading of 0.3 ma. will mean 300 volts, 0.28 would be 280 volts, etc. It is desirable to have a fairly accurate resistor for this purpose, if you want to read voltage to useful accuracy. A precision resistor will be a good investment here but get one that is accurate to plus or minus 5 per cent, in any case. Some resistors may be as much as 20 per cent off, unless you specify otherwise.

50-Mc. Amplifier Adjustment

The 50-Mc. amplifier may now be adjusted, but first we need some kind of dummy (nonradiating) load. The best load is a bank of resistors that will total about 50 ohms and be able to dissipate at least 8 watts. Suitable loads are described in chapter 11. To use such a load properly requires some form of power output indicator, inserted in the line to the load. The s.w.r. bridge, Fig. 11-14, serves this purpose.

Lamps of various kinds can be used, but they are inferior loads. They do have an advantage, however: they give a rough visible indication of power output. Probably the best lamp load is made of 4 or 5 blue-bead pilot lamps (No. 44 or 46) connected in parallel. A 25- or 40-watt lamp may also be used, but it will be far from a 50-ohm load, and very misleading as to tuning of the final plate and loading circuits. If such a lamp is used, short out the loading capacitor, C_5, temporarily.

With the two previous stages having been tuned for maximum amplifier grid current, reconnect the screen resistors. Modulation is not needed at this stage, so Pins 2 and 6 of J_4 may be connected together initially. Plug the meter and 1000-ohm series resistor into J_7 and J_8, to measure amplifier plate current and apply voltage. The meter will now read as if it had a 100-ma. scale. Adjust C_4 quickly for minimum plate current, which should be about 50 to 80 ma., if a load is connected to J_6. If the load is a lamp or bank of lamps, adjust C_4 for maximum brilliance. With the pilot-lamp load C_5 may now be adjusted for maximum brilliance. Retune C_4 and C_5 several times for greatest output. If a regular home light bulb is used for the load, short C_5 temporarily and adjust C_4 for maximum brilliance. Maximum output will occur at approximately minimum plate current, but there may not be exact coincidence, so C_4 should be adjusted for the lowest plate current that gives maximum output.

The 6360 is so designed that there is no need for neutralization if the transmitter is properly designed and built, but a stability check should now be made. Plug the meter back into the grid-current jacks, turn on the transmitter, and *briefly* remove the crystal from its socket. There should be no grid current with the crystal removed. The input to the amplifier will run excessively high under this test, so do it for a short check only.

Another test for stability is to observe the grid current and plate current simultaneously, while watching the output. A perfectly stable transmitter will show maximum grid current, minimum plate current and maximum output at a single setting of the plate capacitor. Some divergence from the ideal is permissible, if other indications given above are achieved.

144-Mc. Adjustment

Thus far we've been talking about the 50-Mc. transmitter. Adjustment procedure is similar for the 144-Mc. model, Fig. 6-17, except that there is a bit more to it. Proceed as above to the point where you have gotten grid current in the tripler

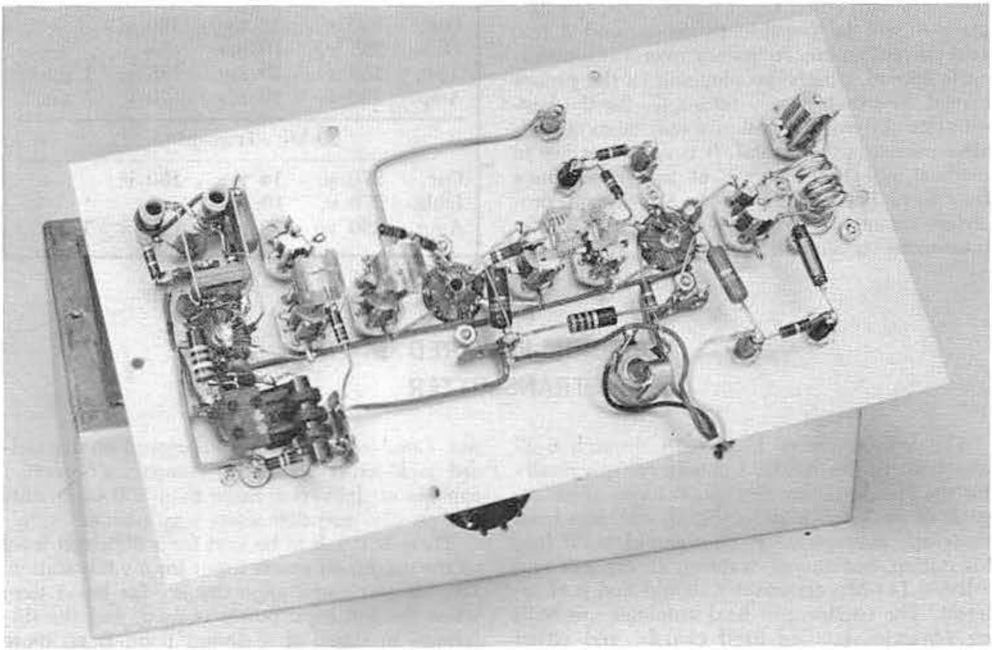


Fig. 6-17—The 144-Mc. transmitter is similar to that for 50-Mc., but it requires one more stage. Oscillator and doubler circuits are at the left-end. Side-by-side coils in the doubler plate and tripler grid circuits come next. The tripler plate and amplifier grid coils, right center, are mounted on the same axis. The amplifier plate and loading circuits are at the far right.

stage. Now connect the screen resistor of the tripler and put the meter in tip jacks J_2 and J_3, to measure amplifier grid current. Apply voltage through pin 2, and tune C_4 for maximum amplifier grid current. This should be at least 2 ma., but it may be as much as 5.

Now plug the meter into J_7 and J_8 and apply plate voltage through Pins 2 and 6. Adjustment from here on is similar to the 50-Mc. amplifier. Because of the drain imposed by the extra 6360 stage, the plate-supply voltage will be a bit lower with the 2-meter transmitter, a fact to keep in mind when figuring the input you will have to modulate.

Once the transmitters are made to work on a given frequency you may want to tune them so that shifting frequency can be done with a minimum of retuning. There is a surplus of grid drive with the tube lineups shown, so "stagger-tuning" is entirely practical. For instance, the 2-meter transmitter can be adjusted so that any frequency between 144 and 146 Mc. can be used merely by inserting the proper crystal and retuning the final plate circuit. Plug the meter into the amplifier grid jacks, J_2 and J_3. With a crystal near 8000 kc. in place, tune for maximum grid current. It will be more than you need. Now put in a crystal for some point near 146 Mc. The grid current will probably be 1 ma. or less, and the output somewhat low, even when the final plate circuit is retuned. Adjust one of the core studs (either L_1 or L_2) *upward slightly*, and see if the grid current rises. Retune C_1 or C_2 slightly to further increase the grid current. Do the same with either C_4 or C_6. By judicious juggling it will be possible to get around 3 ma. grid current on any frequency over a two-megacycle spread, simply by plugging in the proper crystal. You then merely retune C_7 for the lowest plate current that will give maximum output, after changing the crystal. It is not necessary to readjust either C_2 or C_8 at any time, once they have been properly set. Adjustment procedure for spreading the coverage of the 50-Mc. transmitter is similar, but simpler because of the lesser number of stages.

Ideal amplifier grid current in both transmitters is around 3 ma., though either will work well with down to about 1½ ma., or up to 4. More than 4 ma. is likely to reduce the output, and either insufficient or excessive drive will affect the modulation adversely. The amount of grid drive for c.w. operation is much less critical, it being merely necessary to have enough to insure efficient operation. Even 1 ma. will do. Keying the transmitters for c.w. work is done by plugging a key into the cathode jack, J_5.

Plate current may be measured in any stage, to be sure that it is running at safe input. Connect a 10-ohm resistor in series with the lower end of the 470-ohm isolating resistor in the plate circuit of the stage to be checked. Now, connect the 1-ma. meter (with its 1000-ohm resistor in series) across the 10-ohm resistor. This will make the meter read as if it had a 100-ma. scale, just as when you plug into J_7 and J_8. Additional tip jacks shunted with 10-ohm resistors may, in fact, be permanently a part of the transmitters, though there will be little need to use them after the stages are once checked and found to be operating satisfactorily. The accompanying table shows typical voltages and currents measured in the original units.

Stage	*Plate Voltage*	*Plate Current*	*Screen Voltage*	*Grid Current*
		144-Mc. Transmitter		
Osc.	255 v.	12 ma.	140 v.	—
Dblr.	255 v.	10 ma.	—	—
Tplr.	255 v.	50 ma.	125 v.	1 ma.
Amp.	230 v.	70 ma.	170 v.	3 ma.
		50-Mc. Transmitter		
Osc.	270 v.	14 ma.	150 v.	—
Dblr.	270 v.	10 ma.	—	—
Amp.	250 v.	70 ma.	170 v.	3 ma.

A MEDIUM-POWERED 144-MC. TRANSMITTER

The transmitter of Figs. 6-18 through 6-25 was built in the ARRL Laboratory, specifically for the Headquarters Station, W1AW. Some aspects of its design may be slightly different from those the average v.h.f. man would build into his station, but anyone wanting an efficient and reliable 144-Mc. transmitter should find it of interest. The exciter and final amplifier are built on separate standard-sized chassis, and either may be used with other suitable equipment. The exciter makes a fine low-powered r.f. unit by itself, and the amplifier will work well with any exciter capable of delivering over 5 watts output. Combined the two units mount on a standard rack panel, making a compact r.f. section capable of delivering more than 300 watts output on c.w. and 200 watts a.m. phone.

There is much to be said for a 500-watt level as the maximum power input for a v.h.f. station. Overall cost and complexity are far lower than when the full legal power is used, and the difference in signal at a distant point is no more than barely noticeable. The economic advantage gained in staying below 500 watts input is particularly marked when a.m. phone is the most-used mode of operation.

Exciter Design

The exciter is designed to permit shifting frequency over most of the band without extensive retuning. If the stages are peaked near the middle of a two-megacycle range normally used, very little readjustment of the exciter will be required. Even in moving from one end of the band to the other, only repeaking of the tripler and amplifier tuning capacitors will be needed. The tubes run at a conservative level, to assure trouble-free operation in the continuous nightly service encountered in bulletin and code-practice transmissions from the Headquarters station. Double-tuned inductively-coupled circuits throughout give the desired band-pass response, but with selectivity sufficient to attentuate unwanted multiples of the oscillator frequency that might go through to the amplifier and be radiated if simpler circuits were used.

There are three dual tubes in the exciter. The oscillator and multiplier stages are 6AR11 dual pentodes, and the output stage is a 6360 dual tetrode. The oscillator, V_{1A}, is set up for 8-Mc. crystals, though fundamental crystals at 6 or 12 Mc. will also work. The switch S_1 selects one of five crystals. One position may be used for v.f.o. input if the switch S_2 (not in the original unit) is incorporated to short out the cathode r.f. choke, RFC_1.

The oscillator plate circuit multiplies to 24 Mc. Its plate coil, L_1, is inductively coupled to the following grid circuit, L_2, and the coupling is increased by the link, L_3. The second pentode of V_1 doubles to 48 Mc. Its plate circuit and the grid circuit of the push-pull tripler, V_2, are also link-coupled. The tripler plate circuit and the grid circuit of the 144-Mc. amplifier, V_3, are inductively coupled. Amplifier output is taken off through a series-tuned link to a coaxial fitting on the rear wall of the exciter chassis.

Grid current in the tripler or amplifier stages can be measured by connecting a low-range milliammeter between the exposed terminal of C_3 or C_5 and the chassis. This is helpful in initial adjustment, and for trouble shooting, if needed.

The Final Amplifier

It is desirable that a transmitter be capable of running at moderate power, particularly in v.h.f. work, where about 50 watts output is adequate for most communication. The external-anode type of v.h.f. tetrode fits this need admirably, as it will run efficiently at inputs of 100 watts or less, yet it can be pushed up to 500 watts with complete safety.

The 4CX250R shown can be replaced with any tube of this family. Many builders will want to use the 4X150A, which can be found on the surplus market at attractive prices. It will work equally well, except for slightly reduced ratings. The socket specified takes the 4X150A, 4X250B, and 4CX250B and R. With a suitable socket, and possibly modification of the grid circuit to take care of differing input capacitances, any of the many tubes of this general type can be used.

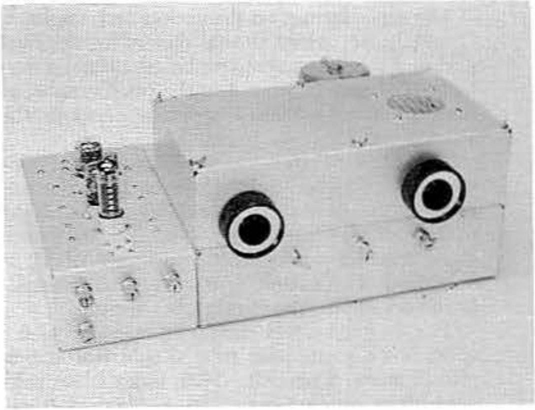


Fig. 6-18—The 500-watt 144-Mc. transmitter is built in separate assemblies using standard chassis sizes, yet it can be mounted on a standard 19-inch rack panel. The exciter, left, may also be used as a low-powered transmitter, capable of up to 10 watts output. Amplifier, right, has built-in bias supply. Its simple strip-line plate circuit is enclosed in a removable cover.

The amplifier tank circuits are made from flashing copper, a readily-available material that can be cut and bent without special tools. Details of the grid inductance, L_{12}, and the plate line, L_{13}, are given in Fig. 6-25. Many plate circuits were made and tested to derive the shape of L_{13}. These included copper pipes from ⅜ to 1 inch in diameter, and copper strips of several widths and configurations. The plate circuit was operated as a pi-network, as well as in the inductively coupled form shown here. When optimum L/C ratio was achieved (tuning to the desired frequency range with the lowest usable capacitance) there was no essential difference in results, so the convenient and safe grounded tank shown was adopted.

Power Circuits

Only two external power supplies are needed for the r.f. portion of the transmitter: one delivering 250 volts d.c. at about 150 ma., and a high-voltage supply giving anything up to 2000 volts at 300 ma. Control of the a.c. voltage input to the final-stage plate transformer by means of a Variac or Powerstat is an excellent way of adjusting the transmitter power level to the needs of the moment. A bias supply for the amplifier is built in. The single 250-volt source handles the exciter stages and final amplifier screen. It should have good regulation. The oscillator screen voltage should be regulated if the transmitter is to be used for c.w., or if the ultimate in oscillator stability is wanted. Otherwise Pins 1 and 3 of the power plug, P_2, can be connected together.

The 4-pin fittings on the exciter and amplifier, J_5 and J_6, are wired so that meters for the final grid and screen current can be connected externally. The meters are not in the photographs, as they are mounted on a separate panel in the W1AW setup.

The filament transformers T_1 and T_2 connected back-to-back give isolation from the a.c. line for the bias supply, and take care of the heaters of the transmitter. The blower motor, B_1, comes on whenever the primary of T_1 is energized. The switch S_4 is connected externally and does not appear in the photographs.

The exciter heaters should be operated at 6.3 volts, but the amplifier tube should run at 6.0, plus or minus 5 percent. With today's line voltages often running over 120, a filament transformer rated at 6.3 volts may give as much as 7. This is much too high for the 250 series tubes, so RFC_4 was introduced to perform a dual role. Mainly, it drops the heater voltage to the desired level, but the r.f. isolation of the heater circuit it also provides certainly does no harm. Because the choke function is not critical, the wire size and/or number of turns can be varied to give 6.0 volts at the 4CX250 tube terminal, with your average line voltage. It is well to measure this with an a.c. meter of known accuracy; vacuum-tube voltmeters are notoriously inaccurate on a.c. readings. The choke was not in place when the pictures were made. It is mounted alongside the tube socket, directly in the air stream from the cooling fan, so heating is no problem with wire sizes as small as No. 28.

Modulation of the transmitter for voice or tone (the latter is used in W1AW bulletin and code-practice transmissions) requires an external audio unit of at least 150 watts output. The audio choke, L_{15}, connected in the screen lead, is shorted by S_3 when c.w. or f.m. is used. Maximum plate voltage is 2000 for c.w. and 1500 for a.m., but the amplifier works well with voltages as low as 750. High voltage is brought in on a separate fitting, J_8. A similar fitting, J_7, inverted alongside the amplifier tube, is used to terminate the high-voltage feed to the shunt plate choke, RFC_3.

Building the Exciter

Layout of parts in the exciter should be fairly clear from the photographs. The principal dimensions for drilling the top surface of the 5 by 10 by 3-inch chassis given in Fig. 6-24. The builder will do well to check his components for minor variations before going ahead with the drilling.

Controls on the front wall of the chassis at the left are the crystal selector, S_1, lower, and the 6360 plate capacitor, C_9, above it. In the middle is the loading capacitor, C_{10}, and at the right is the tripler plate capacitor, C_8. On the rear wall, Fig. 6-13, we see the crystal sockets at the right, the cathode jack, J_1, in the center, the power connector, J_5, at the left, and below it the coaxial output connector, J_2.

The tubes, V_1, V_2 and V_3, are lined up back to front in that order. The oscillator plate and doubler grid coils, L_1 and L_2, are beside V_1. The doubler plate and tripler grid coils, L_4 and L_5, are between V_1 and V_2. All are slug-tuned. The 144-Mc. coils are supported mainly on their own leads. The tripler and amplifier plate-tuning capacitors, C_8 and C_9, are alongside their respective tubes. They and the crystal switch are driven by extension shafts and couplings.

Liberal use is made of tie-point strips for mounting small parts, where they are not connected directly to other components that will support them. Disk-ceramic capacitors are used for bypassing in circuits up through 48 Mc. The 144-Mc. circuits (except the heater circuit of

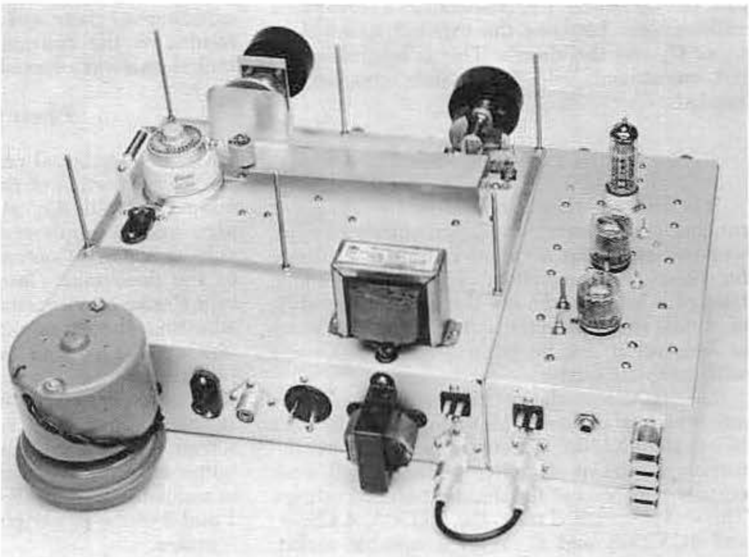


Fig. 6-19—Rear view of the complete transmitter, with amplifier shield removed.

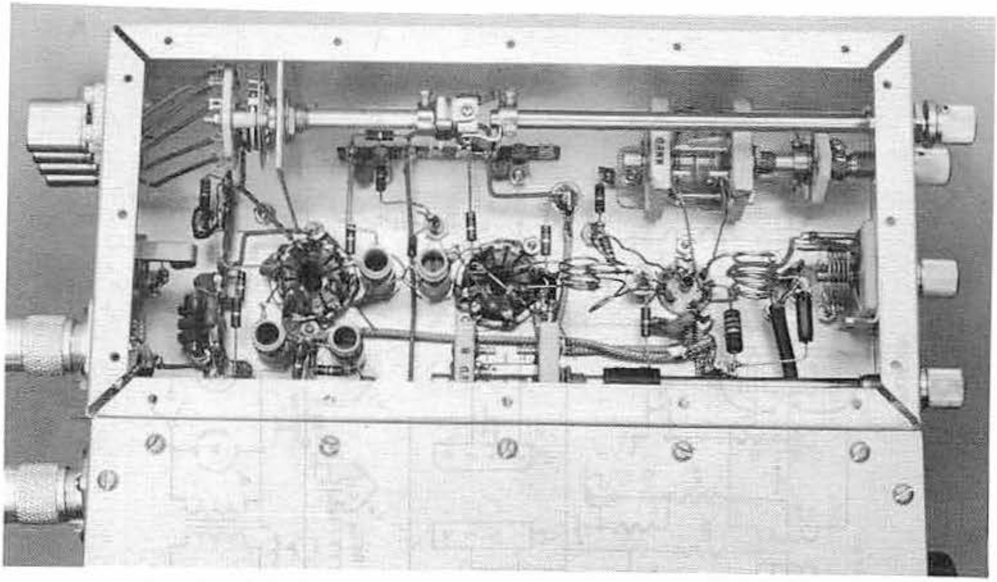


Fig. 6-20—Interior view of the 144-Mc. exciter. The oscillator portion is at the left. Note that all stages are inductively coupled, for maximum protection against spurious frequencies in the output.

V_3) use button-mica or ceramic feedthrough capacitors, as disk ceramics are unreliable at this and higher frequencies. During initial testing of the exciter instability in the output stage was traced to the presence of r.f. in the heater circuits. The heater bypass on Pins 4 and 5 of V_3, a 100-pf. ceramic of the "dogbone" variety, with ¼-inch leads, is series-resonant in the 144-Mc. region. This turned out to be a simple and effective way of getting the heater down to ground potential for r.f., and thus stabilizing the 6360 stage.

Heater leads are made with shielded wire. All power leads can be made this way, though it is not mandatory. Heater voltage for the exciter is brought through the side wall from the transformer, T_1, in the amplifier compartment on a feed-through capacitor, C_{20}. Plate power comes in via the 4-pin connectors, J_5 and P_2, which have separate terminals for the oscillator screen, so that this element can be supplied with regulated voltage if desired.

Adjustment

The exciter should be tested with no more than 250 volts from the supply. Less can be used, and the builder can play safe by inserting a 5000-ohm 10-watt resistor in series with the voltage source temporarily. This will prevent tube damage in case of malfunction, and it will protect the supply in case of a d.c. short.

Start with voltage only on the oscillator plate and screen, leaving other power leads disconnected temporarily. Listen for the oscillator on 8 Mc., or on 24, 48 or 144 Mc., whichever of these frequencies is available for receiving. The note should be a pure crystal tone, and the frequency should vary little or none as L_1 is tuned. The value of C_1 may require change for some inactive crystals, though ordinarily the 10 pf. specified will be satisfactory.

An r.f. indicator is now needed. This can take many forms. A 2-volt 60-ma. pilot lamp with a ½-inch diameter loop soldered across it can be hung over the end of L_1, and the core position adjusted for maximum lamp brilliance. A grid-dip meter in the output-indicating position may be coupled to L_1 for this test. The latter is preferred, since it provides a check on the frequency of the output, which should be in the 24-Mc. region, three times the crystal frequency.

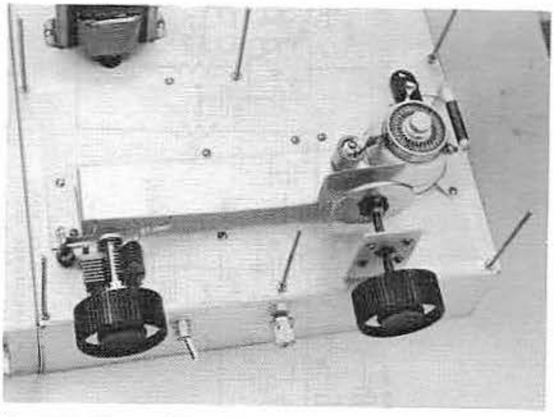


Fig. 6-21—The amplifier plate circuit is mainly a piece of flashing copper. It is grounded for d.c., making for safety in operation and ease of construction. Tuning is by means of a disk capacitor on a brass lead-screw, right. Plate voltage is shunt-fed through the r.f. choke, upper right.

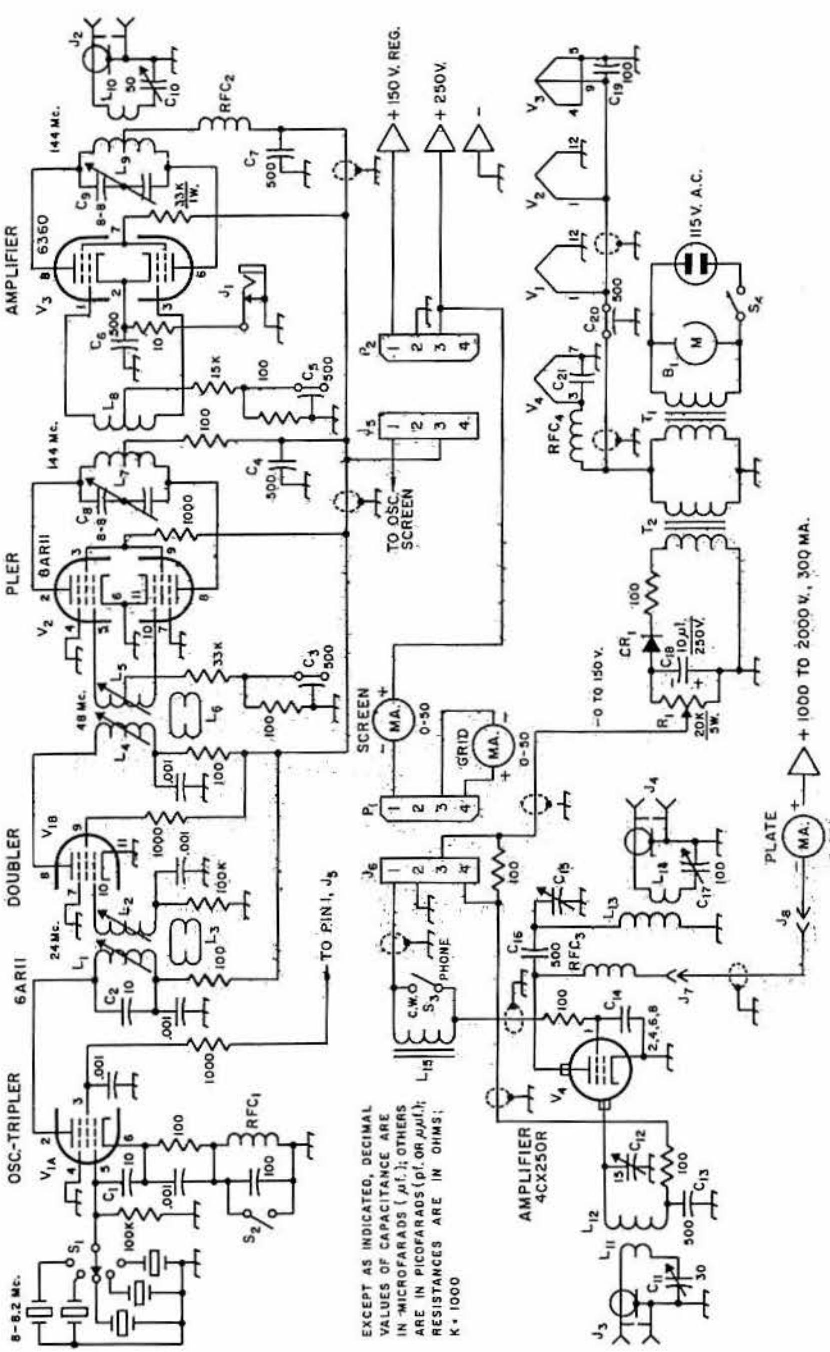

OSC.-TRIPLER
6ARII
DOUBLER
PLER
AMPLIFIER
AMPLIFIER 4CX250R
TO PIN 1, J5
TO OSC. SCREEN
+ 150 V. REG.
+ 250V.
SCREEN
GRID
PLATE
115 V. A.C.
+ 1000 TO 2000 V., 300 MA.
-0 TO 150 V.
PHONE
C.W.
EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μf.); OTHERS ARE IN PICOFARADS (pf. OR μμf.); RESISTANCES ARE IN OHMS; K = 1000

Fig. 6-22—Schematic diagram and parts information for the 500-watt 144-Mc. transmitter.

B_1—Blower, 16 c.f.m. or more.
C_1, C_2—10-pf. dipped silver-mica capacitor.
C_3, C_5, C_{20}—500-pf. feed-through capacitor.
C_4, C_6, C_7, C_{13}—500-pf. button-mica capacitor.
C_8, C_9—Miniature split-stator variable, 8 pf. per section (Bud LC-1659). Do not ground rotor.
C_{10}—50-pf, miniature variable (Hammarlund HF-50).
C_{11}—30-pf. mica trimmer.
C_{12}—15-pf. miniature variable (Hammarlund HF-15).
C_{14}, C_{21}—Bypass built into air-system socket (Eimac SK-620).
C_{15}—Disk-type variable; see text and photos.
C_{16}—500-pf. 5000-volt transmitting capacitor (Centralab 858S-500).
C_{17}—100-pf. variable (Hammarlund MC-100).
C_{18}—10-uf. 250-volt electrolytic.
C_{19}—100-pf. ceramic, with ¼-inch leads.
CR_1—400-volt p.i.v. diode, 100 ma. or more.
J_1—Closed-circuit jack.
J_2, J_3, J_4—Coaxial connector, SO-239.
J_5, J_6—4-pin male power connector.
J_7, J_8—High-voltage feedthrough connector (Millen 37501).
L_1, L_2—3.1 to 6.7 μh, slug-tuned, ceramic form (Miller 4405). Remove 2 turns from L_2.
L_3—Figure-8 loop of No. 24 enamelled wire around slug end of L_1 and L_2.
L_4—1.5 to 3.2 μh. slug-tuned, ceramic form (Miller 4404).
L_5—11 turns No. 24 enamel on ⅜-inch iron-slug ceramic form (Miller 4400), center-tapped.
L_6—Figure-8 loop No. 24 enamel around slug end of L_4 and center of L_5.
L_7—4 turns No. 20 Nyclad, ½-inch diam., ⅜ inch long, center-tapped. Leads are 1 and 1½ inches long.
L_8—4 turns No. 20 Nyclad, ½-inch diam., center-tapped, with 1¼-inch leads. Bend outer turn on each end outward at 45-degree angle, and insert middle two turns about ¼ inch into middle of L_7. See text and photo.
L_9—5 turns No. 18 tinned, ½-inch diam., ⅝-inch long, center-tapped, with 1¼-inch leads.
L_{10}—¾ turn insulated hookup wire, 1-inch diam., around center of L_9.
L_{11}—Loop of No. 18 3 inches long. Adjust shape and position for maximum grid drive; see text.
L_{12}—Copper strip 4 by ¾ inches; See Figs. 6-16 and 19.
L_{13}—Copper strip 9¾ inches long with 1¾-inch tab bent up for fixed plate of C_{15}. See Figs. 6-16 and 19.
L_{14}—Loop of copper 5/16 by 6 inches. Bend as per Fig. 6-19 and adjust position with respect to L_{13} for best output.
L_{15}—10-hy. 50-ma. filter choke.
P_1, P_2—4-pin female connector.
R_1—20,000-ohm 5-watt control.
RFC_1—500-μh. r.f. choke.
RFC_2—Single-layer v.h.f. choke, 1.3 to 2.7 μh. (Millen 34300, 2.7 uh. used).
RFC_3—No. 22 Nyclad closewound 1 3/16 inch on ¼-inch Teflon rod; about 2.2 μh.
RFC_4—Approx 5 feet No. 26 Nyclad closewound on ¼-inch Teflon rod. Vary wire size and/or number of turns to drop heater voltage to 6.0 at tube socket. See text.
S_1—Single-pole 5-position wafer switch.
S_2—Toggle switch mounted close to RFC_1. (Not in transmitter as shown.)
S_3, S_4—Toggle switch. (S_4 not in transmitter as shown.)
T_1—6.3-volt 6-amp. filament transformer.
T_2—6.3-volt 1.2-amp filament transformer.

A high-resistance voltmeter or v.t.v.m. can be used to measure relative grid voltage at the cold end of L_2. A simple wavemeter (see Fig. 11-3) may be used to check the approximate frequency, as the grid voltage will dip sharply as the wavemeter is tuned through the oscillator output frequency. Tune the cores in L_1 and L_2 for maximum drive to V_{1B}.

Now apply plate and screen voltage to V_{1B}, and check similarly for 48-Mc. output. Any of the above methods may be used, and in addition we have provision for measuring tripler grid current built into the exciter. Connect a low-range milliammeter from the exposed terminal of C_3 to ground, and tune all core studs for maximum grid current. This should be about 1 ma, though more is fine if you can get it. Check with a wavemeter to be sure that the energy in L_4 and L_5 on the 6th harmonic of the 8-Mc crystal, and on no other frequency.

Next connect the meter from the exposed terminal of C_5 and ground, and apply plate and screen voltage to V_2. Tune C_8 for maximum amplifier grid current, which should be around 2 ma. Check with a wavemeter to be sure that the drive is on the 18th harmonic of the 8-Mc. crystal frequency. A check on the need for neutralization, if any, should now be made. Tune the plate circuit of V_3 slowly through resonance while watching the amplifier grid current. There should be no drop in grid current as this is done. A downward flicker would indicate feedback, which would require neutralization. This is easily done with a 6360 by soldering ½-inch pieces of insulated wire to the grid terminals, Pins 1 and 3. Bend the ends until they are adjacent to the plate terminals, 6 and 8, respectively, adjusting their position until the change in grid current as the plate circuit is tuned is eliminated. Normally the 6360 does not require neutralization, and the exciter shown here was completely stable without it, after the heater bypass capacitor, C_{19}, was installed.

Now apply plate and screen voltage to the 6360, and connect a dummy load to J_2, preferably through a power-indicating s.w.r. bridge. Adjust C_9 and C_{10} for maximum power output, which should be 6 to 8 watts, with a 250-volt supply. Now, using a crystal frequency near the middle of the range over which you will normally operate, tune all adjustments through C_8 for maximum amplifier grid current, and C_9 and C_{10} for best output.

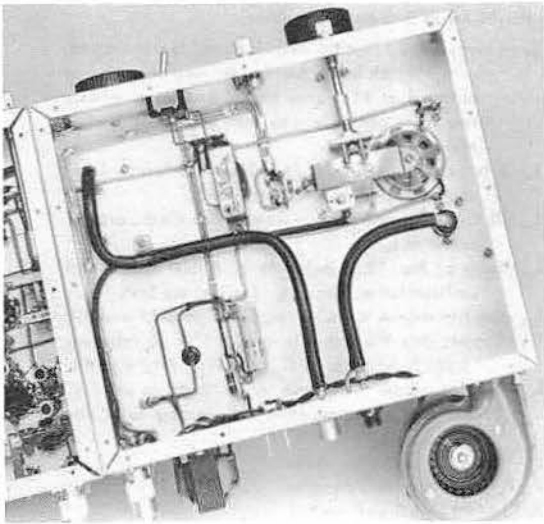


Fig. 6-23—Interior of the amplifier compartment. The grid inductance, upper right, is a strip of copper supported at its left end by a button-mica capacitor and fastened to the grid terminal of the tube socket at the right end. The long piece of coax, left, brings the amplifier output to the rear of the chassis. The shorter section is used as a shielded high-voltage lead. Bias supply components are in the left center portion of the chassis.

It will be seen that input to the 6AR11 pentodes runs only about the rated plate dissipation for the tubes. This makes for long tube life and trouble-free operation. The 6360 also operates conservatively, yet its output is adequate to drive the final amplifier. The exciter may also be used as a low-powered transmitter, and it is well-adapted to portable work, since its total drain is only a little over 100 ma. at 250 volts. The output stage can be modulated with 6 to 10 watts of audio, or keyed in J_1 for c.w. work.

Amplifier Construction

The final stage is built on a 10 by 12 by 3-inch aluminum chassis, which when fastened to the 5 by 10-inch exciter makes a complete 10 by 17-inch assembly that can be rack mounted. Our photographs were made before the panel was added, in the interest of clarity. The construction is extremely simple, and with the drawings of the grid and plate inductances, Fig. 6-25, the builder should have little trouble in duplicating the original. Arrangement of parts, other than in the r.f. circuits, is not important.

The Eimac SK-620A socket has a shield ring enclosing the screen contacts, a feature that may contribute to the exceptional stability of this amplifier. Other air-system sockets leave the screen ring of the tube exposed, and this has been a factor in neutralizing problems encountered with various external-anode tubes of the 150-250 series in the past. The push-pull amplifier for 144 Mc. described elsewhere in this chapter required shield plates alongside the tube sockets, in order to achieve complete stability. Neutralization of this amplifier, if needed, is described at the end of this section.

The grid circuit is a short strap of copper, with its main portion about one inch away from the chassis. One end is supported on a button-mica capacitor, C_{13}, and the other on the grid terminal of the tube socket. The input coupling loop, L_{11}, is supported on a tie-point strip adjacent to grid line, the loop extending underneath the copper strip.

The main portion of the plate line and the stationary plate of C_{15} are a single piece of flashing copper. This is fastened directly to the chassis at the left end, as viewed in Fig. 6-21. Plate voltage is shunt-fed through RFC_3 to the tube anode. A copper strap wrapped around the anode supports the blocking capacitor, C_{16}, which is bolted to both the plate strap and the plate inductance, L_{13}. At the grounded end of L_{13} may be seen the series capacitor, C_{17}, and the output coupling loop, L_{14}. The nature of C_{16} is important. It must be a transmitting-type capacitor, capable of withstanding heat, high r.f. current and high voltage. The TV-type "doorknob" capacitor often used for this purpose on lower frequencies is definitely not recommended for 144-Mc. service.

When the amplifier was placed in service at W1AW it was found that vibration of the plate line caused by the blower motor was a source of operational difficulties, so a ceramic standoff was mounted near the middle of L_{13}, to support it more rigidly on the chassis. We silver-plated all the plate line components, but measurements made carefully before and after show only a perceptible improvement from the plating.

The movable plate of C_{15} is a 2¼-inch aluminum disk mounted on a ¼-20 brass lead screw. A matching nut soldered to a copper plate is bolted to an aluminum bracket to provide a bearing and electrical ground. When the panel is in place a tension spring can be added externally, by slipping it over the brass screw.

The shunt-feed r.f. choke, RFC_3, may be seen in a horizontal position beside the tube, level with the top of the anode. Its back end is connected to a high-voltage feedthrough, J_7. Under the chassis the matching portion of J_7 is connected to a similar fitting, J_8, on the back wall of the chassis, by means of coax used in lieu of high-voltage shielded wire. Another run of coax connects the output fitting, J_4, with the hot end of the output coupling loop, L_{14}.

The shield cover for the amplifier is a standard 7 by 12-inch chassis, notched to pass the shafts of C_{15} and C_{17}, and held in place by wing nuts atop 6 3½-inch 6-32 threaded brass rods. These are fastened at the corners, and at the midpoint of each long side, with hex nuts above and below the main chassis surface. If you do your own metal-work you may be able to make a better shielded plate line than this; the dimen-

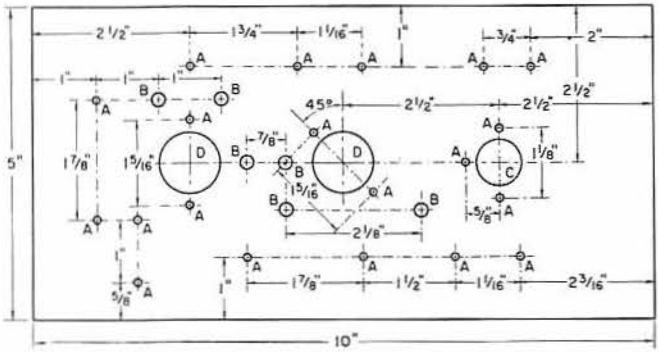


Fig. 6-24—Hole layout for the exciter chassis. Hole sizes are A—1/8 inch, B—1/4 inch C—3/4 inch and D—1 inch. Because of variations in parts sizes, the builder should check with his components before drilling to these dimensions.

sions of ours were dictated by available chassis sizes. The main chassis was polished with emery paper and steel wool along the surface that makes contact with the cover. Good electrical contact is important here, and also at the grounded end of the plate line. The folded-over end of L_{13} is clamped to the chassis with a metal strip and two screws and nuts.

For effective cooling with a small blower it is important that there be very low air leakage out of the main chassis, except up through the tube socket. To this end, the holes in the corners of the chassis, the overlaps at the corners, and all holes made in mounting the various parts were sealed with plastic cement. A screened hole in the top cover allows the warm air to flow out of the plate compartment directly over the tube. A tight-fitting bottom cover is important for good cooling, perhaps more than for shielding.

The built-in bias supply, the audio choke in the screen lead, and the various components other than those in the r.f. circuits can be placed almost anywhere that suits the builder's fancy.

Firing Up

The first step in placing the amplifier in service is to check the grid circuit. Input coupling is best adjusted with a standing-wave bridge connected in the line between J_2 and J_3. A milliammeter should be between terminals 3 and 4 of P_1, to read amplifier grid current. The object now is to obtain optimum coupling into the amplifier.

Apply power to the exciter, which also activates the amplifier bias supply. Leave the screen meter disconnected for the present, so that there will be no voltage on the amplifier screen. Adjust the exciter tuning and loading for maximum amplifier grid current. Now adjust C_{11} and C_{12} for minimum reflected power on the s.w.r. bridge. If this is not zero, try various positions of L_{11} with respect to L_{12}, readjusting their capacitors each time for lowest reflected power. The best power transfer between exciter and amplifier will occur at this point.

Adjust the bias control so that the amplifier grid current is 10 ma. or less, and apply plate and screen voltage to the amplifier. Be sure that the amplifier is loaded at all times, to prevent excessive screen current. Satisfactory operation should be possible with plate voltages as low as 700, with 250 volts on the screen. If lower plate voltage is used for initial testing, the screen voltage should be dropped also, to keep screen current below about 30 ma. Keep a 50-ohm dummy load connected to J_4 at all times, and be sure that C_{15} and C_{17} are adjusted so that power is being delivered to the load. Tube damage is more likely to develop from excessive screen dissipation than from anything that can

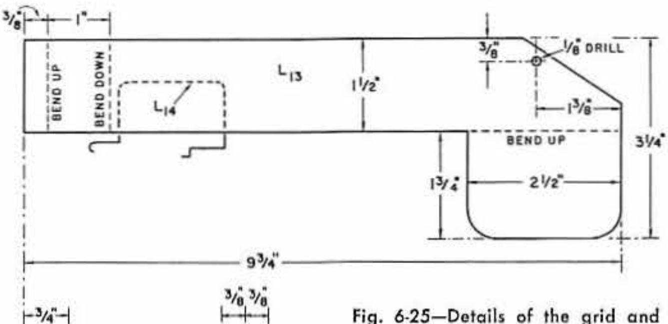


Fig. 6-25—Details of the grid and plate inductances. The material is flashing copper, though brass or copper of heavier gauge may be used. Silver plating is optional. Approximate positions of coupling loops are shown in broken lines.

happen to the plate in normal service, so keep a close watch on the screen meter, and be sure that dissipation is kept below 10 watts.

Adjust the position of L_{14} with respect to L_{13} for maximum output, readjusting the tuning and loading capacitors, C_{15} and C_{17} with each change in coupling. The tuning and the position of the coupling loop will change with various plate voltages, so final adjustment should be made with the plate voltage at the point where maximum efficiency is desired. If an accurate bridge or wattmeter is available, it should indicate operating efficiency in excess of 65 percent. Power output well over 300 watts was measured at 2000 volts, and 200 watts at 1500 volts, with inputs of 500 and 300 watts, respectively.

The amplifier can be run under a wide range of plate and screen voltages, bias and driving power, so long as none of the maximum ratings for the various elements is exceeded. With fixed screen supply, best efficiency will be obtained by juggling the grid bias, checking output meanwhile. Keep the final plate current below 250 ma. and the screen current under 30. Screen current will be progressively lower as the plate voltage is raised, and may even go negative at plate voltages above 1000 or so, particularly with low drive. If a separate variable screen supply is used, there may be some advantage in using voltages above 250, so long as the screen dissipation is kept low.

Neutralization can be added, if necessary, as follows: A feed-through bushing (National TPB) is mounted under L_{13}, so that it projects through the chassis under L_{12}. A loop of wire about ½ inch on a side is connected from the bushing rod to the chassis, under L_{12}. A brass capacitor plate about ½ by 1 inch is soldered to the top, under L_{13}. Vary the position of the loop with respect to L_{12}, and the plate with respect to L_{13}, to achieve minimum r.f. feedthrough, with the exciter running and the amplifier having only heater voltage applied. Check with a sensitive r.f. indicator coupled to J_3.

A 40-WATT TRANSMITTER FOR 220 MC.

The crystal-controlled transmitter shown in Figs. 6-26 and 6-28 will run 30 to 40 watts at 220 Mc. Referring to Fig. 6-27, a simple overtone oscillator circuit uses one half of a 12AT7 dual triode. The crystal may be between 8.15 and 8.33 Mc. or 24.45 and 25 Mc. In either case, the frequency of oscillation is in the latter range, as the crystal works on its third overtone. The second half of the 12AT7 is a tripler to 73 to 75 Mc. This stage has a balanced plate circuit, so that its output may be capacitively coupled to the grids of a second 12AT7, working as a push-pull tripler to 220 Mc.

Though the oscillator-tripler circuit works well as shown, a revision of the transmitter was later made to use the 6CX8 oscillator-multiplier circuit of Figs. 6-14. The latter arrangement provides better oscillator stability and more grid drive than the overtone circuit shown in Fig. 6-27. The circuit remains the same from the plate of V_{1B} on.

The plate circuit of the push-pull tripler is inductively coupled to the grid circuit of an Amperex 6360 dual tetrode amplifier that runs straight through on 220 Mc. Similar inductive coupling transfers the drive to the grid circuit of the final amplifier stage, an Amperex 6252 dual tetrode. This tube is a somewhat more efficient outgrowth of the 832A, which may also be used, though with lower efficiency and output. Base connections are the same for both tubes.

The grid return of the 6252 is brought out to the terminal strip on the back of the unit, to allow for connection of a grid meter. Both this point and the tip jack in the 6360 grid circuit have 1000-ohm resistors completing the grid returns to ground, so that operation of the stages is unaffected if the meters are removed.

Instability in tetrode amplifiers for v.h.f. service may develop as a result of the ineffective bypassing of the screen. In the case of the 6360 stage stable operation was obtained with no bypassing at all, while on the 6252 a small mica trimmer is connected directly from the screen terminal to ground. It is operated near the minimum setting.

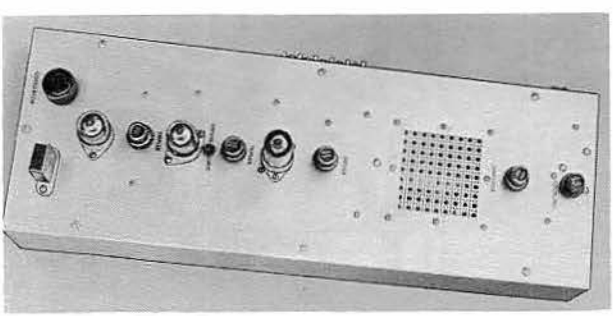


Fig. 6-26—Top view of the 220-Mc. transmitter. Final amplifier tube is inside the chassis, below the screened ventilation hole. Power connections, keying jack and output terminal are on the back of the chassis

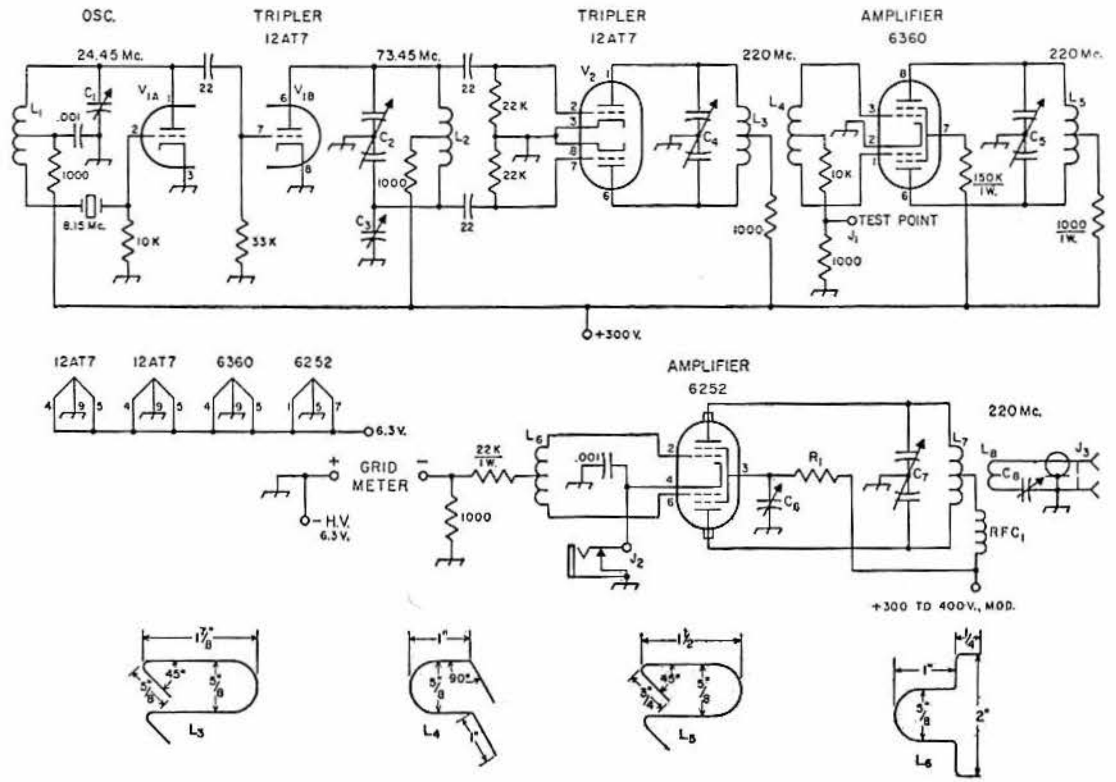


Fig. 6-27—Schematic diagram and parts information for the 220-Mc. transmitter. Capacitor values below 0.001 μf. are in pf. Resistors ½ watt unless specified. The 6CX8 oscillator circuit of Fig. 6-14 may be substituted for that shown above, with improved stability and drive.

C_1—50-pf. miniature variable (Hammarlund MAPC-50-B).

C_2 C_4, C_5—8-pf. miniature butterfly variable (Johnson 160-208).

C_3, C_6—3-30-pf. mica trimmer.

C_7—Butterfly variable, 1 stator and 1 rotor (Johnson 167-21, with plates removed).

C_8—15-pf. miniature variable (Hammarlund MAPC-15-B).

J_1—Tip jack, insulated.

J_2—Closed-circuit phone jack.

J_3—Coaxial chassis fitting, SO-239.

L_1—15 t. No. 20 tinned, ½-inch diam., 1 inch long (B & W Miniductor No. 3003). Tap at 4 turns from crystal end; see text.

L_2—12 t. No. 18 tinned, ½-inch diam., 1 inch long, center-tapped.

L_3, L_4, L_5, L_6—U-shaped loops No. 18 enam., center-tapped. Dimensions given on drawing.

L_7—2 t. No. 14 enam., 1-inch, 1-inch diam., leads ⅝ inch long. Center-tapped, space turns ½ inch apart.

L_8—1 t. No. 18 enam., inserted between turns of L_7. Cover with insulating sleeving.

R_1—23,500 ohms, 2 watts. (Two 47,000-ohm 1-watt resistors in parallel.)

RFC_1—25 t. No. 28 enam. on 1-watt high-value resistor.

Construction

The transmitter is built on an aluminum plate 6 by 17 inches in size. This screws to a standard chassis of the same dimensions, which serves as both shield and case. Cut-outs about three inches square are made in the chassis and base plate, above and below the tube, to allow for ventilation. These openings are fitted with perforated aluminum or screening to preserve shielding. The case should be equipped with rubber feet, to avoid marring the surface it rests on, and to allow air circulation around the tube.

The tube sockets and all the controls except the tuning capacitor of the oscillator are mounted along the center line of the cover plate. The 220-Mc. stages are inductively coupled, using hairpin loop tank circuits the dimensions of which are given in Fig. 6-27. The tuning range of these circuits is affected by the widths of the loops as well as their length, so some variation can be had by squeezing the sides together or spreading them apart.

It is important that the method of mounting the 6252 socket be followed closely. An aluminum bracket about 2⅞ inches high and 4 inches wide supports the socket. Note that the socket and tube are on the *same* side of the plate. Holes are drilled in the plate in line with the control grid terminals to pass the grid leads.

These holes are ⅜-inch diameter, and are equipped with rubber grommets to prevent accidental shorting of the grid leads to ground. The shape of the grid inductance should be such that it leads pass through the centers of the holes. The socket is supported on 5⁄16-inch metal pillars. It may be necessary to bend the socket lugs slightly to keep them from shorting to the mounting plate. The heater lead comes to the top of the plate, and the cathode lead bends around the bottom of it.

Power leads are made with shielded wire, and are brought out to a terminal strip on the back of the chassis. These leads and the coax to the output connector should be long enough so that the plate on which the transmitter is built can be lifted off the chassis and inverted as shown in the photograph.

Adjustment

Initial test should be made with a power supply that delivers no more than 250 volts, and as little as 150 to 200 volts can be used. If the voltage is more than 250, insert a 5000-ohm 10-watt resistor in series with the power lead temporarily. Plate voltage should be applied to the various stages separately, starting with the oscillator, making sure that each stage is working correctly before proceeding to the next.

A milliammeter of 50- to 100-ma. range should be connected temporarily in series with the 1000-ohm resistor in the oscillator plate lead. When power is applied the current should be not more than about 10 ma. Rotate C_1 and note if an upward kick occurs, probably near the middle of the range of C_1. At this point the stage is oscillating. Lack of oscillation indicates too low feedback, or a defective crystal. Listen for the note on a communications receiver tuned near 24 Mc., if one is available. There should be no more than a slight change in frequency when a metallic tool is held near the tuned circuit, or when the circuit is tuned through its range. The note should be of pure crystal quality. If there is a rough sound, or if the frequency changes with mechanical vibration, the oscillator is not controlled by the crystal. This indicates too much feedback, and the tap on the coil, L_1, should be moved near the crystal end.

The proper amount of feedback is the lowest tap position that allows the oscillator to start readily under load. If 24-Mc. crystals are used, the tap can be lower on the coil than with 8-Mc. crystals. When 8-Mc. crystals are operated on the third overtone, as in this case, the frequency of oscillation may not be exactly three times that marked on the crystal holder.

Now apply plate voltage to the second half of the 12AT7, again using a temporary plate meter connected in series with the 100-ohm decoupling resistor that feeds plate power to L_2. Current will be about 10 ma., as with the oscillator. Tune C_2 for maximum output. This can be determined by brilliance indication in a 2-volt 60-ma. pilot lamp connected to a 1-turn loop of insulated wire coupled to L_2. Check the frequency of this stage with a wavemeter.

Now connect a low-range milliammeter (not more than 10 ma.) between the test point, J_1, and ground. Apply power to the push-pull tripler, again using a temporary milliammeter connected in the lead to the plate coil, L_3. Tune the plate circuit for maximum indication on the grid meter. Plate current will be about 20 ma. Adjust the position of L_3 with respect to L_4 for maximum grid current. Now go back over all previous adjustments and set them carefully for maximum grid current. Adjust the balancing padder, C_3, retuning C_2 each time this is done, until the combination of C_2 and C_3 that gives the highest grid current is found. Check the frequency to be sure that the stage is tripling to 220 Mc.

Now apply power to the 6360 plate circuit, again using the temporary meter to check the current. Connect the low-range milliammeter between the grid-metering terminal on the connector strip and ground. Set the screen trimmer, C_6, near minimum, and tune the 6360 plate circuit for maximum grid current. With 300 volts

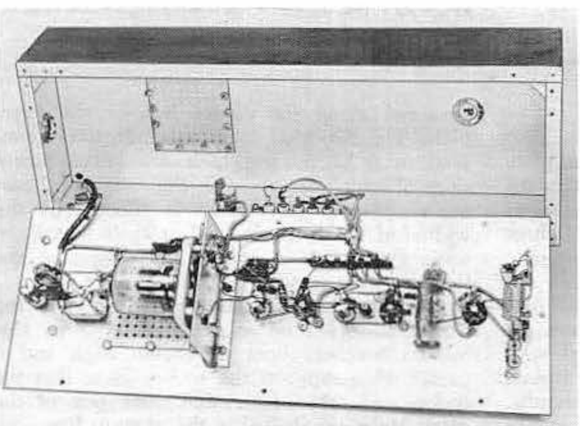


Fig. 6-28—Interior view of the 220-Mc. transmitter. All r.f. components are mounted on an aluminum plate, which is screwed to the top of a standard 6 x 17-inch chassis.

The crystal oscillator is at the far right. Next to the left is the first tripler plate coil, mounted over its trimmer, with the mica balancing padder, C_3, above. The 12AT7 tripler, the test point, J_1, the tuning capacitor C_4, the tripler plate and amplifier grid loops, L_3 and L_4, the 6360 socket, the 6360 plate and amplifier grid loops, the 6252, and its tuned circuits follow in that order.

on the preceding stages, it should be possible to get at least 4 ma. Adjust the spacing between L_5 and L_6 carefully for maximum grid current, retuning C_5 each time this is done. Plate current should not exceed 55 ma.

Check for neutralization of the final amplifier by tuning C_7 through resonance while watching the grid-current meter. If there is no change, or only a slight rise as the circuit goes through resonance, the stage is near enough to neutralization to apply plate power. The 6252 has built-in cross-over capacitance, intended to provide neutralization in the v.h.f. range, so it is likely to be stable at this frequency. If there is a downward kick in the grid current at resonance, adjust the screen trimmer until it disappears. If best neutralization shows at minimum setting of the screen trimmer it may be desirable to eliminate the trimmer.

With an antenna or dummy load connected at J_3, final plate voltage can be applied. Tune the final plate circuit for maximum output, with a meter of 100 ma. or higher range connected to read the combined plate and screen current. This meter may be connected in the power lead, or it can be plugged into the cathode jack. In the latter position it will read the combined plate, screen and grid currents. Tune for maximum output and note the plate current. If it is much over 100 ma., loosen the coupling between L_7 and L_8. The input should not be over 50 watts at this frequency.

A final check for neutralization should now be made. Pull out the crystal or otherwise disable the early stages of the transmitter. The grid current and output should drop to zero. If they do not, adjust the screen trimmer until they do. Make this test only very briefly, as the tubes will draw excessive current when drive is removed. When perfect neutralization is achieved, maximum output will be found at a setting of C_7 at which plate current is at a minimum and grid current at maximum.

Operation

All stages should be run as lightly as possible, for stable operation and long tube life. No more than 300 volts should be run on the exciter stages, and if sufficient grid drive can be obtained, lower voltage is desirable. The 6360 stage runs with rather low drive, and its efficiency is consequently poor, but it delivers enough power to drive the 6252, even when run at as low as 250 volts, if all stages are operating as they should.

Observe the plates of the tubes when the transmitter is operated in a darkened room. There should be no reddening of the plates. If one side of any of the last three stages shows red and the other does not it is evidence of unbalance. This can usually be corrected by adjustment of the balancing trimmer, C_3, in the first tripler plate circuit. Lack of symmetry in lead lengths or unbalanced capacitance to ground in any of the r.f. circuits may also lead to lopsided operation.

Though the 6252 is rated for up to 600 volts on the plates, it is recommended that no more than 400 be used in this application, particularly if the stage is to be modulated for voice work. In the latter case, the plate-screen current of the 6252 is run through the secondary of the output transformer on the modulator having an output of 20 watts or so.

EXCITERS AND AMPLIFIERS

The preceding r.f. units in this chapter are mainly for use as complete transmitters, requiring only power supplies and speech equipment, if used, to make them ready for service. This does not preclude lifting ideas from them for use in exciters or amplifiers of one's own design, of course. The items to follow will suggest more such ideas. They were designed for use with no specific companions units. Some could be tied in with parts of transmitters just described, or with one another—or they can be adapted for use with entirely different equipment, so long as the power levels match.

Perhaps you have built the 144-Mc. transmitter of Fig. 6-13, and are now looking for an inexpensive amplifier to boost your power to about the 100-watt level. The 829B amplifier of Fig. 6-33 may be what you need. If you built both the 50- and 144-Mc. units and now want a big signal on both bands you're a good prospect for the powerhouse amplifiers of Fig. 6-38. It is worth noting that though these amplifiers are designed for service at 1 kilowatt input, they work well at much lower levels. They are useful for any mode.

Would-be v.h.f. sideband operators will find the heterodyne unit of Fig. 6-29 of interest. Though shown for operation with a 50-Mc. s.s.b. exciter, it requires only suitable modification of the mixer cathode circuit and the oscillator-multiplier frequencies to work with h.f. sideband gear.

A HETERODYNE EXCITER FOR 144 MC.

This 3-tube r.f. unit was built primarily to develop s.s.b. drive on 144 Mc. from a 50-Mc. sideband source, but it can be used advantageously in several other ways, since it will reproduce a 50-Mc. signal of any type on 144. If the 50-Mc. rig is v.f.o. controlled the 144-Mc. signal

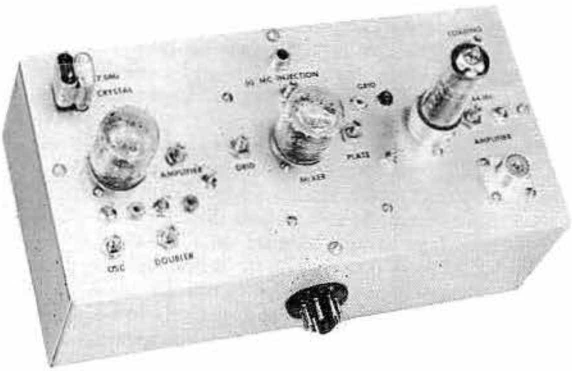


Fig. 6-29—Conversion unit for duplicating a 50-Mc. signal on 144 Mc. Two tubes at the left and center comprise the oscillator, multiplier, amplifier and mixer stages. At the right is the 144-Mc. output amplifier. Provision is made for metering all stages by means of tip jacks and test points. Note crystals taped together to prevent loss of the one not in use.

will have the same stability as the 50-Mc. one, and the dial calibration is the same for both bands.

The conversion process is similar to that involved in a 144-Mc. receiving converter intended for use with a 50-Mc. tunable receiver. In fact, the first stages of this unit could be used in that way, following the general layout of the 144-Mc. transverter described in Chapter 7. The heterodyning signal is on 94 Mc. This is fed to the grids of a mixer, the cathodes of which are driven on 50 Mc. by the exciter used for work on that band. The two frequencies add, providing 144-Mc. output in the mixer plate circuit. (The mixer also would give 94 *minus* 50, or 44 Mc., but this product is rejected by the tuned circuits.)

Circuit and Layout

The schematic diagram, Fig. 6-30, and the bottom view, Fig. 6-31, may be "read" from left to right. First we have a simple triode crystal oscillator, V_{1A}, on 47.0 or 47.5 Mc., depending on the crystal, Y_1. The 47-Mc. plate coil, L_1, and its tuning capacitor, C_1, are in the upper left corner of the picture. The second triode of V_1 is a doubler to 94 or 95 Mc. Its tuned circuit, L_2C_2, is seen adjacent to the oscillator, but with its axis perpendicular to L_1. Inductively coupled loosely to L_2 is L_3, the grid circuit of a 94-Mc. amplifier, V_{1C}. On the right side of the first tube is the amplifier plate circuit, L_5C_3, straddled by L_6, the split grid coil of the mixer, V_2. Below the mixer tube is the 50-Mc. input circuit connected to the mixer cathodes.

From here on the layout and circuit look like any other low-powered 144-Mc. transmitter. The amplifier grid coil, L_8, is purposely made too small to resonate in the 144 to 146-Mc. region with the input capacitance of the 6360, V_3. Being on the high-frequency side of resonance, it offers little feedback coupling to the output circuit, even though there is no shielding between the two. The amplifier plate circuit, L_9C_7, is at the far right. Output is taken off through a series-tuned link, $L_{10}C_8$.

Positioning of the various coils is important. Note that coils are placed so that unwanted coupling between circuits is kept down, even with a fairly compact layout. It is suggested this principle be followed unless the builder is willing to cope with a new set of neutralization problems.

The oscillator and doubler circuits are standard practice. In the grid circuit of the 94-Mc. amplifier, the input capacitance of the 6M11 pentode was too high to permit resonating L_3 at 94 Mc. in the usual way. Some checks with a variable series capacitor showed that a coil the same size as in the previous plate circuit could be resonated at 94 Mc. with about 10 pf. in series, so the fixed capacitor shown in Fig. 6-24 was used. Only a small amount of energy is needed for the mixer grids, so neither the tuning nor the coupling between circuits is at all critical.

Getting the 94-Mc. amplifier to operate in a stable manner is mainly a matter of achieving ground potential for r.f. at the screen. This is done with the series circuit, L_4C_4, the setting of which is not particularly fussy. Coupling between L_5 and L_6 should be adjusted to the minimum that will provide satisfactory output from the mixer. Make sure that both circuits actually tune, as it is possible to get enough output with one or the other not actually peaking. Best rejection of unwanted frequencies will not be assured unless the circuits are tuned to the desired frequencies.

Coupling between L_7 and L_8 should also be as loose as it can be and still provide adequate drive for the 6360. Drive requirements depend on the class of operation of the output amplifier. For anything but Class-C conditions adequate drive is very easily achieved. Here again, be sure that L_7 actually tunes *through* the desired frequency, in order that rejection of unwanted frequencies will be at a maximum.

Construction is on a standard 5 by 10-inch aluminum plate and 3-inch chassis. A layout drawing, Fig. 6-32, is given for those who wish to make an exact duplicate. To check every circuit during the adjustment phase of the project, an unusual combination of feedthrough bypasses and tip jacks is used. Oscillator plate current is measured by plugging a meter into J_1 and clip-

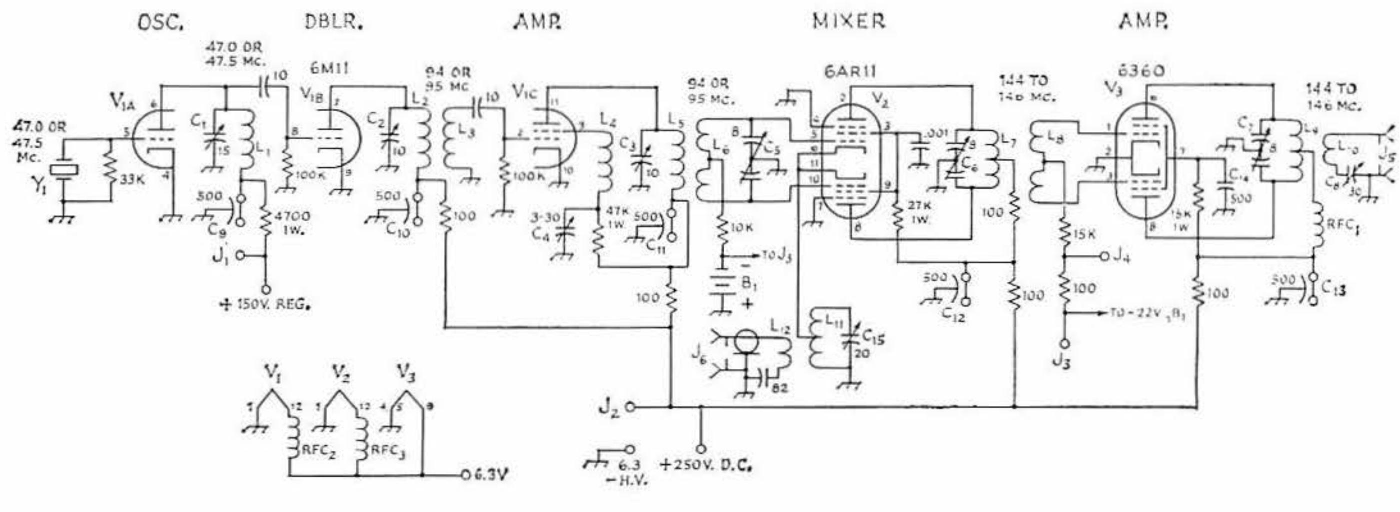


Fig. 6-30—Schematic diagram and parts information for the heterodyne unit. Decimal values of capacitance are in microfarads (μf.); others in picofarads (p.f.). Unless otherwise indicated, fixed capacitors are ceramic. Resistors are composition, ½-watt, unless specified.

B_1—22½-volt miniature battery (Eveready No. 412).
C_1—15-pf. miniature trimmer (Hammarlund MAC-15).
C_2, C_3—10-pf. miniature trimmer (Hammarlund MAC-10).
C_4—3-30-pf. mica trimmer.
C_5, C_6, C_7—8-pf. per section miniature butterfly (Hammarlund MACBF-8).
C_8—30-pf. miniature trimmer (Hammarlund MAC-30).
C_9, C_{10}, C_{11}, C_{12}, C_{13}—500-pf. feedthrough bypass (Centralab FT-500). Cover exposed ends with ⅜-inch lengths of spaghetti when not in use for metering.
C_{14}—500-pf. button-mica.
C_{15}—20-pf. miniature trimmer (Hammarlund MAC-20).
J_1, J_2, J_3, J_4—Insulated tip jack.
J_5—Coaxial receptacle, SO-239.
J_6—Phono jack or coaxial receptacle.
L_1—6¼ turns No. 24, ½-inch diam., 32 t.p.i. (B & W 3004).
L_2, L_3—3¾ turns, like L_1.
L_4—6 turns No. 20, ¼-inch diam., one inch long.
L_5—3 turns, like L_1.
L_6—2 turns each side, No. 20, ½-inch diam., 16 t.p.i., center-tapped. Make from single piece of B & W 3003, cutting all but one plastic strip. Leave ⅜-inch space at center.
L_7—4¼ turns No. 20, ½-inch diam., 16 t.p.i., center-tapped (B & W 3003).
L_8—Like L_6, except ½-inch space at center.
L_9—Like L_6, except $\frac{3}{16}$-inch space at center.
L_{10}—One turn insulated hookup wire inserted halfway into center of L_9. Leads are 1½ and 2 inches long.
L_{11}—5 turns No. 24, ½-inch diam., 32 t.p.i., tapped 2 turns from ground end (B & W 3004).
L_{12}—2 turns made from same piece as L_{11}, spaced 4 turns away; cold ends adjacent.
RFC_1—No. 28 enam. close-wound full length of ½-watt 1-meg. resistor. (And ready-made 144-Mc. r.f. choke is suitable.
RFC_2, RFC_3—Ferrite-bead r.f. choke, 1.3 μh. (National R-45-251) No. 24 enamel wound full length of 1-watt resistor also usable.
Y_1—Third-overtone crystals, 47.0 and 47.5 Mc. (International Crystal Mfg. Co. F-605). Tape together, opposite ends up, to prevent loss of one not in use.

ping to the exposed terminal of C_9. All other plate currents may be read by plugging one side of the meter into J_2 and connecting to C_{10}, for doubler plate current, C_{11}, for 94-Mc. amplifier plate current, C_{12}, mixer plate and screen current, or C_{13}, amplifier plate and screen current. Amplifier grid current, if any, is checked at J_3 and J_4. A table of operating conditions is given later.

Bias for the mixer and output amplifier is obtained from a small 22½-volt battery. Builders may prefer some other bias source, but the battery does the trick simply and inexpensively. There is no current drain, and it may even be charged a bit when the amplifier runs into Class-C conditions, so life should be long and voltage constant. Just be careful not to short out the battery when working on the unit.

The 6360 amplifier operated satisfactorily without external neutralization, but a small amount was added when a slight reaction on amplifier grid current was noted as the plate circuit was tuned through resonance with voltage off. The grid and plate leads are crossed over inside the 6360, providing inherent neutralization in the v.h.f. range, so only a tiny amount of additional capacitance is needed. A half-inch wire is soldered to each grid terminal, and bent over toward the adjacent plate terminal. The position is adjusted until reaction on amplifier grid current is eliminated. For circuit simplicity, this neutralization is not shown in Fig. 6-30.

Adjustment and Use

It should not be taken for granted that the heterodyning approach is for the sideband operator alone. Given any of the popular small 50-Mc. transmitters, homebuilt or commercial, this heterodyne unit will duplicate its signal on the 144-Mc. band at a comparable power level. You'll need no big batch of crystals or two separate v.f.o. units to give coverage of both bands. If you're a Technician or Novice at present, use only the 47.5-Mc. crystal in the oscillator-multiplier. Heterodyning from 50-Mc. frequencies will start your coverage at 145 Mc. A crystal that gives operation on 50.2 Mc. will put you on 145.2 Mc., and so on. A v.f.o. that covers 50 to 51 Mc. (not for Novice use, of course) will give you coverage of 145 to 146 Mc., which can be extended to 144 Mc. with the insertion of the 47-Mc. crystal at a suitable time.

The output stage of the conversion unit can be run as a linear amplifier for sideband, c.w. or a.m., or it can be driven into Class-C conditions for higher efficiency on c.w. Plate modulation may be applied in the usual way for high-efficiency a.m. service. The linear way will probably be the more attractive to most users, however, as it eliminates the heavy and power-consuming audio equipment. If your 50-Mc. rig is plate modulated, you can make provision for switching the audio power over from its final stage to that of the conversion unit.

Initially we ran the 50-Mc. energy into the mixer grids and applied the 94-Mc. injection to the cathodes, but it was easy to saturate the grids with the swinging drive from the 50-Mc. sideband rig. With the circuits swapped around as shown, the mixer takes the full output of an HX-30 (about 2 watts a.m. or 6 watts s.s.b.) without flat-topping. Output is several times what it was with the other arrangement, and linearity is extremely good. Every circuit tunes uncritically, and it is possible to set up almost on-the-button merely by peaking the circuits to approximate frequencies with a grid-dip meter.

The various operating voltages are brought to a terminal strip visible in the upper center portion of Fig. 6-31. In firing up the unit apply plate power to one stage at a time, beginning with the oscillator. This stage works simply, showing the usual sudden downward kick in plate current from about 12 to 5 ma. when the crystal starts oscillating. Set C_1 so that oscillation starts every time voltage is applied.

If you have a grid-dip meter you can set all following circuits close enough without applying power to the unit. The dip meter can also be used to indicate power output relatively from the various stages, and to determine that output is on the desired frequencies.

The pentode amplifier should be checked for stability by removing power from the preceding stages briefly and watching the doubler plate current while tuning C_3. Should any fluctuation appear, adjust C_4 to stop it.

We are now ready to "mix" and to obtain output on 144 Mc. Feed 50-Mc. power into J_6. With power on V_1 and V_2, check for output on 144 Mc. at L_7. A pilot lamp connected to a loop of insulated wire wrapped around L_7 may be used temporarily as an output indicator. When output has been obtained, connect a one-ma. meter to J_3 and J_4, and look for amplifier grid current. Leave plate and screen voltage off the 6360 for the moment.

The lead from J_3 can be removed from the negative terminal of the bias battery and connected to the chassis, to make it easier to obtain grid current for purposes of adjustment, if necessary. Peak all adjustments for maximum grid current, making sure that this drive is on the desired frequency. You'll need something larger than a one-ma. meter if everything is working correctly, or you can reconnect the bias battery once you have obtained a reasonable current reading. Operation of the amplifier from here on is exactly like it would be in a conventional transmitter.

When the conversion unit is used for sideband or a.m. the 6360 operates as a Class AB_1 linear amplifier. Thus the drive must be kept below the level at which grid current starts to flow. In driving an amplifier like the 144-Mc. 4CX250 amplifier described in this chapter, it is not necessary to drive the 6360 into grid current for any class of service. On c.w., for example, it is possi-

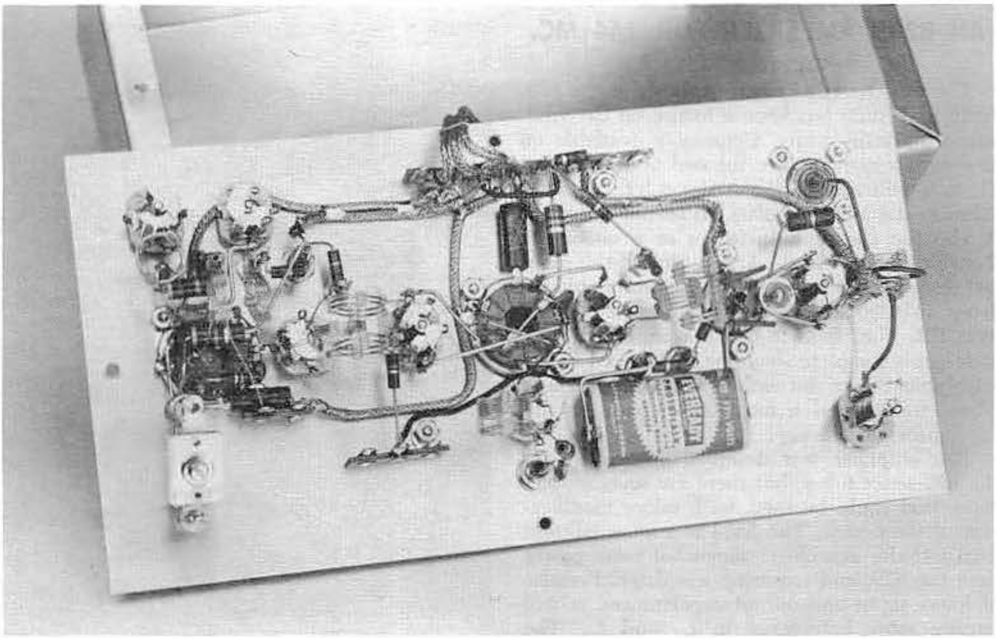


Fig. 6-31—Interior of the 144-Mc. heterodyne exciter. 47- and 94-Mc. circuits are at the left, the mixer in the center, and 144-Mc. amplifier at the right.

ble to develop 600 watts output from the 4CX250s with the 6360 stage running Class AB_1 (no grid current). If a harder-to-drive final stage is used it may be necessary to push the 6360 into Class-C conditions for full-power c.w. work. This will also be necessary if the 6360 is to be plate modulated.

In practice, it is convenient to use the output control on the 50-Mc. exciter as the sole means of controlling the operation of the conversion unit, whether the mode of operation be sideband, c.w. or a.m. Keying for c.w. is done in the 50-Mc. exciter, and modulation of the signal is also done there. We have encountered no linearity problems in the mixer or its following amplifier at any level of operation needed with the 4CX250 push-pull amplifier running at power output levels from 50 to 600 watts.

The conversion unit is plugged into a power supply designed for the 2-band station described in Chapter 7. Power is left on the setup during all operating time, as the current drain without 50-Mc. drive is well below the rated dissipation of all tube elements. Any power supply capable of delivering 250 to 300 volts at 100 ma. and 150 volts, regulated, should be satisfactory. Some typical operating conditions are:

Oscillator plate current: 12 ma. without crystal oscillating; 5 ma. with.
Doubler plate current: 8 ma.
Amplifier plate current: 10 ma.
Mixer plate and screen current: 15 ma. with no 50-Mc. drive; up to 20 ma. with maximum drive.
6360 Amplifier plate and screen current: 25 ma. with no 50-Mc. drive; 48 ma. for operation as linear amplifier; 70 ma. max. for Class-C c.w.
Amplifier grid current: None, except for Class-C operation; about 1.5 ma. max.
Output: 6 watts c.w., sideband or plate-modulated a.m.; 2 watts a.m. linear.

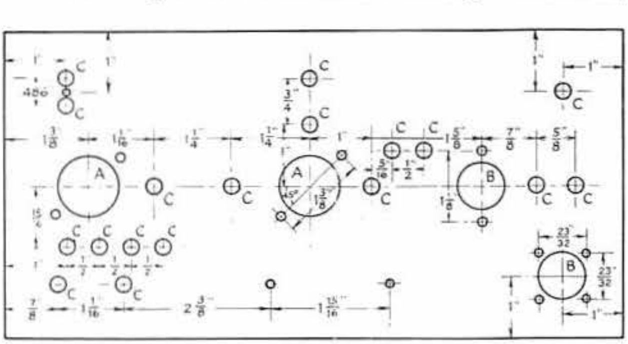


Fig. 6-32—Layout drawing showing principal hole locations and sizes, for those wishing to make a duplicate unit. Hole sizes: A—1 inch, B—¾ inch, C—¼ inch; others ⅛ inch. Chassis and plate are 5 by 10 inches.

AN 829B AMPLIFIER FOR 144 MC.

Fig. 6-33—The 829-B amplifier, with its shield cover in place. Air circulation is provided by the screened holes and cover.

The dual tetrode known variously as the 829, 829B and 3E29 has been a fixture on the v.h.f. scene for many years. Commonly available on the surplus market since the end of World War II, it is still one of the better v.h.f. amplifier tubes in the 100-watt class. At surplus prices, it is also the cheapest. Inclusion of a rather old 2-meter amplifier in the first edition of this manual showed that there is still a considerable interest in this tube, so this modern version by W1CER, Figs. 6-33 to 6-37, is presented here. It features complete shielding, a recessed socket with shield ring, for isolation of the grid and plate circuits, and a metal strap plate circuit, for improved efficiency.

This amplifier was designed specifically for the 829-series tubes, but there are several other types that could be used, with minor modification of the design. The 5894 is a more efficient dual tetrode, capable of somewhat more power than the 829, and requiring less drive. Because of lower input and output capacitances, it will require more inductance in L_2 and L_3. The 832A, a smaller version of the 829 taking lower power and less drive, is also usable.

Construction

The amplifier is built on a 3 by 5 by 10-inch aluminum chassis, with an aluminum cage on top, 9½ inches long, 4 inches wide and 4½ inches high. Holes in the sides and rear of the top compartment, at the tube end, allow for air circulation. The cover is perforated aluminum, permitting the heat to rise from the tube, as cool air moves in from the side holes.

The 829B socket is an E. F. Johnson Type 122-101, designed for recessed mounting. Leads from the socket terminals 1, 4 and 7 to ground are ⅜-inch wide strips of copper or brass, to reduce lead inductance. The .001-μf. capacitors at Pins 3 and 5 are returned to Pin 4, using the shortest possible leads. The grid coil, L_2, is mounted directly on the socket terminals, with the link, L_1, inserted between turns at the center. A 3-lug terminal strip attached to the rear wall supports C_1 and L_1. A 5-terminal barrier strip on the outside rear wall is used for power supply connections.

Coaxial connectors for input and output are on opposite sides of the rear of the chassis. A UG-106/U shield hood covers the back of J_2, to isolate it from J_1 and prevent stray coupling between the input and output. The lead from J_2 to the feed-through terminal and the high-voltage lead from the barrier strip to its terminal up front are made with coaxial cable.

Details of the plate circuit assembly and top enclosure are given in Fig. 6-37. The top edges of the plate line, L_3, are soldered the full length of the stator posts of C_2, for minimum stray inductance at this point. The tuning capacitor is supported on a plastic mounting block, which has narrow slots for L_3. These can be cut in the plastic with a keyhole saw, after drilling starting holes at the top. See detail B. If Teflon of suitable thickness is available, it would be ideal for this support, as it is impervious to heat of the order encountered here. Plexiglas and other clear plastics are usable.

Teflon shafting would also be best for the rod that is to run from C_2 out through the front panel. Wood dowelling is also suitable. Do not use metal stock, as it would be closely-coupled to L_3. The rotor of C_2 must be isolated from ground.

The low-impedance end of L_3 is supported on a 1-inch ceramic pillar. Mount a No. 6 spade bolt at the exact center of the U bend in L_3, thread the standoff onto this, and then bolt the bottom of the insulator in place. The coupling loop, L_4, is supported on the stator post of C_3 and the feed-through bushing to which the coax to J_2 is connected, on the underside of the chassis. C_3 is on the front wall of the shield enclosure, so L_4 is soldered to it after the cover is in place.

The plate line was made of sheet brass, and then silver plated. Flashing copper will work equally well. If not plated, it should be polished thoroughly, and then coated with clear lacquer to reduce tarnishing. The lacquering should be done only after the assembly job is complete. It will be seen from detail C, Fig. 6-37, that there are two strips of thinner stock bolted to the ends

829-B Operating Conditions

Service	E_p (*Max.*)	I_p	E_{sc}	I_{sc}	E_g (*minus*)	I_g
Class C–c.w.	750 v.	160 ma.	200 v.	17 ma.	50 v.	7 ma.
Class C–a.m.	600 v.	150 ma.	200 v.	16 ma.	60 v.	7 ma.
Class AB_1–s.s.b.	600 v.	110 ma.	200 v.	26 ma.	18 v.	0
no sig.		40 ma.	(reg.)	4 ma.		

Maximum plate input for a.m. linear: about 40 watts, Class AB1

of the stiff material of L_3. Holes for these bolts should be larger than needed, so that the line, the straps, and the Fahnstock clips for the plate connections can be assembled loosely at first, then tightened in a position such that no strain is placed on the tube plate pins. Be sure that the tube is seated properly in the socket before the final tightening of the line assembly.

Adjustment and Use

The amplifier may be driven in Class-C service with any exciter delivering 3 to 10 watts output. Operating conditions and maximum plate voltages for c.w., a.m. and s.s.b. service are given below. The 829B works well at lower plate voltages, and is often operated at about 400 to 450 volts in v.h.f. applications. The maximum

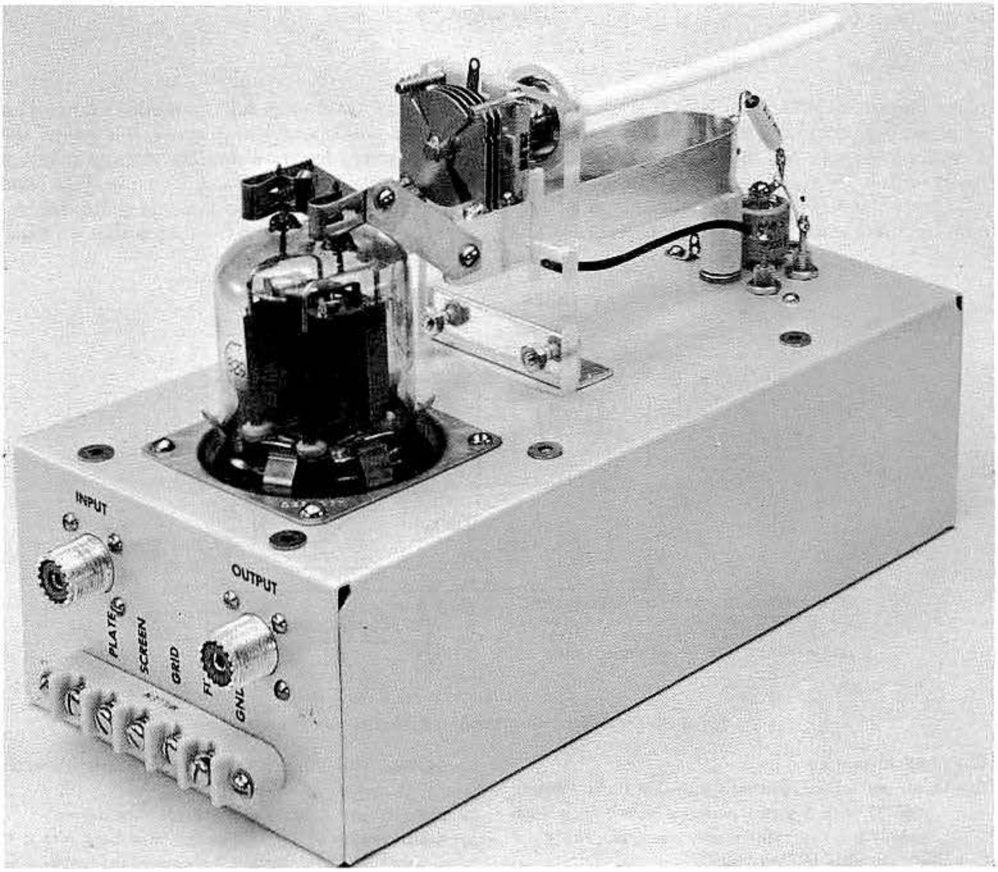


Fig. 6-34—Top of the amplifier chassis, as seen from the rear with the shield cage removed. The output link with its black spaghetti tubing is just below the U-shaped plate tank inductor. The loading control, C_3 is mounted on the shield cage and is not shown here.

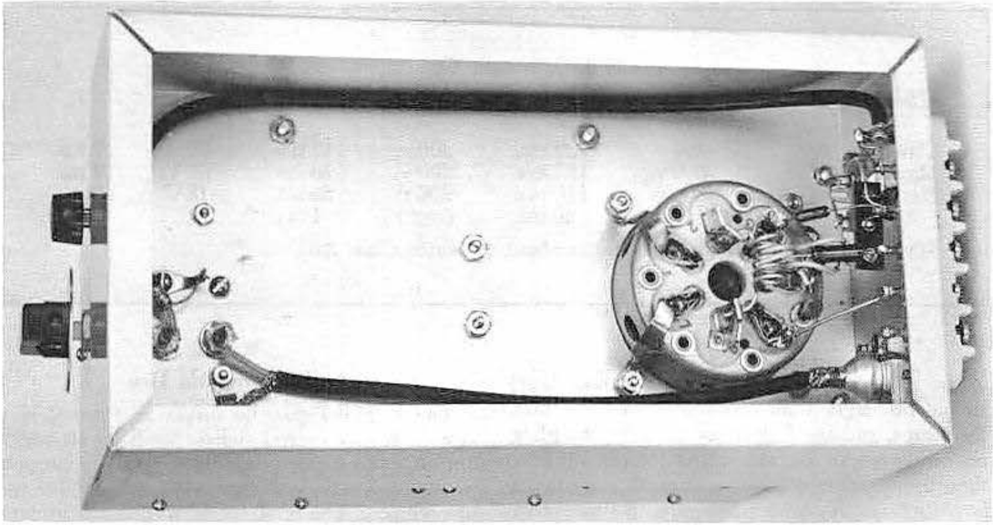


Fig. 6-35—Looking into the bottom of the chassis. The feed through bushings for plate power and r.f. output are at the left. Coax cable is used for the high voltage d.c. lead. Wide copper straps ground the filament and cathode pins of the tube socket. A hood over the back of J_2, lower right, helps isolate the input from the output.

plate current at 450 volts is 200 ma., and this amplifier delivers about 55 watts output this way. A suitable supply for this voltage level can be made with a TV receiver power transformer.

Many amateurs look for a linear amplifier that can be used with the small a.m. transmitters commonly used in v.h.f. communication. This amplifier will operate as a linear, but unless the exciter is very low-powered the step-up may not be attractive. Output in a.m. linear service is no more than half the maximum safe plate dissipation for the tube used. This means that an 829B

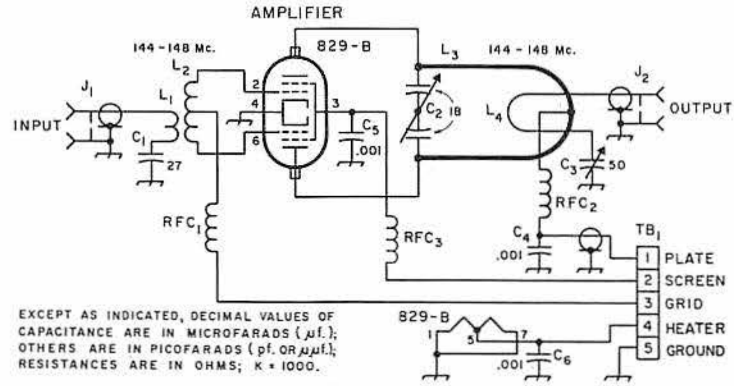


Fig. 6-36—Schematic diagram of the 2-meter amplifier.

C_1—27-pf. silver mica.

C_2—18 pf. per section, butterfly variable (E. F. Johnson 167-22 with 3 stator plates removed from each side. Also, two rotor plates are removed).

C_3—50-pf. variable (Millen 20050).

C_4—0.001-μf. transmitting ceramic (Centralab 858S).

C_5, C_6—0.001-μf. 1000-volt disk.

J_1, J_2—SO-239 connector.

L_1—2 turns No. 22 insulated hookup wire in center of L_2.

L_2—5 turns No. 20 tinned wire, 5/16-inch diameter ½ inch long (see text).

L_3—Plate inductor. See Fig. 6-37 for dimensions.

L_4—6-inch length of No. 12 enam. wire bent into a U with 1¼-inch spacing between sides (cover with spaghetti tubing).

RFC_1, RFC_2—2.7-μh. choke (Millen 34300-2.7).

RFC_3—0.8-μh r.f. choke (Millen 34300-.82).

TB_1—5-terminal barrier strip (Millen 37305).

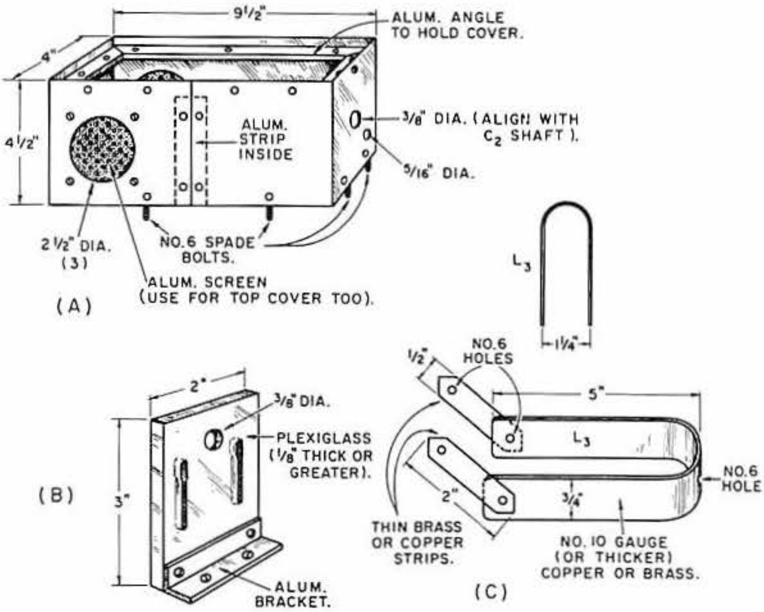


Fig. 6-37—A—General layout of the shield box is shown at A. The box is made from No. 16 gauge aluminum stock. B—Details of the mounting block which supports C_2 and L_3. C—Dimensions for L_3 and its connecting strips (see text).

linear is limited to about 15 watts output on a.m., which may be good enough for use with a 1-watt transistor rig, but not very attractive at higher levels of exciter power. For more on linear amplifiers, their uses and limitations, see the preceding chapter, and "Tips on Linears" later in this one.

Screen voltage should be regulated, in linear service, either a.m. or s.s.b. For c.w., f.m. or high-level a.m., the screen can be supplied through a dropping resistor from the plate voltage source. The value will depend on many factors, but should be about 10,000 ohms at low plate voltages, rising to 35,000 at the high end of the range. Grid bias may also vary, and it may be obtained from a bias supply, or from a grid resistor (connected between RFC_1 and Terminal 3 of the barrier strip) or both. In s.s.b. or a.m. linear service, it preferably should be regulated and adjustable.

Any tetrode amplifier can be run under widely varying conditions, so it can be adjusted to give optimum results with the power supplies you may have available, for modes of emission you are most interested in. The "typical operating conditions" listed in tube tables are guidelines, not laws. But when the tables say "Maximum Ratings," they *mean* it!

To adjust the 829B amplifier, apply heater voltage, and then connect the exciter to J_1. Connect a milliammeter between Terminals 3 and 5, and turn on the exciter, noting the grid current. Adjust the position of L_1 with respect to L_2, and the turn spacing of L_2, for maximum grid current. Now tune the plate circuit slowly through its range, watching the grid current. There may be a slight rise at resonance, but no downward dip. The latter would indicate need for neutralization, which was not required in this version. Grid current should run 7 to 12 ma. for Class-C service. It may be more in the static condition, as it will drop some when the amplifier is actually running, and loaded.

If neutralization is needed, run wires from the grid terminals of the socket up to the top of the chassis on feed-through bushings, and then bring wires up alongside the tube envelope adjacent to each plate. The wires are crossed over under the chassis, and the desired feedback is obtained by varying the position of the top wires with respect to the tube plates.

A lamp load may be connected across J_2 for a rough indication of power output, though a good dummy load and a power-indicating watt-meter or s.w.r. bridge is much to be preferred. Apply plate and screen power, tune C_2 and C_3 for maximum indication, and then adjust the position of L_1 with respect to L_3 carefully, retuning each time the loop is moved. Coupling should be the loosest that will give satisfactory power transfer. The lamp load will be of no value in the adjustment, as it represents a load of far different impedance than will be used ultimately with the transmitter.

KILOWATT AMPLIFIERS FOR 50 AND 144 MC.

The amplifiers shown in Fig. 6-38 were designed for versatility. Though capable of running at the maximum legal power for amateur stations, they operate efficiently at much lower levels. They work well as linears, for use with a.m. or s.s.b., or they can be modulated or keyed in high-efficiency Class-C service. Though the tube type shown is expensive when purchased new, an effective substitute is commonly available on the surplus market at much lower cost. Operated as a rack-mounted pair, as pictured, the amplifiers offer convenient band-changing from 50 to 144 Mc., merely by snapping on the appropriate heater voltage switch, and changing the air connection from one to the other.

The external-anode type of transmitting tube has many variations. The family originated with the 4X150A many years ago, and tubes of the early type are still available, and widely used. A later version, with improved cooling, is the 4X250B, capable of higher power but otherwise very similar to the 4X150A. More recently the insulation was changed from glass to ceramic, and the prefix became 4CX. All the general types thus far mentioned were made with variations in basing and heater voltage that will be apparent to any reader of tube catalogs. The 4CX250R used here is a special rugged version, otherwise very similar to the 4CX250B, and interchangeable with it for amateur purposes. Similar types are supplied by other makers as the 7034/4X150A 7203/4CX250B and 7580. There is another version for linear-amplifier service only, called the 4CX350A.

If one then goes to other basing arrangements similar power capabilities may be found in the 4CX300A, 8122 and others, but differences in tube capacitance might require modification of the circuit elements described here. The air-system sockets (required for all external-anode tubes mentioned) may be the same for all types in the second paragraph, but those just above require different sockets.

Both amplifiers take a kilowatt on c.w. or s.s.b. with ease. The 144-Mc. model must be held to 600 watts input for plate-modulated service to stay within the manufacturer's ratings. On 50 Mc. the three tubes in parallel loaf along at 1000 watts in the low-duty-cycle modes. The permissible input on a.m. phone is 900 watts. Class C efficiency is on the order of 75 per cent, over a wide range of plate voltages. It is possible to run all the way from 800 to 2000 volts on the amplifier plates without altering screen voltage or drive levels appreciably.

Mechanical Layout

The amplifiers are similar packages, to mount together harmoniously, though this is of only incidental interest to the fellow concerned with one band or the other. They are built in standard 4 by 10 by 17-inch aluminum chassis, mounted open side up and fitted with shield covers. In the author's station a single blower is used for all transmitters. This explains the air-intake sleeve seen on the back of each amplifier. An air hose from the remote blower is pushed into the amplifier being used.

The transmitters are all hooked up together, to meters, power circuits, audio equipment and

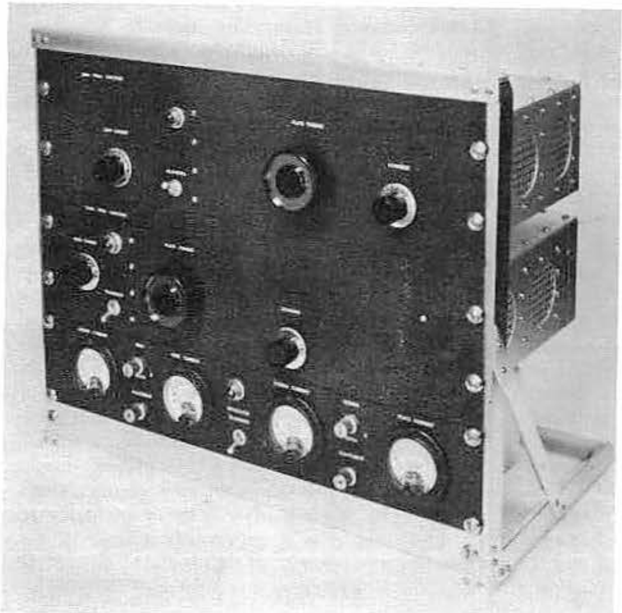


Fig. 6-38—The kilowatt amplifiers for 50 and 144 Mc. in a rack made from aluminum angle stock. At the bottom is a meter panel with controls for meter and mode switching.

power supplies common to all. Changing bands involves mainly the switching on of the desired heater circuits, and the insertion of the air hose in the proper intake sleeve. Separate antenna relays are provided for each final stage, and power switching and plugging and unplugging are largely eliminated.

Tube sockets are the air-system type, mounted on 4-inch high partitions with folded-over edges that are drawn up tightly to the top, bottom, front and back of the chassis with self-tapping screws. Air is fed into the grid compartments at the left side, as viewed from the front. Its only path is through the sockets and tube anodes, and out through screened holes in the right side of the chassis. Panels are standard 5¼-inch aluminum. Controls for the amplifiers are similar, though their locations are slightly different. No attempt was made to achieve symmetry through mechanical gadgetry, since the unbalance of the front panels is not unpleasing. The rack shown in Fig. 6-38 was made up from aluminum angle stock to fit the job. Several screen and bias control arrangements were tried before the circuit shown in Fig. 6-43 was settled upon. Meters read driver plate current, and amplifier grid, screen and plate currents. Switches enable the operator to check the grid and screen currents to each tube in the 144-Mc. amplifier separately, and the screen currents in the 50-Mc. amplifier likewise. A mode switch provides proper screen operating conditions for a.m., linear or c.w. service.

The 50-Mc. Amplifier

The use of three tubes in parallel in the 50-Mc. amplifier was an experiment, tried with the expectation that parasitics, unbalance, excessive tank circuit heating and all manner of troubles would develop. These problems never materialized; use of paralleled tubes seemed to introduce no problems on its own, and extensive

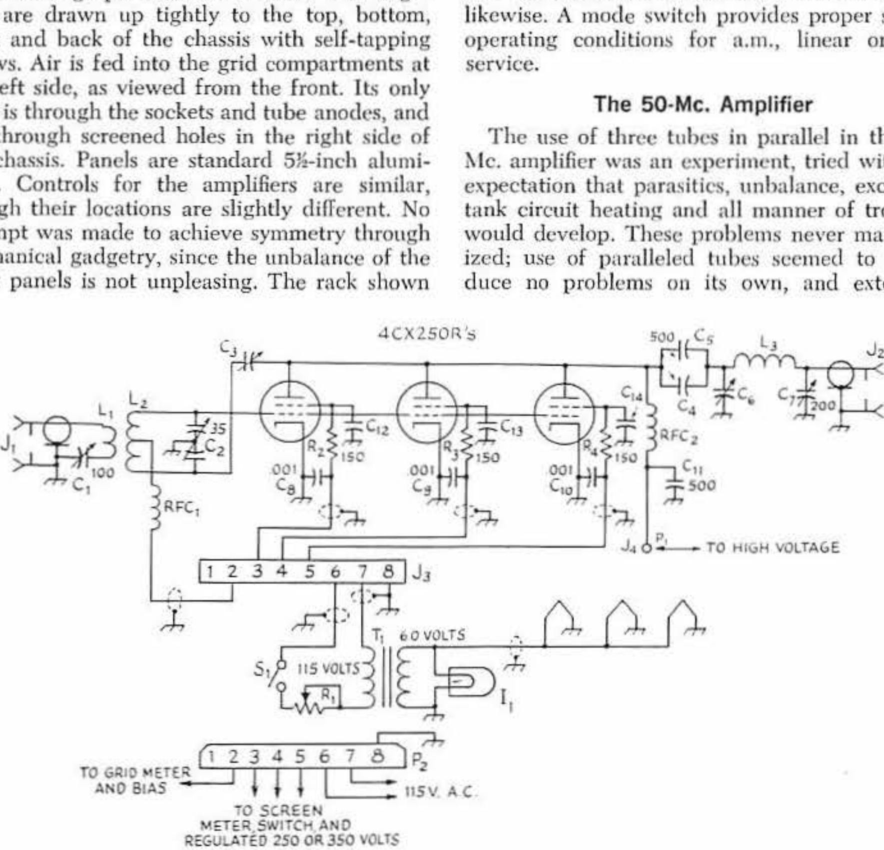


Fig. 6-39—Schematic diagram and parts information for the 50-Mc. amplifier.

C_1—100-pf. miniature trimmer (Hammarlund MAPC-100).

C_2—35-pf. per section split-stator (Hammarlund HFD-35X).

C_3—Neutralizing capacitance—see text.

C_4, C_5, C_{11}—500-pf. 5000-volt transmitting capacitor (Centralab 858S-500).

C_6—Tuning capacitor made from 3-inch aluminum disks —see text and Fig. 6-40.

C_7—200-pf. variable, .03-inch spacing (Johnson 167-12 or 200L15).

C_8, C_9, C_{10}—.001-μf. disk ceramic.

C_{12}, C_{13}, C_{14}—Bypass built into special air-system socket.

I_1—Green-jewel pilot lamp holder.

J_1, J_2—Coaxial chassis receptacle.

J_3—8-pin male power fitting.

J_4—H.v. power connector female (half of Millen 37501).

L_1—1 turn insulated wire about 1-inch diam. Make from inner conductor of coax running to J_1. Strip jacket and braid back about 4 inches. Insert between center turns of L_2.

L_2—8 turns No. 14, ⅝-inch diam., 1¼-inches long, center-tapped.

L_3—3 turns 2 inches diam., 3 inches long, ¼-inch copper tubing.

P_1—High-voltage power connector, male (half of Millen 37501).

P_2—8-pin cable connector to match J_3, female.

R_1—20-ohm 10-watt slider-type resistor. Set so that heater voltage is 6.0 at socket.

R_2, R_3, R_4—150-ohm ½-watt resistor. Connect at socket screen terminal.

RFC_1—No. 32 enamel wire, close-wound full length of 1-watt resistor, 10,000 ohms or higher.

RFC_2—No. 28 d.s.c. or enamel-wound 1¾ inch on ½-inch Teflon rod. Space turns 1 wire diam. for 8.3 μh. For winding information see Chapter 13.

S_1—S.p.s.t. toggle.

T_1—6.3-volt 8-amp. Adjust R_1 to give 6.0 volts.

experience with the amplifier has confirmed the worth of the idea. This happy state of affairs involves a few basic considerations that should be stated here.

1) Paralleling straps in the grid and plate circuits were made "three of a kind." The two going to the outer grids were bent identically, and then the one for the middle tube was bent back on itself as necessary to use the same total length of strap. The same was done in the plate circuit.

2) The grid circuit was split-stator tuned, to get a reasonably-sized grid coil, even with the combined input capacitance of the three tubes plus circuit capacitance—some 60 pf. or more. This also provided a means for easy neutralization.

3) The pi-network plate circuit is tuned with a handmade disk capacitor. This has a far lower minimum *C* than the more conventional tuning capacitor, and it is devoid of the side bars and multiple ground paths that are so often the cause of parasitics in v.h.f. amplifiers. No parasitic resonances were found in this amplifier, other than one around 100 Mc. introduced apparently by the r.f. choke. This caused a blowup when grid-plate feedback developed with a similar choke in the grid circuit. The problem was solved easily by use of a low-*Q* choke of different inductance in the grid circuit. Do *not* use a high-quality r.f. choke for RFC_1!

4) All power leads except the high-voltage one are in the grid compartment, and made with shielded wire. Where the high voltage comes into the plate compartment it is bypassed at the feed-through fitting.

5) The plate circuit is made entirely of copper strap and tubing, for highest possible *Q* and low resistance losses. It may be of interest that the entire tank circuit was silver-plated after the photographs were made. Efficiency measurements made carefully before and after plating showed identical results.

Looking at the interior view, Fig. 6-40, we see the grid compartment at the left. The coaxial input fitting, J_1 in Fig. 6-39, is in the upper left corner of the picture. Coax runs from this, out of sight on the left wall, terminating in a loop, L_1, made from its inner conductor. This is inserted between turns at the center of the grid coil, L_2. The series capacitor, C_1, is just visible on the left chassis wall. It is not particularly critical in adjustment, so no inconvenience results from its location away from the front panel.

Screen voltage, bias, and 115 volts a.c. come through an 8-pin fitting, J_3, mounted betweeen the air intake and the heater transformer, T_1. On the front panel are the heater switch, S_1, and the pilot-lamp holder.

The three air-system sockets (Eimac SK-600, SK-620, SK-630, Johnson 124-110-1 or 124-115-1, with chimneys) are centered on the partition, spaced so that there is about ¼ inch between their flanges. The small angle brackets that come with the sockets should be tightened down with their inner ends bearing against the ceramic chimneys, to hold them in place. Note that the 150-ohm isolating resistors R_2, R_3, and R_4 are connected right at the screen terminals.

Both grid and plate straps are cut from flashing copper ⅝-inch wide. Lengths are not critical, except that all grid straps should be the same length, and all plate straps identical. The plate straps are made in two pieces soldered together in T shape, to wrap around the anode and join at the coupling capacitors, C_4 and C_5. These T-shaped connections could be cut from a sheet of copper in one piece, with a little planning.

The copper-tubing plate coil, L_3, is mounted on stand-off insulators not visible in the picture. Connections to the coupling capacitors, the tuning capacitor, C_6, and the loading capacitor, C_7, are made with copper strap. It will be seen that these various pieces are bolted together, but the were also soldered. The connection from C_7 to the output fitting, J_2, is a single strap of copper, bolted and soldered to L_3.

The disk tuning capacitor can be made in several ways. Flashing copper is easy to work, and the 144-Mc. capacitor was made of this material. A more sturdy disk can be made from ⅛-inch aluminum. Those shown in Fig. 6-40 were 3-inch meter cutouts from an aluminum panel. Disk-type neutralizing capacitors (if you can find them; they're not common catalog items these days) provide ready-made disks and lead screws for tuning. For the latter we used 3-inch ¼-20 brass screws from a neighborhood hardware store. A panel bushing with brass nuts soldered to it provided the lead-screw sleeve. The stationary disk is supported on ½-inch-diameter Teflon rod, a material also used for the r.f. choke form. Teflon works easily and is unexcelled for insulating applications where high temperatures are encountered. We found it reasonably priced, in various diameters, at a local plastics house.

The plate r.f. choke, RFC_2, is important. You'll probably have to make it to get one of sufficiently good quality. For more on this see information under Fig. 6-39 and "R. F. Chokes for the V.H.F. Bands," Chapter 13. Two coupling capacitors were paralleled because we've experienced trouble with exploding capacitors in pi-network plate circuits in the past. Maybe one would have handled the job, but two do for sure.

Some Possible Variations

It is always risky to suggest variations on a design unless they have been checked out in use, as bugs may develop in unforeseen ways. The following are ideas only, to be used at the builder's risk, since they have not been tested by the designer.

You might not care for three tubes in parallel. Two should work well, handling a kilowatt except in a.m. linear or plate-modulated service. Many builders report success with 2 tubes.

For those who can afford it, a vacuum variable

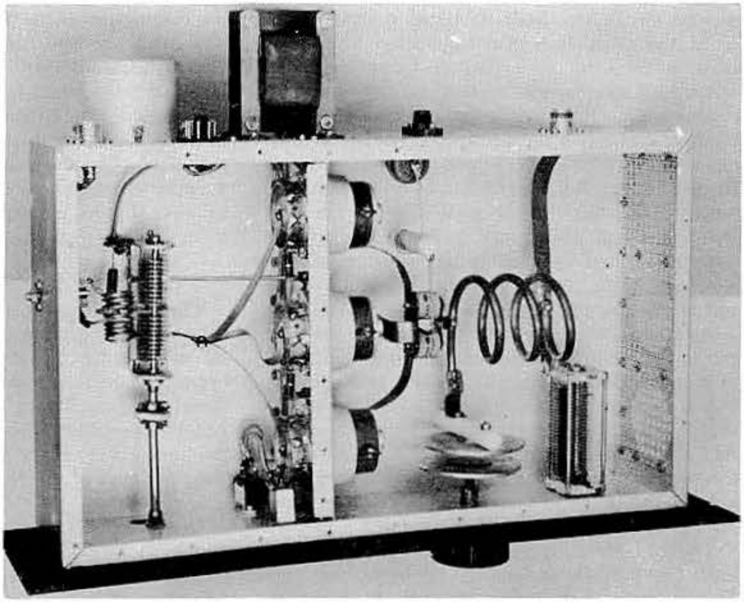


Fig. 6-40—Interior of the 50-Mc. amplifier. Note method of paralleling grid and plate connections. Cylinder at upper left is for detachable air hose.

capacitor should be ideal for C_6. One with about 10 pf. maximum capacitance should do nicely.

For lower tube cost, 4X150As from surplus should work without mechanical changes. Use plenty of air, if you intend to push the ratings of the 150As. A 100-c.f.m. blower is not too much. The ability of the anode structure to withstand heat is the main difference between the 150A and later versions of this tube, and some people have gotten away with 250 ratings with 150-type tubes. In this connection, the 50-Mc. amplifier will take a kilowatt at 1200 to 1500 volts, if your power supply will handle the current. This approach, plus plenty of air, is preferable to using plate voltages much in excess of the 4X150A ratings.

The 144-Mc. Plumber's Special

Use of 1⅝-inch copper tubing for a 2-meter tank circuit is by no means new.* We simply went one step further and made the entire circuit from standard plumbing components. All the heavy metal you see in the plate compartment of Fig. 6-41 came from the plumbing counter of the local Sears store. The picture and Fig. 6-41 should be largely self-explanatory.

At the tube end of the plate line, L_4 in Fig. 6-42, we have brass castings normally used to join sections of the copper pipe. They make a nice sliding fit over the tube anodes. For tighter fit, cut thin brass shim stock and insert as much as needed between the anode and the sleeve. The end of the fitting can be slotted and then clamped firm on the anode with a hose clamp, as an alternative. The short at the B-plus end of the line is made with two T fittings, with their flanges cut down to ½ inch and slipped over a short section of the pipe that is not visible. Joints throughout the assembly were silver-soldered with a torch, but conventional soldering should do equally well. The flanges at the open ends of the T fittings are cut down to about ¼-inch in length.

The last instruction and the information about the plate line given under Fig. 6-42 apply only if the fittings are identical to those obtained by the builder. Since there are several types of fittings available from plumbing supply houses, the following overall dimensions should be heeded: tube end of the plate line to center-line of short—10⅜ inches; spacing of pipes center to center—3½ inches.

In using tube types other than those specified, it may be that some change in plate circuit inductance will be needed. A simple check will show if this is needed. Slip the castings and pipe together without soldering, and assemble the plate circuit temporarily. Check the tuning range by means of a grid-dip meter. No plate or heater voltage is needed for this rough check, but it is well to have the coupling loop in place, and a 50-ohm resistor connected across J_2.

The coupling loop, L_5, is cut from a single piece of flashing copper ½ inch wide. This delivered slightly more output to the load than was obtained with loops of wire of various lengths tried. The loop should be positioned so that the bottom edge is approximately flush with the bottom of the pipes. Optimum coupling to a 50-ohm load is achieved when the closed end of the "U" is about ¼ inch lower than the open end. Looking down at the plate-line assembly, the coupling loop is centered between the pipes.

The loop and plate line are supported on

*"High-Efficiency 2-Meter Kilowatt," *QST*, Feb. 1960, p. 30. "Top Efficiency at 144 Mc. with 4X250Bs," Breyfogle, *QST*, Dec. 1961, p. 44.

Teflon rod insulators. The r.f. choke is also wound on Teflon. Note its position *outside* the U of the plate line. First mounted inside the loop, it went up in a furious burst of smoke when high power was applied to the amplifier.

Our tuning disks are 3-inch sheets of flashing copper. For nicer appearance and better mechanical stability, use ⅛-inch aluminum as in the 50-Mc. model. Three-inch brass ¼-20 screws are threaded through the pipe fittings. The rear one is held in place with a lock nut, and the other is rotated by the tuning knob, a bakelite shaft coupling, and a length of ¼-inch Teflon rod running in a panel bushing.

A third disk is mounted adjacent to the rear portion of the tank circuit. Its position is adjusted to achieve perfect balance in the tank circuit, but in practice this turned out to have no measurable effect. It is felt that a really good choke at RFC_1, and careful adjustment of C_1, can practically eliminate the effect of any slight unbalance if the point of connection of RFC_1 to the tank circuit is not bypassed to ground.

The 144-Mc. grid circuit, L_1L_2, looks like two coils, but actually is a coiled-up half-wave line. This is somewhat more compact than a half-wave line with its conductors out straight, and it seems equally effective. The grids are connected to the outer ends and the tuning capacitor to the inner. The point of connection of the bias-feed resistors should be determined in the same way as with the usual half-wave line: by coupling in 144-Mc. energy and touching a pencil lead along the inductance while watching the grid current. The correct point for final connection of the resistors is that at which no reaction on grid current is observed. Isolating resistors here, and for feeding screen voltage to the sockets, are preferable to r.f chokes. The inner conductor of the coaxial line is used to make the coupling loop, L_3, which is placed between the inner ends of the grid circuit.

Balanced drive is maintained by adjustment of the differential capacitor, C_1, connected in parallel with C_2, and mounted on the side of the chassis adjacent to it. The series capacitor, C_3, is out of sight under the tuning capacitor, which is mounted on standoff insulators. It is adjusted by inserting a small screwdriver in a hole in the side of the chassis, but if we were doing it again we'd mount C_3 on the side wall, just under C_1, to make it more readily adjustable. Note that the rotor of C_2 is ungrounded.

About Neutralization

These amplifiers were tested without neutralization and we almost got away with it, but use of all modes, particularly a.m. linear and s.s.b., imposes strict requirements on stability. Conventional cross-over neutralization employed in the 144-Mc. amplifier is omitted from Fig. 6-42 in the interests of clarity. The schematic representation, C_3 in Fig. 6-39, is not very informative either.

In the 50-Mc. amplifier the lead visible in Fig. 6-40, attached to the rear stator terminal of C_2, runs to a polystyrene feedthrough bushing (National TPB) mounted in the partition between the rear and middle sockets. Even this bushing's wire stub projecting into the plate compartment turned out to be too much "C_3" and it was trimmed off 1/16th inch at a time, until minimum feedthrough was indicated on a wavemeter coupled to L_3 and tuned to the driving frequency.

Similar feedthrough bushings are used in the 144-Mc. amplifier, but here a small wire had to be added to each one. The wire connected to the grid of the front tube is aimed toward the anode of the rear tube, and vice versa. Small sheets of thin brass or copper should be fastened under the adjacent edges of the sockets, and bent up at right angles to the partition. These ⅜-inch high barriers act to shield the

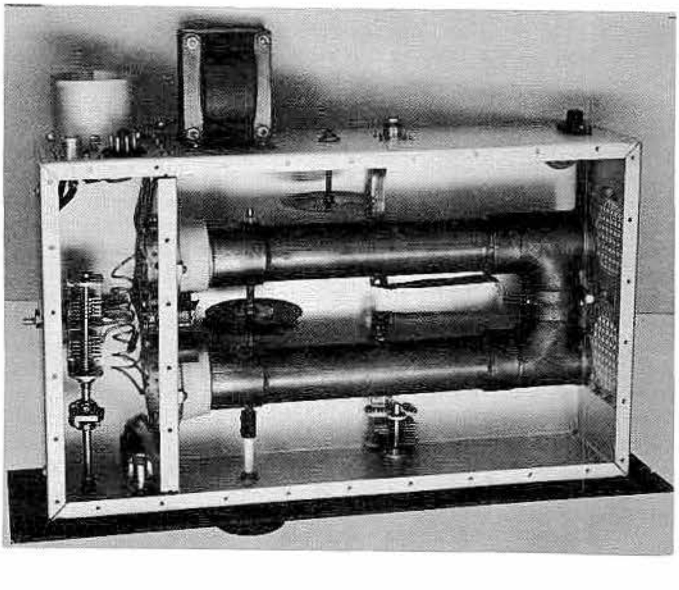


Fig. 6-41—Interior of the 144-Mc. amplifier, showing the plate circuit made from standard plumbing components. Brass pipe junctions make connection to the anodes, and T fittings are modified to form the short at the end of the line.

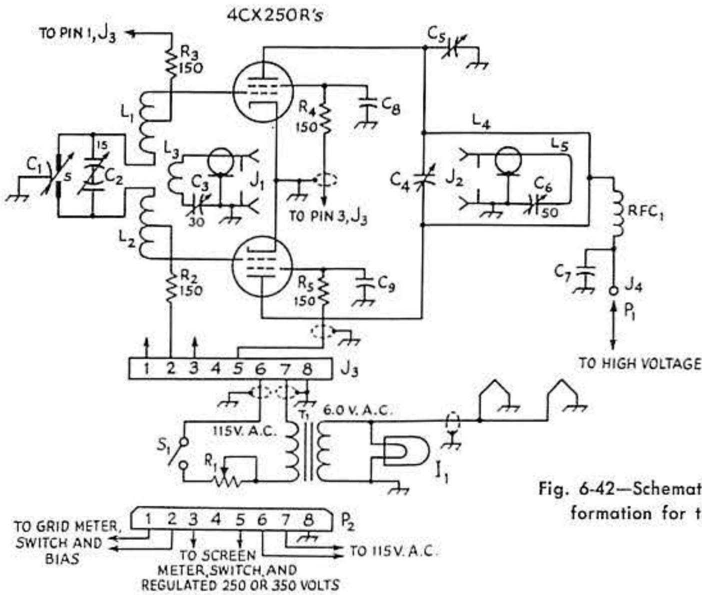


Fig. 6-42—Schematic diagram and parts information for the 144-Mc. amplifier.

C_1—5-pf. differential trimmer (Johnson 160-303 or 6MA11).

C_2—15-pf. per section split-stator (Hammarlund HFD-15X). Leave rotor ungrounded.

C_3—30-pf. miniature trimmer (Hammarlund MAC-30).

C_4—Tuning capacitor made with 3-inch disks. See text and Fig. 4.

C_5—3-inch disk movable with respect to L_4. See text and Fig. 6-41.

C_6—50-pf. variable (Hammarlund MC-50).

C_7—500-pf. 5000-volt (Centralab 858S-500).

C_8, C_9—Bypass capacitor built into air-system socket.

I_1—Green-jewel pilot lamp holder.

J_1, J_2—Coaxial chassis receptacle.

J_3—8-pin male chassis connector.

J_4—High-voltage power connector, female (half of Millen 37501).

L_1, L_2—3½ turns No. 14, ⅝-inch diam., turns spaced ½-inch. R_2 and R_3 tap on about 1 turn in from grid end. See text.

L_3—1-turn inner conductor of coax from J_1, about ¾ inch diam. Remove jacket and braid about 3 inches. Adjust position with respect to L_1, L_2 for maximum grid current.

L_4—Plate line 1⅝-inch copper pipe, with junctions and T fittings. Exposed portion of pipe is 8 inches long. Cut right end of T fittings to ¼-inch shoulder, and joined ends to ⅜-inch shoulders.

L_5—½-inch strap of flashing copper, U portion 4 inches long and 1¼ inch wide. Make loop and connections from single piece. Support L_4 and L_5 on standoffs of ceramic or Teflon.

P_1—High-voltage connector, male (half of Millen 37501).

P_2—8-pin female cable connector to match J_3.

R_1—20-ohm 10-watt slider-type. Adjust for 6.0 volts at socket.

R_2, R_3, R_4, R_5—150-ohm ½-watt resistor.

S_1—S.p.s.t. toggle.

T_1—6.3 volt 8-amp. Adjust R_1 for 6.0 volts.

RFC_1—2.15 μh. r.f. choke. No. 22 enamel closewound $1\frac{3}{16}$ inch on ¼-inch Teflon rod.

screen rings of the tubes from the feedback "capacitors" and assure that the coupling is from grid to opposite plate, and not to the screen.* Length and position of the feedback wires are adjusted for minimum feedthrough of driver energy to the plate circuit, as described above. About a half inch of wire was needed in addition to the terminal stub in this case.

When used as linear amplifiers the tubes must be biased to permit them to draw considerable plate current with no drive, so perfect neutralization is a "must." Properly neutralized, the amplifiers will be stable when run at or near maximum safe plate dissipation with no drive, even when the grid and plate circuits are swung through their entire ranges. If they will not pass this test the amplifiers are not ready to be used for linear service.

Controls and Metering

Almost everyone who builds his own equipment has a favored way of controlling it, so the system shown schematically in Fig. 6-43 may not suit everyone. It is for use in a station where power supplies are actuated by closing the primary circuits to all that the operator wants to have come on for transmitting purposes. They are mounted away from the transmitting position, and a cable carries the various voltages to the r.f. position. At the left, J_1, J_3, J_4 and J_5 are terminals carrying all voltages from the power-supply position. These are distributed through meters, controls and output fittings, J_6, J_7 and

*Air-system sockets are now available with built-in shielding of the screen ring. The Eimac numbers are SK-620 and 630.

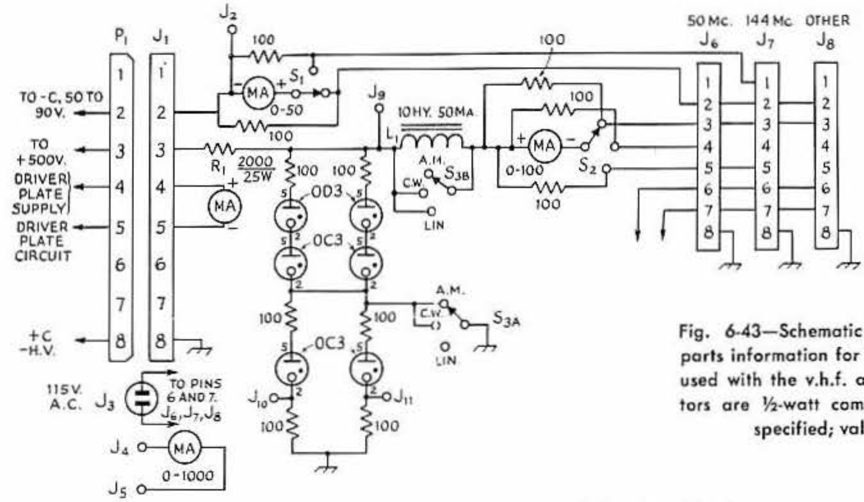


Fig. 6-43—Schematic diagram and parts information for the control unit used with the v.h.f. amplifiers. Resistors are ½-watt composition, unless specified; values in ohms.

J_1—8-pin male power connector.

J_2, J_9, J_{10}, J_{11}—Tip jack.

J_3—A.c. connector, male.

J_4, J_5—High-voltage feedthrough connector (Millen 37501).

J_6, J_7, J_8—8-pin female power connector.

L_1—10 hy. 50-ma. choke. Must be shorted out for other than plate-modulated service.

P_1—8-pin female cable plug.

R_1—2000-ohm 25-watt resistor. Value may be reduced to as low as 1000 ohms if regulation at high values of screen current is desired, provided current measured in J_{10} and J_{11} does not exceed 40 ma. under low-screen-current conditions.

S_1—Single pole 2-position switch.

S_2—Single-pole 3-position switch.

S_3—Double-pole 3-position switch.

J_8, to various transmitters. Circuit breakers at the supply position are used to turn everything off when the station is closed down.

Adjustable bias, 50 to 90 volts negative, is brought in through Pin 2 to a 50-ma. meter and appropriate shunts that keep the circuit that is not being metered closed. The switch S_1 enables the operator to read the grid currents separately in the 144-Mc. amplifier. Grid voltage may be read when required, at J_2.

Similarly, a 500-volt positive source is connected through Pin 3, a voltage-regulating system, an audio choke, a 100-ma. meter and a 3-position switch, S_2, to the screens. Currents can be read separately here, too, and this facility is important in determining that all tubes are running within ratings. The VR system is switched by S_{3A} to provide regulated 250 or 350 volts to the screens. Ganged to it is S_{3B}, which shorts the audio choke for all modes except plate-modulated a.m. This must be done, as the choke will cause trouble on the other modes. The series-parallel VR-tube bank is by no means an ideal regulating system, but it prevents soaring of the screen voltage under conditions of low or negative screen current. These occur only in linear operation, and on c.w. when the key is up. It is not particularly important that screen voltage be held constant for high screen current, as in plate-modulated a.m. and key-down c.w. conditions with low plate voltage. The screen voltage will be kept down by the heavy load on the supply at such times. Actually a single string of three regulator tubes will do the job quite well, and both amplifiers have been worked successfully with this simpler screen arrangement. Current through the regulator tube strings can be measured between J_{10} or J_{11} and ground.

Operation

Because a variety of tubes may be used, with a wide range of conditions as to plate voltage and drive, we're not going to be too specific here. If you follow the tube manufacturer's recommendations for the plate voltage you intend to use you won't be far wrong. All tubes of this class are quite versatile as to drive level and plate voltage; unless you are running close the maximum plate-input ratings the principal factor to watch is screen dissipation, as far as safety of the tubes is concerned. Set up your amplifier with a dummy load and then try the various conditions given in tube data sheets, observing the operation on all meters. In this way you'll soon learn your way around. A few words of preliminary advice may, however, be in order.

First, don't feel that you have to run a kilowatt right off the bat. Put a Variac in your final plate supply primary and run the voltage down for initial testing, or use a lower-voltage supply until you become familiar with the way the rig works. Watch the screen current closely, particularly at low plate voltage or with high grid drive or light loading. The provision for checking individual screen currents is important, otherwise you may learn too late that one tube has been taking all or most of what you have seen on a meter that reads total screen current only. In the push-pull amplifier it may be advantageous to balance screen currents by C_1, rather than grid currents, if balance of both screen and grid curents does not occur at one setting.

Tune up for Class C and get the feel of the amplifiers before trying linear operation. Then, if linears are unfamiliar to you, read up on them elsewhere in this chapter before jumping in. Use a scope; there is no sure way to set up and operate a linear without one. The Heath Monitor Scope, HO-10 or SB610, is ideal for this job because of its built-in tone oscillator and in-the-transmission-line features. Running a linear, either sideband or a.m., without a scope check is inviting trouble.

Finally, if you must use an a.m. linear, don't expect 70 per cent efficiency from it. Don't expect 50. Expect and see that you *get*, no more than 35 per cent from a Class AB_1 linear, or no more than about half the rated plate dissipation for the tubes used. This means 350 watts out of our 50-Mc. amplifier with a kilowatt in, even though you can get 750 watts out of it in Class C. For the 144-Mc. amplifier, 200 watts out with 700 in is about the safe maximum for a.m. linear service. These are optimum figures; you may get less, but you can't get more and be *linear*.

For higher plate efficiencies go to s.s.b., c.w. or plate-modulated a.m. In any of these modes these amplifiers will give you the biggest legal signal around, if that's what you want. Or they'll throttle down nicely to 300 watts input or less, merely by lowering the plate voltage. They'll work efficiently at much lower inputs if the screen voltage is dropped appropriately. Chances are that you'll still have a signal that will stand out in most neighborhoods, on either 6 or 2, and you'll have no worries about over heating.

TIPS ON LINEAR AMPLIFIERS

It is no small wonder that the linear amplifier appears attractive to the neophyte looking for his first step up the v.h.f. power ladder. At first glance it seems almost too good to be true. A Class AB_1 linear, the type most often used, requires *no driving power at all.* Class AB_1 is operation without the amplifier drawing grid current at any time. With the amplifier consuming no power from the driver stage, only a mere handful of exciter is needed. You could use a one-watt transistor rig, and have output to spare.

This applies whether the amplifier runs 100 watts input or 1000, so it can be seen that the linear is most attractive in the high-power bracket. The inevitable price to be paid is low efficiency. Thus there is hardly any point in building a linear for less than about 200 to 300 watts input; you won't get enough step-up in power to make the project worthwhile. And since any amplifier is a fairly expensive undertaking, it may be well to build it for kilowatt capability, even if you don't expect to push it that far right away. The amplifiers of Fig. 6-38 through 6-43 can be run as low as about 300 watts input if you wish. At this level they deliver about 100 watts to the antenna—no mean signal on a v.h.f. band. There is plenty in reserve when you need it, and the final tubes hardly know they're working.

As its name implies, a linear amplifier is one which reproduces the wave form of its driver stage exactly, but at higher power level. This requires considerable attention to details. Everything has to be *right*, or the signal quality suffers, and it will occupy far more space in the band than a signal should. Grid bias, drive level and antenna loading are all critical. Regular use of an oscilloscope is a must. Meters alone are not enough, if you want to be sure that your signal is above reproach.

About Driver Stages

Obviously the driver stage is important in the linear picture. If we are going to amplify it in exactly its original form, the signal had better be good to start with. A distorted splattering signal fed to a linear results in more of the same; lots more! The exciter should be stable and its output stage as perfectly modulated as we can make it. Since the driver operates at very low level, this is not hard to do. If an exciter is being built especially to drive a linear, it might be well to go with a neutralized-triode output stage, with no more than about 5 watts input. A Class-A modulator employing inverse feedback and some form of output limiting would be good. Peak limiting is important, to keep the average modulation percentage high and prevent overmodulation.

Most v.h.f. transmitters will have a lot more output than is needed, so the drive applied to the amplifier must be reduced in some way. Detuning the driver output circuit or the amplifier grid circuit will not do, as it may leave the driver without a proper load, and impair its modulation quality. A simple solution is to connect a 50-ohm dummy load parallel with the driver output. A coaxial T fitting is connected to the driver output receptacle. The dummy load is connected to one side of the T, and the amplifier grid input to the other. The amplifier grid circuit still may have to be detuned slightly, if the exciter output is more than 2 or 3 watts, but this will not be harmful for only a small reduction in drive. Driver output may also be reduced by lowering its plate or plate-and-screen voltage, though it is well to check the quality to be sure that linear modulation characteristics are being obtained in the driver.

Checking Signal Quality

The Health Monitor Scope, Model HO-10 or SB610, is ideal for use with a v.h.f. linear, as it may be left connected to the transmission line for continuous monitoring. Some modification may be necessary for effective use of this scope on 144 Mc., though it works nicely on 50 Mc. and lower bands as is. Two coaxial receptacles of the SO-239 type are mounted on the back of the scope, with their inner terminals joined by a wire about 1½ inches long. The transmitter is connected to one receptacle and the antenna coax to the other. The unshielded wire inside the scope causes an appreciable impedance bump in a 144-Mc. line. This may be corrected by connecting a coaxial T fitting to one of the terminals, and using its two arms to make the above connections from transmitter to antenna line. Internal scope connections and functions remain intact, and the impedance bump is held to manageable proportions.

The scope, milliammeters in the grid, screen and plate circuits of the amplifier, and a power-indicating device in the coaxial line are useful in setting up the linear for maximum effectiveness. The power meter will tell you if you are getting all you should from the amplifier. If you're getting too much, the scope will tell you. The meters are necessary to assure operation at both safe and optimum conditions.

The tube manufacturers' data sheets give typical operating conditions for various classes of service, usually including a.m. linear. These are the best guides available and you'll do well to follow them closely, especially when just learning your way around with a linear. They do not tell the whole story, however. They are merely "typical"; there may be other combinations that will work well, if you know how to read the indications your meters and scope provide. Conversely, it may be possible to radiate a less-than-admirable signal, when meter indications alone seem to be in order. You'll need that scope!

In using the 6- and 2-meter linears of Fig. 6-38 the plate voltage can be almost anything, provided that the amplifier is adjusted carefully whenever the plate voltage is changed. From 800 to 2000 volts has been used on 4CX250Rs and Bs. Screen voltage should be what the sheet calls for; in this case 250 volts for Class C and 350 volts for Class AB_1. Bias should be variable and adjusted so that the tube or tubes will draw the recommended no-drive plate current. In this instance it's about 100 ma. per tube. It is well to start with bias on the high side (no-drive plate current low) to be on the safe side until set up correctly.

With the amplifier running in this fashion, feed in enough drive to make the plate current rise and output start to appear. Tune the final plate circuit and adjust the loading control for maximum output, as indicated by the height of the scope pattern or by the power-indicating meter in the transmission line. Disregard the final plate current, so long as it is at a safe value (Do not tune for dip; tune for maximum output.) Run up the drive now to the point where grid current just starts to show, and then back it off slightly. Readjust the plate and loading controls for maximum output. Be sure that you're putting every watt you can into the transmission line for this amount of grid drive. Maximum loading is a must for linear operation.

Try modulating the driver, while watching the scope pattern. It should look like the patterns shown in Fig. 6-44, A and B. These are envelope patterns, which are most readily obtained with the Monitor Scope. Unmodulated carrier is shown at A. The Heath scope has a built-in tone oscillator. Using this or a steady whistle into the microphone should produce a pattern like the one at B, when the modulation level is 100 per cent. The peaks and valleys are sharp, and the valleys (negative peaks) just reach the zero line. Positive peaks are just twice the total height of the unmodulated envelope, Pattern C shows effects of excessive grid drive or too-light loading, or both. Note the flat-topping, and the lower height of the positive peaks. If you don't have some form of negative-peak limiting, watch out for excessive modulation in that direction. That's where the splatter comes from first if audio and r.f. operation is clean otherwise. In watching your voice modulation beware of the bright flashes at the zero line of the modulation pattern that indicate over-modulation on negative voice peaks.

Practice the adjustment routine with a dummy load connected to the transmitter, and you'll soon get the hang of it. Deliberately over-drive the amplifier and see how quickly you can detect the results (pattern 6-44C) on the scope pattern. Observe the meter action, too. You'll see that you can't draw *any* grid current without spoiling the picture. You'll also see that when the scope picture is right the plate current stands still on all modulation peaks. The screen current will probably be just a bit negative. Output will absolutely not exceed 35 per cent of the input. If it does, you've got some meter inaccuracies, or you're cheating on the interpretation of the scope pattern. The scope is the final authority; you *have* to believe it.

Now, once over lightly again. Loading is all-important. Keep it at the maximum output you can get for a given value of grid drive. Recheck it for every frequency change or change in plate voltage. Grid current will *always* be zero. Grid drive can be lower than optimum as regards output, but never more than optimum. (You can read grid *voltage* for a reference on amount of grid drive, if you like.) The scope will tell you very clearly the minute you go too high. So will the sound of the signal, but this may be hard to determine, if your receiver overloads on your own signal. Most receivers will. Final plate current will rise with increasing grid drive, but it must stand still during modulation. If it kicks on modulation peaks, you've got distortion, and very likely splatter.

All adjustments react on one another to some extent, and each time you change any operating condition you have to go through the routine completely again. This sounds as if you'd spend the rest of your life tuning the rig, but once you

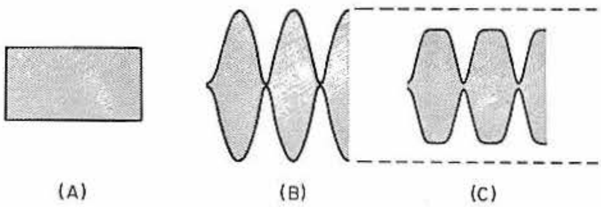


Fig. 6-44—Typical oscilloscope envelope patterns. Unmodulated carrier is shown at A. The single-tone pattern for 100 per cent modulation is shown at B. Peaks should rise to twice the enevelope-pattern height, and valleys should just reach the center line. The effects of excessive drive or too-light loading, or both, on a linear amplifier are shown at C. Note the flat-topping and small increase in amplitude over the unmodulated envelope.

get the hang of it you can make the necessary corrections in seconds.

Using Other Modes

Since a.m. linear is the most critical of all, it is in order to switch to any other mode without making any adjustments, if you want to switch instantly. A good linear is more versatile than this, however. It's possible to do a lot better than the a.m. conditions on sideband, and still stay in the AB_1 mode. Efficiency on c.w. will shoot up markedly with just a slight increase in grid drive, with no other changes. Same for f.m., which is identical to c.w., as far as the tubes in the final are concerned. If you want the ultimate in c.w. or f.m. output, switch to 250 volts on the screen, and run up the grid drive some more. Drive level is very uncritical, so about all you have to watch for is to keep the final input below the kilowatt level, and avoid swinging the plate current on f.m. Readjustment of the plate tuning and loading will be needed for top efficiency. Plate-modulated voice service is quite similar to the c.w. conditions, except that the maximum plate voltage permissible is lower with most tubes. Grid drive requirements are usually slightly higher for good plate modulation conditions than for c.w. or f.m., and the bias should be juggled for best modulation characteristics. Scope indication should be like Fig. 6-44B.

V.h.f. Stations, Transceivers and Transverters

Buy or build? This question faces every new amateur, and it is likely to remain with him as he advances in the art. We have no quarrel with the fellow who chooses the all-commercial route, so long as he is well-informed and uses his station intelligently and with consideration for others. Buying is the quick and often easy way to get started in amateur radio. There are still sound arguments for building one's own, however, and plenty of hams, new or old, still play the game that way.

First, there is the matter of cost. Admittedly, parts cost money these days, but if the job is done wisely the v.h.f. enthusiast can build a complete station for much less than similar facilities would cost ready-made. Then, nearly all commercial gear is a compromise in one or more ways. When you are going to build it yourself, you can design your station to do what *you* want it to do, and to look the way *you* want it to look. You don't pay for anything that you don't need. A station that works from 80 through 6 meters, for example, is a poor investment for the fellow with no interest in anything but v.h.f. work. It's a sure thing that a v.h.f.-only rig will deliver a lot more 6-meter watts per dollar and better v.h.f. reception than the multiband variety.

But perhaps most important is the nature of the hobby itself. Despite all the easy approaches to it, ham radio is still a *technical* avocation. The fellow who learns his way around is going to get more out of hamming than the mere purchaser of boxes. When you collect the parts (and perhaps make a few of them), put a station together with your own hands and skill, and make it work to your satisfaction, you have accomplished something. The end result is *your* station in a way that no commercial package can ever be, and you will be a better ham for having done the job!

You may hear that "nobody builds ham gear any more." Before accepting this as gospel, consider the story of the equipment shown in Fig. 7-1. Described serially in *QST* in 1961, this sectional two-band v.h.f. station caught on quickly. Before the year was out, demand for the *QST* issues containing the series had cleared the shelves at ARRL Headquarters of a stock that would have lasted for years, ordinarily. A reprint of the four *QST* articles was made and offered for sale at 50 cents. It sold by the thousand, and is still a fast-moving item. Today this series stands as the most-used v.h.f. material ever published, and one of the most successful constructional items in all *QST* history.

Station Planning

Too often, amateur stations "just grow," rather than developing along planned lines to make the best use of the considerable financial outlay they usually represent. This applies to equipment purchased ready-made, as well as to that built at home. We accumulate transmitters, receivers, converters, modulators, and so on down the line, with little thought as to their integration into a working unit for v.h.f. communica-

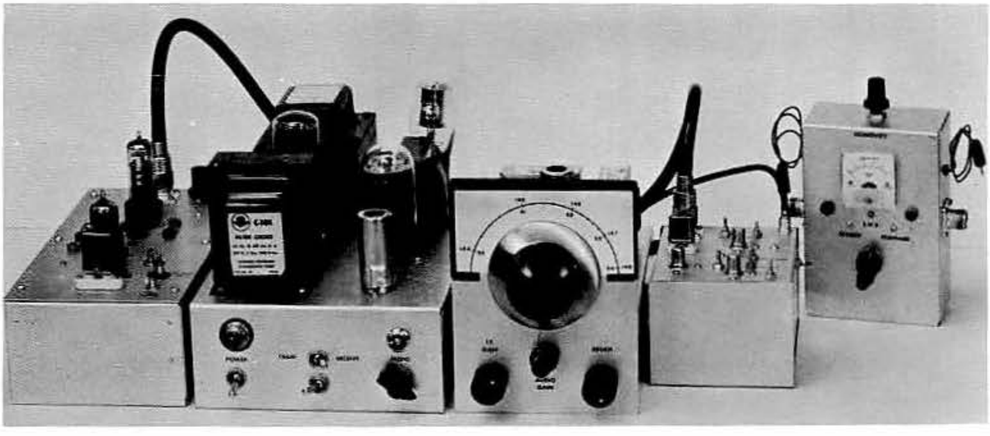


Fig. 7-1—The complete v.h.f. station, shown here set up for 50 Mc. The transmitter, left, and converter, right, have companion plug-in units for 144 Mc. The control unit, left center, contains the power supply and modulator, and all units of the station draw their power from it. The simple tuner, right, may be omitted if the builder has a communications receiver. At the far right is an s.w.r. bridge that doubles as a test meter.

tion. Some commercial gear leans to the opposite extreme—the one-box station that may be neat and unobtrusive, but is often lacking in versatility.

Amateurs are individualists. We like our stations to be unique, tailored to our special needs. With some advance planning we should be able to assemble a station that is both effective and versatile, without its necessarily becoming elaborate or tremendously expensive. With these objectives in mind, most of the equipment we describe here is built unit-style, with few built-in heavy items like power supplies and modulators. These tend to be static in design; long-term investments that can be used with a succession of r.f. units we may wish to build and try. Subassembly design has much to recommend it, and cost is by no means the only consideration. The ability to try different circuits without becoming involved in the kind of rat's nest all too often seen in amateur stations should rate high in our planning.

A COMPLETE TWO-BAND STATION FOR THE V.H.F. BEGINNER*

The station shown herewith was designed to get the newcomer off to a good start on the v.h.f. bands. You may not need all of it. If you already have a good communications receiver, for example, you will have only a passing interest in the simple tuner at the center of Fig. 7-1, described in detail in Chapter 4. Maybe you're a one-band man. In that case you'll want only one of the transmitter look-alikes described in Chapter 6, one of which occupies the left side of this picture. One of the converters, described in Chapter 4 and shown to the right of the tuner, will take care of your receiving needs.

The important point is that whether you go for 50 or 144 Mc., whether you want to build your station piece by piece as you can afford it or get it all over in one project, whether you have a communications receiver or must provide a means of receiving with your v.h.f. converters, this station gives you just the items you need. Nothing necessary is omitted, and nothing in the way of useless glamour is included. Each unit is intended to do its job well, and to allow for improvement of the station later on.

*The equipment shown here was described originally in a 4-part series in QST for July through October, 1961. Demand for these issues created by the series sold out back issues at once, so a reprint of the 4 articles was made. It is still available from ARRL Headquarters for 50 cents for the complete series. Included with each reprint is a set of templates for drilling the principal surfaces of the tuner, transmitters and converters.

The transmitter r.f. units are stable and efficient. They include provision for c.w., and may be adapted to variable-frequency control. They will make fine exciters for high power later on. The modulator and power supply use quality components, and are handy items around any ham shack. Control circuits are included, so that the question of how to use the gear in actual communication (so often left unanswered in items supposedly for the beginner) is completely taken care of. The receiving system is a little different from anything you've seen in modern v.h.f. articles, but it does the job. You can receive c.w. with it, as well as a.m. or f.m. phone, and it can even produce readable s.s.b. signals with a bit of care. The converter "front ends" for 50 and 144 Mc. are excellent performers, and if you decide later to use a communications receiver in place of the tuner, they will give you v.h.f. reception second to none.

Last, but no means least, nearly every v.h.f. station description tells the builder to use a standing-wave bridge in tuning up the transmitter and adjusting the antenna—but few home-built s.w.r. bridges will work on 6 or 2. This station includes a v.h.f. s.w.r. bridge. It is described in detail in Chapter 11.

The complete station is shown in block-diagram form in Fig. 7-2, with much the same arrangement of component parts as in the photograph, Fig. 7-1. How the various units work

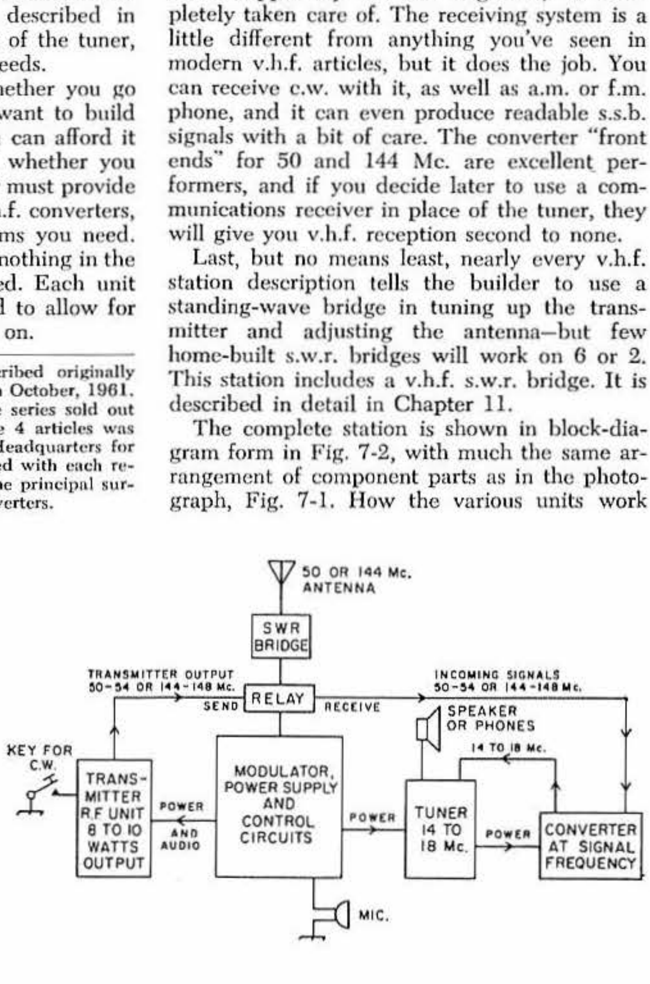


Fig. 7-2—Block diagram of the two-band v.h.f. station. A central unit contains the speech equipment, power supply and control circuits. The antenna connects to a send-receive relay on the back of this unit through a standing-wave bridge. The transmitter r.f. assemblies for 50 or 144 Mc. plug into the left side of the control unit, and a tuner for 14 to 18 Mc. into the right side. Converters for 50- or 144-Mc. reception plug into the right side of the tuner. The various units may be interconnected with cables, instead of being plugged together, if operating convenience so dictates.

together should be self-evident. All except the control unit are described elsewhere in this book, so we'll run over each only briefly here.

Receiving

Some means of listening is usually the first requirement of the newcomer, so we will consider reception first. It is almost standard practice in v.h.f. circles to employ a converter of some sort, which changes the signal on 50–54 Mc. or 144–148 Mc. to some lower frequency before it goes through the detection process. There are several reasons for this, but perhaps the most important is selectivity. It is difficult if not impossible to attain the desired degree of selectivity at 50 Mc. or higher, but the difficulty decreases with frequency. This is the main reason for the use of so-called double-conversion receivers, even on our lower amateur bands.

In a communications receiver, a 14-Mc. signal, for example, may be converted to 455 kc. or lower, where it is more readily amplified than at the original frequency. In our receiver we convert from 50 or 144 Mc. to 14 Mc., and our amplification and detection take place at the latter frequency. This is not quite as good as if it were done in the manner of the communications receiver, which would include a second conversion, but it does have advantages for the home constructor, not the least being simplicity. We can tune 14 to 18 Mc. with our little tuner, without the tracking problems that bedevil the designer of a superheterodyne-type 14-Mc. receiver, and the whole works involves only a broad-band amplifier, a detector, and a simple audio system. These jobs can be handled easily with three tubes.

Ahead of this we use crystal-controlled converters, which amplify the signal and then convert it to some frequency between 14 and 18 Mc., at which point our tuner takes over. If you decide to go to the communications-receiver method of reception later on (a desirable step if you can afford it), these converters will give you v.h.f. reception of the highest caliber. The simple tuner need not be abandoned, however. It can serve for portable operation, or for use under any circumstances where the ultimate in sensitivity and selectivity are not required. For the full story on the tuner, and the converters that go with it, see Chapter 4.

The Transmitters

You can build a transmitter for 50 or 144 Mc. with fewer parts and simpler circuits than the ones shown here. You might even develop the same power output for a bit less money than we have spent. But simplicity and low cost can be delusions. We started with v.h.f. crystals, for example, and came up with a one-tube 6-meter rig and a two-tuber for 2. They were unstable, both as to warm-up drift and frequency shift under keying and modulation, so they were ruled out. The tube lineup we finally used would work with fewer tuned circuits, but it might then radiate strong unwanted harmonics, and be something of a neighborhood nuisance.

These transmitters are designed to their job well. You will be justifiably proud of the quality of their signals, on voice or c.w., and they are well-adapted to increasing power, when you're ready for that next step. They are built just as much alike as possible, to keep parts procurement and the need for spares to a minimum. Their construction and operation are described in Chapter 6.

THE MODULATOR, POWER SUPPLY AND CONTROL UNIT

One of the problems often encountered in putting together a complete station is a satisfactory means of controlling it. Complete control of our station is built into the modulator and power supply portion. It is the central item of the station, designed so that the transmitter r.f. assemblies plug into its left side and the receiving gear into the right side. In the audio portion a 12AX7 dual triode speech amplifier drives a 6L6G modulator. The microphone may be either crystal or high-impedance dynamic. The power supply for the entire station is included, as are the circuits for send-receive switching. A coaxial antenna change-over relay is mounted on the rear wall. The standing-wave bridge and test meter is described in Chapter 11.

The chassis is 7 by 12 by 3 inches in size. Layout of parts is not critical. If the general physical arrangement shown in the photographs is used there should be no problems encountered in building the unit. Looking at it in Fig. 7-1, we see the speech amplifier tube in the foreground. To the left is the filter choke. In back of the 12AX7 is the 6L6G modulator, and in line thereafter are the modulation transformer and the voltage-regulator tube. At the rear of the picture are the rectifier tube and the power transformer.

On the front wall at the left are the main a.c. switch, S_1, and a red pilot light. The upper of the two toggle switches is the dual send-receive control, S_2. This switches the high voltage from transmitter to receiver, and also applies a.c. to the coaxial relay, which is mounted on the back of the unit (see bottom view, Fig. 7-3). The second switch, just below the send-receive control, is used to apply voltage to the receiver while the transmitter is on, if desired. This enables the operator to monitor his transmissions, and also can be used for duplex operation (above 51 Mc.) if separate transmitting and receiving antennas are used. At the right are the microphone connector and the gain control for the speech amplifier.

From the bottom view it is obvious that there is plenty of room for the parts. All leads that are not part of the components themselves are made with shielded wire (Belden 8885). This may not be entirely necessary, but it is a good precaution against r.f. feedback and hum troubles. Liberal use of terminal strips makes for a neat and trouble-free unit. Note that there are octal power sockets on each side of the chassis. These carry the heater and plate voltages for the transmitters, J_3, left, and receiving gear, J_2, right, as the unit is viewed from the front.

In the bottom view the coaxial relay is seen on the rear wall of the chassis. Note that the a.c. terminals are bare in this picture. Before the unit is put into service these leads should be covered securely with plastic tape or insulated sleeving. The coaxial connectors come in close proximity to them when the cables are connected, and a shock is likely if the relay terminals are not protected. In the upper left of the picture is the power transformer. Below it are the regulator tube socket and one of the electrolytic filter capacitors. This capacitor was added during the testing of the equipment, only the dual 8-μf. capacitor at the upper right having been included originally. A triple 8-μf. 450-volt capacitor or three separate 8-μf. 450-volt capacitors can be used equally as well. The modulator and speech-amplifier components are at the lower right of the bottom view. The main control switch and pilot socket are in the upper right.

In the schematic diagram, Fig. 7-4, the main control switch is shown in the *off* position. When it is closed, the power circuits are activated, applying filament and plate voltage to the rectifier, and heater voltage to the modulator and to whatever equipment is plugged into it. The send-receive switch, S_2, is shown in the *receive* position, which is the way it should be left when the station is turned off. With the power on, 150 volts, regulated, is applied to the amplifier and detector tubes in the tuner and to all tubes in the converter, through Pin 6 of socket J_2. The audio stage in the tuner receives its high-voltage d.c. through Pin 2 of J_2. When S_2 is in the *send* position, a.c. goes to the coil of the coaxial relay, K_1, and high-voltage d.c. to the transmitter through Pin 2 of socket J_3, and to the speech amplifier and modulator tubes.

Note that the plate current of the transmitter output stage flows through the secondary of the modulation transformer, T_1. The fluctuating audio voltage from the modulator, also in this secondary winding, adds to and subtracts from the d.c. voltage that reaches the amplifier plate through Pin 6 of J_3. This, in simple terms, is the

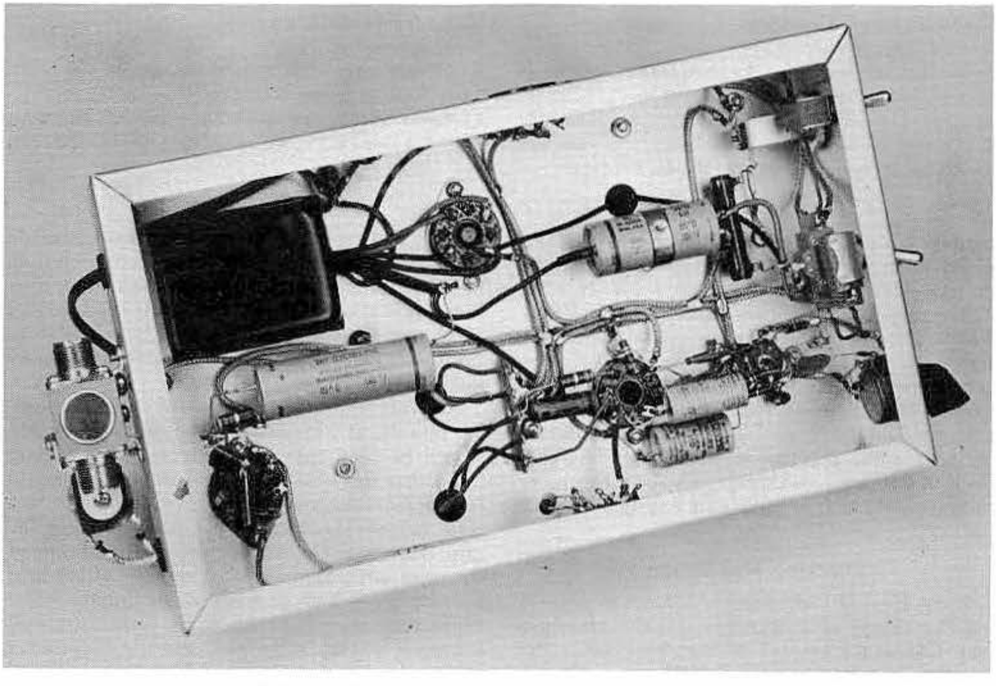


Fig. 7-3—Interior of the modulator and power-supply assembly. Note the antenna changeover relay mounted on the back of the chassis.

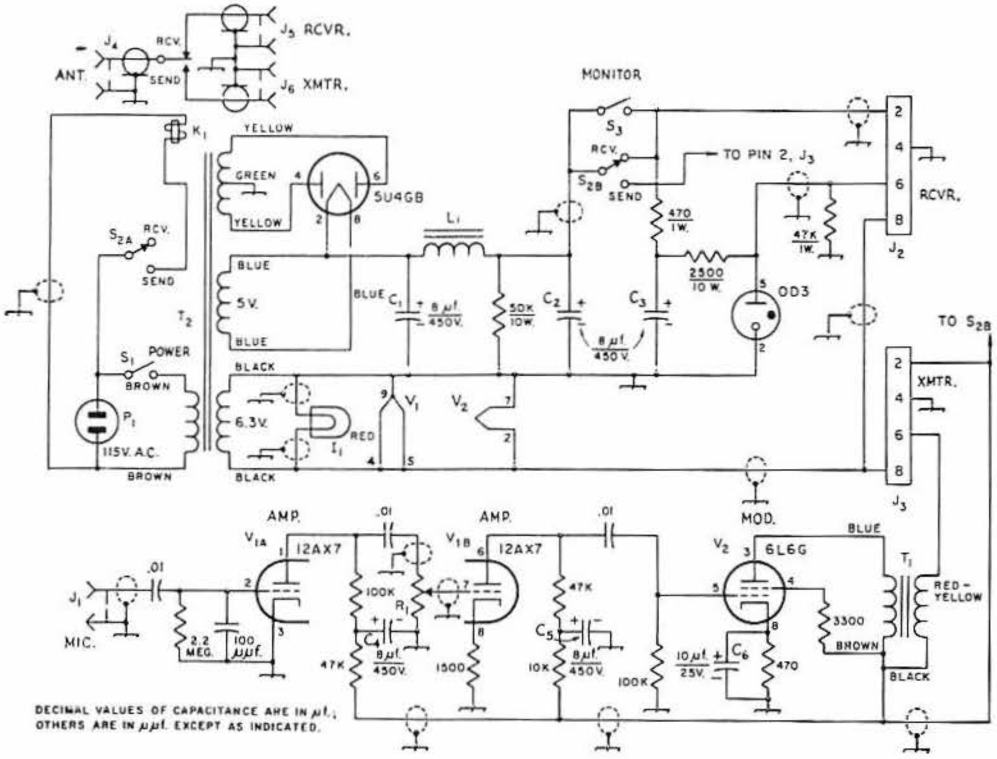


Fig. 7-4—Schematic diagram and parts information for the modulator and power supply. Capacitor values in μf. unless otherwise indicated. Resistors ½ watt unless specified.

C_1–C_5 incl.—8 μf. 450-volt electrolytic. C_1, C_2 and C_3 can be separate or combined in one housing.
C_6—10 μf. 25-volt electrolytic.
I_1—Pilot lamp and socket.
J_1—Microphone connector (Amphenol 75-PC1M).
J_2, J_3—Octal socket (Amphenol 77 MIP-8).
J_4, J_5, J_6—Coaxial fittings on relay K_1.
K_1—Coaxial antenna change-over relay, 115 volts a.c.
L_1—4.5 hy. 200-ma. filter choke (Stancor C-1411).
P_1—115-volt plug.
R_1—0.5-megohm control, audio taper.
S_1, S_3—Toggle switch, s.p.s.t.
S_2—Toggle switch, d.p.d.t.
T_1—20-watt modulation transformer, pri. 10,000 ohms, sec. 3, 5, and 8000 ohms (Triad M3X).
T_2—Power transformer, 270-0-270 volts, 200 ma.; 5 v., 3 amp.; 6.3 v., 4 amp. or more (Stancor P-8172).

modulation process: making the transmitter amplifier plate voltage vary in relation to the audio voltage developed in the speech amplifier.

The switch S_3 is shown in the open position, which allows the receiver to go off when the transmitter comes on. Closing S_3 keeps the receiver operating during transmitting periods, for monitoring or duplex work. In using the outfit this way you will probably have to use earphones on the receiver to prevent audio feedback. Keep the audio gain control on the tuner turned down low, or your ears will take a beating.

Checking Modulation

Some kind of lamp load is helpful in observing the effects of modulation on the transmitters. Connect a crystal or high-impedance dynamic microphone to the modulator, and with the audio gain turned down, adjust the transmitter for maximum output indication as described in Chapter 6. If the lamp load is made of several blue-bead pilot lamps connected in parallel, the bridge can also be used as an auxiliary indication of power output. If the load is a 115-volt lamp the mismatch may be too high to use the bridge effectively. For more on dummy loads, see Chapter 11.

Advance the audio gain slowly while speaking into the microphone. As the gain is increased it will be seen that the brilliance of the lamp indication increases with speech. There should be appreciable brightening, but the plate and grid currents should not vary. Adjustment of the grid drive and the loading affect the ability of the transmitter to modulate properly. If the grid current is too high or too low, modulation may cause the currents to fluctuate, indicating that the voice quality will suffer and the transmitter may cause interference outside its normal passband. Most effective modulation will be obtained at the highest gain setting that can be used without causing the plate or grid current to fluctuate.

It is possible that the modulation may be low, even if the transmitter is working properly, due to limited output from the modulator. The modulator will deliver 7 to 8 watts of audio without severe distortion. This means that the transmitter should not run much over 15 watts input if full modulation is to be achieved. If you get reports of "low modulation" from fellows you work, reduce the transmitter input slightly by detuning the loading capacitor and readjusting the plate tuning for the point that gives the greatest output with the least plate current. A current of 60 to 70 ma. will be about all that the modulator will handle well, though on c.w. it will be possible to increase the loading to the point where the final stage runs 20 watts input or more. This is worth having, though the difference between it and the 15 watts that can be fully modulated will be just barely noticeable at the receiving end.

TRANSISTOR PORTABLE STATION FOR 50 MC.

Though "working portable" from the high spots has been an integral part of the v.h.f. game since the earliest times, it remained for the age of transistors to bring lightweight portable v.h.f. stations into the realm of full practicality. The 50-Mc. station in Fig. 7-5 is complete, including even the antenna system and microphone, yet it is a mere 5-pound handful. It delivers a good-quality voice signal of up to one watt, and its receiver will pick up any a.m. signal that you could hear on the best home-station setup, yet the station will operate for many hours on its self-contained pack of D cells. You can run the rig from the car battery when it is convenient to do so, but you can also take it to any spot you can reach on foot, and have it ready to go in minutes.

What's Inside

The handmade sheet aluminum box is 4¾ by 6 by 9 inches in size. Inside are separate units for transmitting, receiving, modulation and power, any of which can be changed without dismantling the others. The receiver is a small imported pocket broadcast set, with a crystal-controlled converter ahead of it. The converter is seen in the left foreground of Fig. 7-6. Its adjustments are reached through holes in the right side of the case, as viewed from the front. The broadcast receiver, attached to the front panel, shows in the left rear corner of Fig. 7-6. At the right rear, in back of the battery pack, is the transmitter r.f. unit. Just above this assembly is a readymade 1-watt audio amplifier, modified for modulator service. Each of these units will be described in detail.

With the transceiver, in Fig. 7-5, are the microphone, a 35-foot "long-wire" antenna, a dummy load for testing the transmitter, and a small antenna coupler built into a plastic parts box. These items, a Minilog and miscellaneous small tools and spare parts are carried in a zippered plastic "gym bag" 5 by 9 by 12 inches in size. A lightweight portable 3-element beam that makes this little station "really talk" is described in Chapter 9.

Transmitter R.F. Unit

The transmitter is shown in Figs. 7-7 and 7-9, with its circuit diagram in Fig. 7-8. Parts are

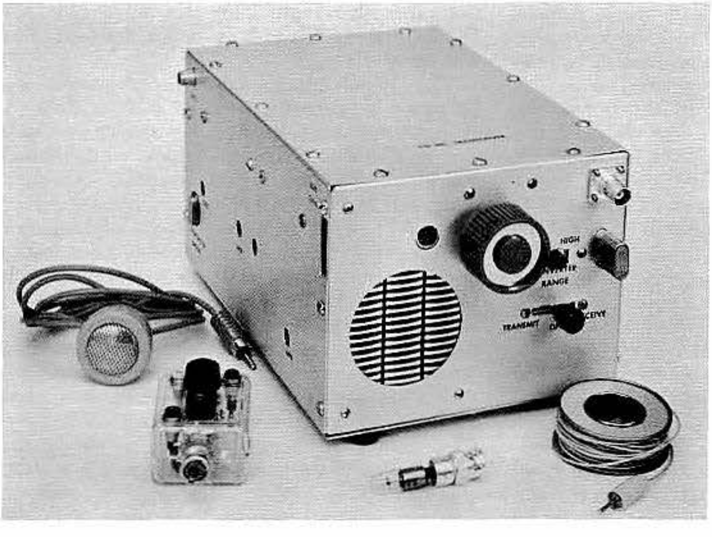


Fig. 7-5—The 50-Mc. portable station, complete with all necessary operating accessories—total weight: under 5 pounds. Accessories grouped around the transceiver: the microphone, miniature antenna coupler, pilot-lamp dummy load and a 35-foot "long-wire" antenna.

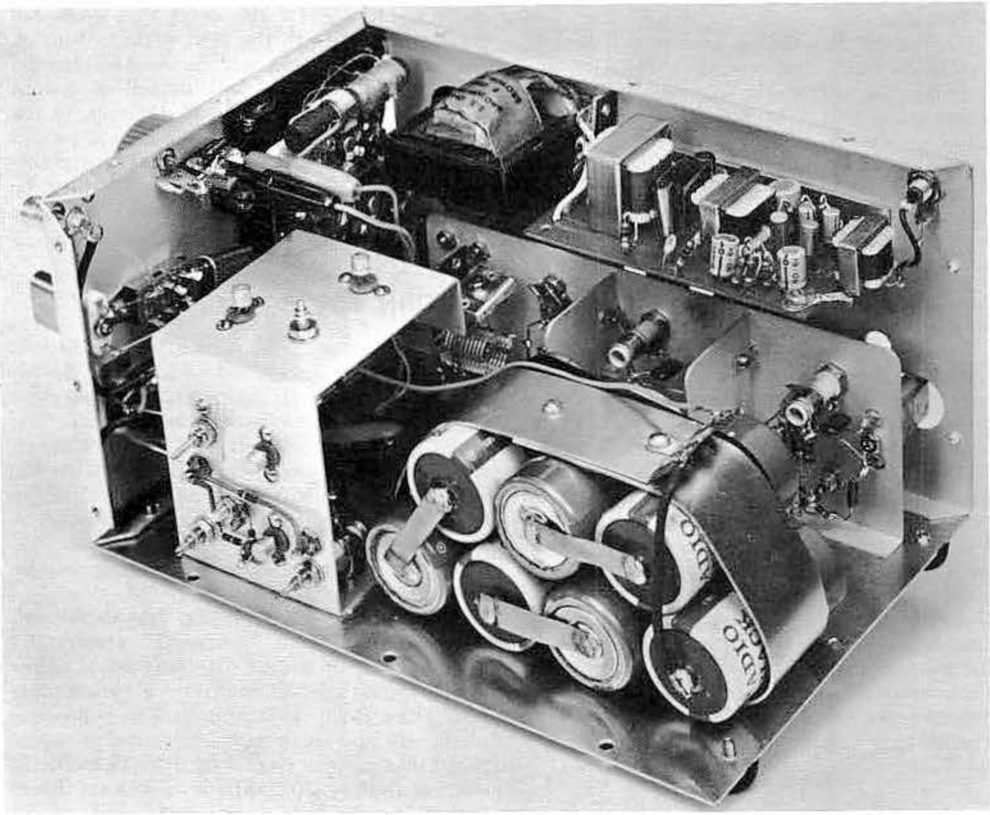


Fig. 7-6—Interior of the 50-Mc. transistor transceiver, with top plate, right side and rear panel removed. Parts of the switching circuits and the small broadcast receiver used for the i.f. system are seen on the front panel. The C-shaped subassembly at the left is one of several converters tested in the transceiver. The transmitter r.f. assembly is seen in back of the package of 7 "D" cells. At the upper part of the left-side panel is the modulator.

mounted on an aluminum plate made from a sheet 3 by 7 inches, with ⅜ inch folded up at the bottom. This fastens to the transceiver bottom plate with self-tapping screws. The oscillator and buffer stages use 2N706 transistors. These are at the left side of Fig. 7-7 and the right of Fig. 7-9. Aluminum shields 2 by 2½ inches are mounted on spade lugs at 2 and 4 inches in from the left side, as seen in Fig. 7-7. Leads from L_2 and L_4 run through these shields and are insulated from them by sleeves of polyethylene made by removing the conductors from small pieces of RG-58 or 59 coax.

The output stage has two silicon v.h.f. power transistors in parallel. Several types are usable, but the least expensive we've found are the Archer (Radio Shack) 27R131 shown here. RCA's 2N3553, 2N3866 and 2N4427 also work well, and can be run at higher input if desired. 2N706s work well in the final, but will not stand amplitude modulation voltage peaks encountered with 12 volts on the collectors. Nearly all silicon v.h.f. transistors will do for the oscillator and buffer, but the 2N706 has a higher dissipation rating than most. They also can be obtained for as low as five for a dollar from surplus sources.

The safe dissipation rating for transistors can be raised by even the simplest of heat sinks. A strip of thin brass or flashing copper $^3/_{16}$ inch can be bent into keyhole shape and slipped over the 2N706 case for this purpose. The brass plates holding the final stage transistors together (Fig. 7-7) serve the dual purpose of heat sink and parallel collector connectors. Dimensions are not critical, but ours are 0.041 by ⅜-inch brass, about 1¾ inches long. Aluminum would be equally good. Be sure that these do not touch the mounting plate or the socket-mounting screws at any point, as the collector and case are connected together in power transistors, and thus the case has the supply voltage on it. Center-to-center spacing of the holes should be the same as that of the transistor sockets, one inch in this case.

Various output circuits were tried, with the series-tuned center-tapped arrangement shown in Fig. 7-8 working out best for this setup.

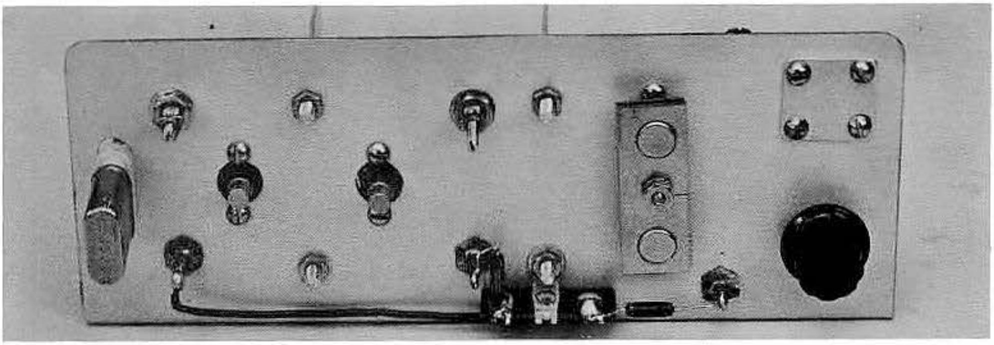


Fig. 7-7—Transmitter portion of the 50-Mc. transistor rig. At the left side are the crystal oscillator and buffer stages. The two transistors in the output stage are connected in parallel by means of two brass plates, which also serve as a heat sink. The amplifier collector circuit is tuned by means of the knob at the lower right, the surface of which is slotted to permit adjustment with a screwdriver, through a hole in the left side of the transceiver case. The crystal and the two tuning slugs are also provided with access holes.

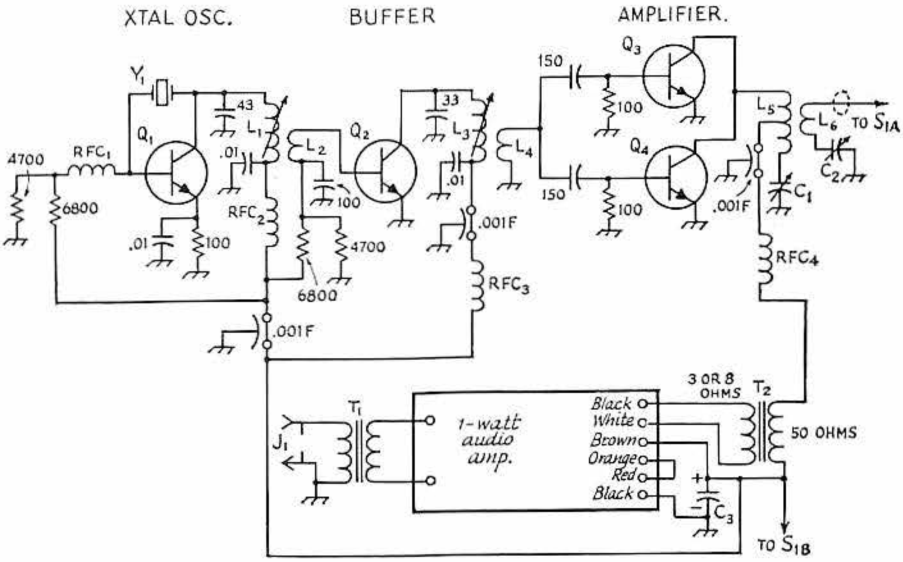


Fig. 7-8—Schematic diagram and parts information for the transmitter portion of the 50-Mc. transceiver. Resistors are all ¼ watt. Decimal values of capacitance are in μf; others in pf. Suffix F indicates feed-through type. All others not described are Mylar or dipped-mica, 50-volt rating or more. The modulator is shown in outline form only, since it is a ready-made unit. Lead colors given are for Radio Shack audio amplifier, type 277-038, having a 1-watt rating.

C_1—35-pf. subminiature variables (Millen 25035E).

C_2—180-pf. mica trimmer (Arco 463)

C_3—5-μf. 25-volt electrolytic.

J_1—Phono jack.

L_1, L_3—5 turns No. 24 enamel, closewound on ¼-inch iron-slug form. (Miller No. 4501, with 3 turns removed or wind on No. 4500 form.)

L_2, L_4—2 turns insulated wire wound near bottom end of L_1 and L_3, respectively.

L_5—10 turns No. 20, 16 t.p.i., ½-inch diam., center-tapped (B & W No. 3003 Air-Dux 416T, PIC 1730).

L_6—2 turns insulated wire around center of L_5.

Q_1, Q_2—2N706 or equiv. See text.

Q_3, Q_4—Silicon v.h.f. power transistor, 1-watt or higher dissipation. See text.

RFC_1-RFC_4, incl.—8.2-μh. iron-core r.f. choke (Millen J300-8.2).

T_1—Input transformer, high-impedance microphone to amplifier input, 200k to 1000 ohms (Archer 27-1376).

T_2—Output transformer; 45 to 50-ohm primary, 3.5 or 8-ohm secondary. Connect low-impedance winding to amplifier output, and run final-stage collector current through 50-ohm winding (Knight 54D4147).

Y_1—Third-overtone crystal, 50.11 to 54 Mc. (International Crystal Mfg. Co. Type F-605).

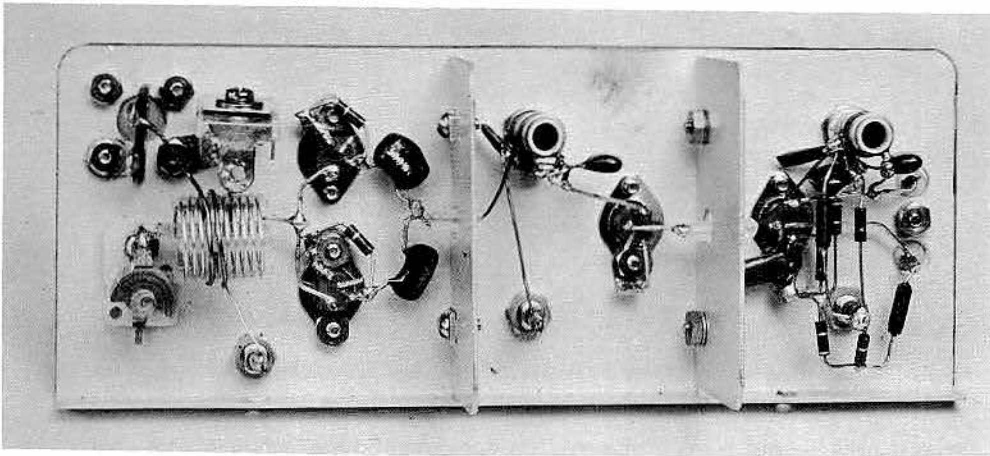


Fig. 7-9—Back view of the transmitter, with output stage at the left. Partitions isolate the three stages; crystal oscillator at the right, buffer at the middle.

output is taken off through a series-tuned loop, L_6, wound around the midpoint of L_5 The series capacitor C_2 is a high-minimum mica trimmer, visible directly over the tank coil in Fig. 7-9. It can be adjusted for optimum transfer to a 50-ohm load and left set thereafter, since adjustment is not critical.

Modulation

The audio amplifier used for the modulator (Radio Shack 277-038, 1-watt rating) has an output transformer with a low-impedance secondary. This must be replaced with one suitable for modulation purposes, or a step-up transformer can be added. We chose the latter, as it was easier to find than one designed specifically for modulator service. An input transformer to match the high-impedance microphone must also be added. The extra transformers, T_1 and T_2 in Fig. 7-8, are visible in Fig. 7-6, mounted at opposite ends of the amplifier. The modulation transformer is connected "back to back" with the output transformer of the amplifier, and has the collector current of the final stage of the transmitter running through its 50-ohm winding.

The amplifier has p.n.p. transistors, so it is set up for positive ground, as is the broadcast receiver. In using the transceiver in negative-ground cars (U.S. standard) the "ground" side of the amplifier and broadcast receiver must be isolated from the transceiver case. The amplifier is mounted on an aluminum bracket, making sure that the mounting screws do not come in contact with the positive-voltage circuits of the module. Parts of the amplifier circuit that connect to the positive lead (brown lead in the unit used here) are bypassed to the transceiver case with an electrolytic capacitor, C_3 in Fig. 7-8.

The amplifier unit is intended for 12-volt service, but it works well at lower voltages. Its output tracks with the input to the final stage of the transmitter as the supply voltage is changed, so the modulation percentage remains about the same regardless of the power source used.

Signal quality and modulation percentage depend on many factors. With our operating conditions the best modulation is obtained with audio applied only to the collectors of the amplifier stage, and with the final collector circuit detuned slightly on the high-frequency (low capacitance) side of resonance. When tuned for maximum output the stage shows little upward modulation, and when C_1 is detuned to the

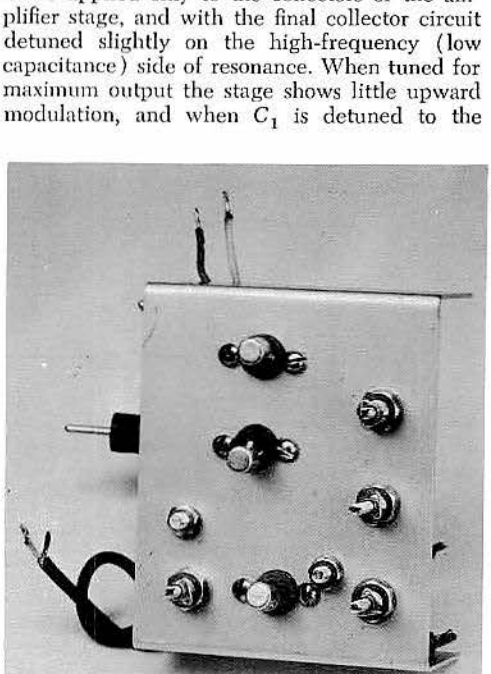


Fig. 7-10—Front view of the converter portion of the 50-Mc. transceiver. Core studs at the right side are for adjusting the r.f. amplifier collector circuit, the mixer base circuit, and the oscillator collector circuit. The r.f. stage input circuit is at the lower left.

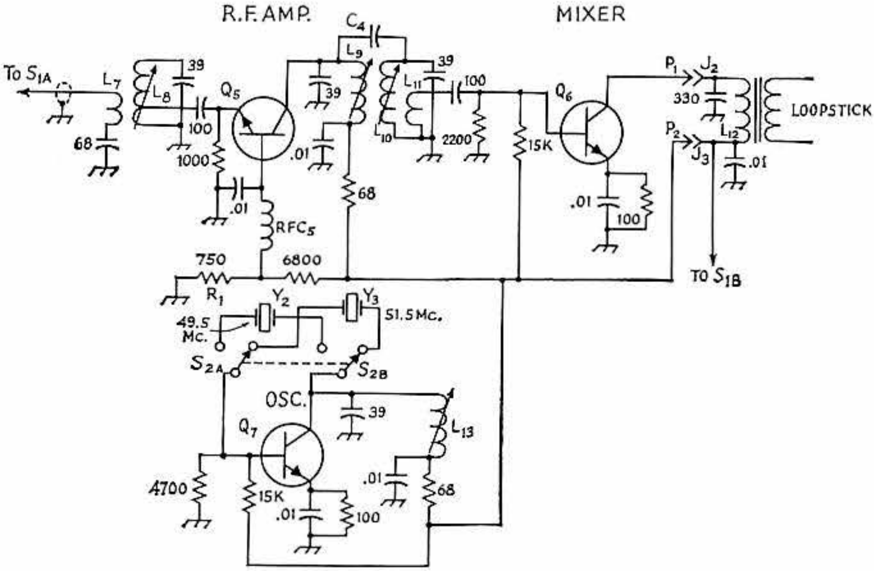


Fig. 7-11—Schematic diagram and parts information for the transistor converter. Decimal values of capacitors are in μf; others in pf. All are Mylar or dipped mica, 50-volt rating or more. Resistors are ¼-watt composition. Parts are numbered serially following those of the transmitter.

C_4—Leads of insulated hookup wire twisted together 4 turns. See text.

J_2, J_3—Insulated tip jack (Johnson 105-800).

L_7—2 turns of the inner conductor of the lead to S_1, wound over bottom turns of L_8. See text and Fig. 7-12.

L_8, L_9, L_{10}, L_{13}—6 turns No. 24 enamel, on ¼-inch iron-slug ceramic form (Miller 4500, or 4501 coil with 3 turns removed.) L_8 is tapped at 2 turns from ground end. If made from prepared coil, unwind, clean insulation at tap point, solder on tap, and rewind. Space out turns on any coils if needed to obtain resonance within core range.

L_{11}—2 turns No. 24 enamel, wound over bottom turns of L_{10}.

L_{12}—About 8 turns No. 24 enamel, wound over turns of built-in loopstick of broadcast receiver. Position and number of turns not critical. 330-pf. capacitor also uncritical.

P_1, P_2—Insulated solderless tip plug (Johnson 105-300).

Q_5, Q_6, Q_7—Silicon v.h.f. transistor (RCA 40235 used; 40236 through 40240 also tried).

R_1—680 and 68-ohm ¼-watt in series. Check different values for optimum amplifier performance.

RFC_5—8.2-μh r.f. choke (Millen J300-8.2).

S_2—Two-pole two-position slide switch.

Y_2—Third-overtone crystal, 49.5 Mc. International Crystal Mfg. Co. Type F-605.

Y_3—Same as Y_2, but 51.5 Mc., or as desired; see text.

high-capacitance side the quality is poor and the modulation distinctly downward. The amount of detuning needed depends on the collector voltage, increasing with voltage level.

Output capability is about one watt at 9 volts and two watts at 13 volts, but the stage must be detuned to one-half and one watt, respectively, for good modulation. About 300 milliwatts output is possible, with good modulation, at 6 volts.

Transmitter Adjustment

Tuneup is very simple. Checking individual stages for current drain is desirable, and adjustments can be made at lower than rated voltages initially. Operation at 6 volts is similar to that at higher voltages, and it may be safer in the check-out phase. Apply voltage through the oscillator feedthrough capacitor only, at first, and check the current drain. As the slug in L_1 is moved there will be a downward dip in collector current as the crystal begins oscillating, to around 10 ma. at 6 volts. At 9 volts the oscillator current is 15 to 20 ma. Output is enough to light a 2-volt 60-ma. pilot lamp dimly, if a loop of wire is soldered to its terminals and slipped over L_1. Set the slug in L_1 for the highest output at which the oscillator starts readily each time voltage is applied.

Now apply voltage to the buffer through RFC_3, and check current drawn by Q_2. It will rise as the oscillator is tuned toward maximum output, and the pilot lamp load should glow fairly brightly when coupled to L_3. Adjust the stud in L_3 for maximum output. Current drain will be 20 to 30 ma. with the stage working correctly.

Check the amplifier similarly, applying voltage through RFC_4. The current to this stage will be practically nil until drive is applied, after which it is proportional to the drive level. A 6-volt 150-ma. pilot lamp (brown bead, No. 40, 40A or 47) makes a good dummy load when the rig is intended to work into 50 ohms. Other lamps will light up, but they do not approximate

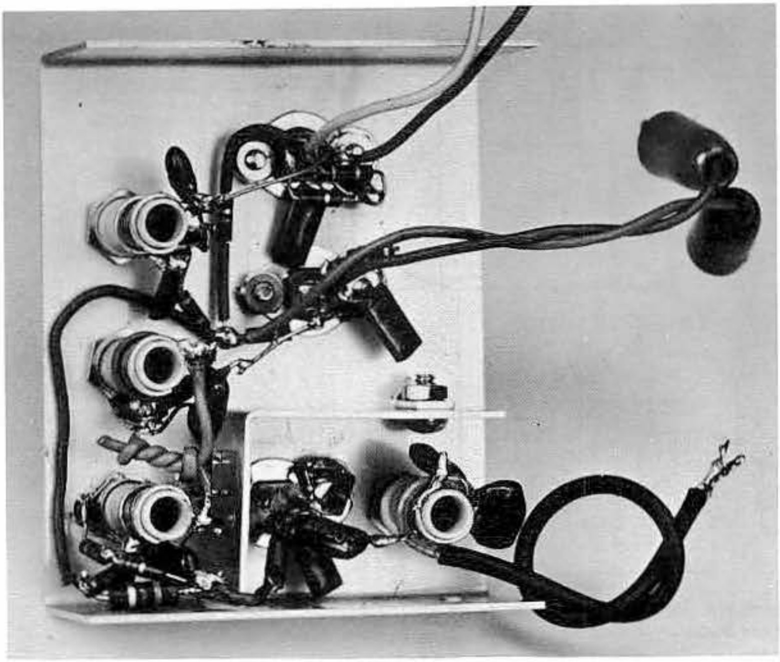


Fig. 7-12—Rear View of the 50-Mc. converter. The r.f. amplifier transistor socket and the input coil are isolated from the rest of the converter by an L-shaped shield, lower right. Leads at the top run to the crystal switch. Those with tips attached plug into jacks connected to the mixer collector winding on the loopstick. Coax at the lower right goes to S_{1A}.

50 ohms at normal brilliance. Solder short wires to the base and plug these into the BNC fitting, or temporarily solder the lamp across the coax lead connected to L_6. A lamp mounted in a BNC fitting is a desirable accessory.

Tune C_1 and C_2 for maximum lamp brilliance, at first. Recheck the settings of L_1 and L_3 also. The lamp will light very brightly at 9 to 10 volts, indicating about one watt output. Peak C_2 for maximum output, and leave it that way. When modulation is to be applied, detune C_1 on the low-capacitance side while talking into the microphone, detuning only enough to get a good upward modulation indication in the lamp. Note the final collector current under these conditions. At full output it will be 150 to 200 ma., at 9 volts, with detuning to 125 to 150 ma. for best modulation. At 12 volts the best setting will be around 150 to 175 ma.

If you have several 2N706s, try various ones in the oscillator and buffer stages, selecting the ones that drive the final collector current to the highest value at the maximum-output tuning condition.

When the detuning procedure outlined is followed the resulting modulation characteristics are at least as good as those of any small pentode or tetrode tube transmitter for the v.h.f. bands. Voice quality is good and "talk power" is high, as there is some inherent clipping effect that tends to prevent excessive modulation and splatter.

The Receiver

Use of a simple crystal-controlled converter working into a pocket broadcast receiver for the i.f. and audio system gives more than adequate sensitivity, and the selectivity is good. A friction-drive vernier, to be described later, provides smooth tuning. There are weaknesses however, as in any very-simple approach. The main problem is spurious responses. Image rejection is inherently low, with such a low intermediate frequency, but this is turned to an advantage by setting up the converter injection so that it can be on either the high or the low side of the signal frequency.

Converter Circuit Features

The schematic diagram, Fig. 7-11, makes most circuit details self-evident. Most silicon v.h.f. transistors work well in these stages. The r.f. amplifier, Q_5, is a common-base stage. Its collector circuit is band-pass coupled to the mixer, Q_6. The mixer collector circuit is a few turns of wire wound over the built-in antenna (loopstick) of the broadcast receiver. The oscillator, Q_7, has one crystal (Y_2) wired to a selector switch, S_{2A}. The other side of the switch, S_{2B}, is connected to a crystal socket on the front panel, so that crystals may be plugged in for Y_3, to do any of several jobs. The crystal socket is omitted from Fig. 7-11 for simplification.

When crystal Y_2 is selected by S_2 the injection frequency is 49.5 Mc. Beating with incoming signals, this produces intermediate frequencies between 500 and 1500 kc. for a signal range of 50 to 51 Mc. The broadcast receiver may not go down to 500 kc. unless its oscillator padder is fudged a bit, but the lowest frequency usable for voice in this country, 50.1 Mc., comes in at 600 kc. If you don't care about tuning as high as 51.1 Mc. the crystal frequency for Y_2 can be modified to suit your desires.

Use of a crystal on 51.5 Mc. for Y_3 permits tuning of the first megacycle of the band in the reverse direction on the broadcast dial. The low end appears at 1500 kc. and 51 Mc. is at the 500-kc. end. This provides a quick solution to image problems that may crop up locally, since image rejection is much better at the 1500-kc. end of the receiver's tuning range. Mobile services around 48 Mc. ride through strongly as images when Y_2 is used, but disappear when Y_3 is switched in. A local MARS net just below the band edge takes over the receiver when Y_2 is used, but gives no trouble with Y_3. On the other hand, Y_3 puts most of the band occupancy in the part of the dial where tuning rate is least favorable. Signals in the upper half of the band (if there are any) appear as images in the tuning range when Y_3 is used. So it boils down to using whichever crystal does the best job under conditions of the moment.

Plugging other crystals in for Y_3 provides coverage of any one-megacycle segment in or near the 50-Mc. band. For ranges other than 50 to 51 Mc. the r.f. circuits must be repeaked for optimum reception, but this is done readily enough by moving the core studs in L_8, L_9 and L_{10}. Repeaking these lower in frequency gives 48 to 49 Mc. with Y_2. Running them out and switching in Y_3 gives 52 to 53 Mc.

Receiver Construction

From Figs. 7-5 and 7-6 it will be seen that the broadcast receiver is mounted on the front panel of the transceiver, with the back of its case removed and the speaker facing forward. No specific dimensions can be given as there is an almost unlimited variety of small receivers available. We recommend that one of the better types be used; a.v.c. action and audio quality are considerably better in most 8-transistor models than in the very cheap 6-transistor ones. The one used here is Radio Shack's 8-transistor job, priced around eight dollars.

Most pocket sets use p.n.p. transistors, and so have opposite battery polarity to that required for the n.p.n. transistors in the transmitter and converter. This poses no real problem, as the receiver cases are plastic and there is no "ground" as such. We drilled holes near the four corners of the case for mounting. With some sets it may be necessary to install wire screening inside the speaker hole to prevent pickup of broadcast stations, but this was not needed with the receiver used here.

A vernier drive for the broadcast receiver dial can be made quite simply. A ¼-inch panel bearing (E. F. Johnson 115-255) is used with a drive shaft of ¼-inch tubing or rod. A small rubber grommet is slipped over the shaft in a position to bear against the edge of the small circular dial of the receiver. The mounting hole for the bearing can be filed slightly oval in shape, to permit adjusting the pressure of the grommet on the dial. You can select your own tuning rate by trying different sizes of grommets. We liked the smallest: one with a ⅛-inch center hole and intended for mounting in a ¼-inch hole. This has to be stretched some to get it on the ¼-inch shaft, but it holds firm and works fine. One grommet will stand up for months of operation.

The converter chassis is a C-shaped piece of aluminum, cut to 2½ by 5 inches and then bent over one inch top and bottom. The physical layout is not particularly critical, except that the holes for the three coils (left side of Fig. 7-12) should be ¾ inch center to center. They are on a vertical line ⅜ inch in from the side of the plate, with L_9 ½ inch up from the bottom. Next above it is L_{10}, with the oscillator coil, L_{13}, at the top. The r.f. input coil, L_8, is ½ inch in from the other side, and the socket for Q_5 is centered approximately between L_8 and L_9. The sockets for Q_6 and Q_7 are along a vertical line 1⅛ inches over from that of the three coils. Q_6 is midway between the center lines of L_{10} and L_{13} (⅜ inch up from L_{10}), and Q_7 is the same distance above the level of L_{13}. The r.f. amplifier is isolated from the rest of the converter by means of an L-shaped shield mounted on spade lugs. The amplifier collector lead runs through this shield to L_9. The converter assembly is held on the bottom plate by two self-tapping screws.

The antenna coupling winding, L_7, is made from the inner conductor of the RG-174/U coax used for the lead to S_{1A}. Strip the braid back about two inches and leave the polyethylene intact except for about ⅛ inch at the end, for soldering to the series capacitor. Wrap the insulated conductor around the winding in the same direction as the bottom turns of L_8, and solder the braid and one side of the series capacitor to a ground lug under the coil mounting. Leave some surplus length in the coax, so that the converter can be removed with the connection to S_{1A} left intact for minor adjustments.

The common positive supply lead and the mixer collector lead are fitted with solderless tip plugs (E. F. Johnson 105-300) which fit into matching jacks (105-800), to permit easy disconnecting for converter removal. (This maker's tiny Rib-Loc plugs and jacks would be fine here.) The jacks are soldered to a tie-point strip visible in Fig. 7-6, just adjacent to the top of the broadcast receiver. The oscillator base and collector leads running to S_{2A} and S_{2B} are made just long enough to reach the terminals of the switch, and must be unsoldered to remove the converter.

Packaging and Power

Presumably the components of the transceiver could be fitted into some standard-size case, but the metal work involved in making your own is not extensive. The front and back panels are 4¾ by 6 inches, with ⅜ inch folded over on all sides. Metal size before bending: 5½ by 6¾ inches. Sheets for the sides are cut 5½ by 9¾ inches, and bent up to 4¾ by 9. Top and bottom plates are 6 by 9 inches. Self-tapping screws hold the case together. Access

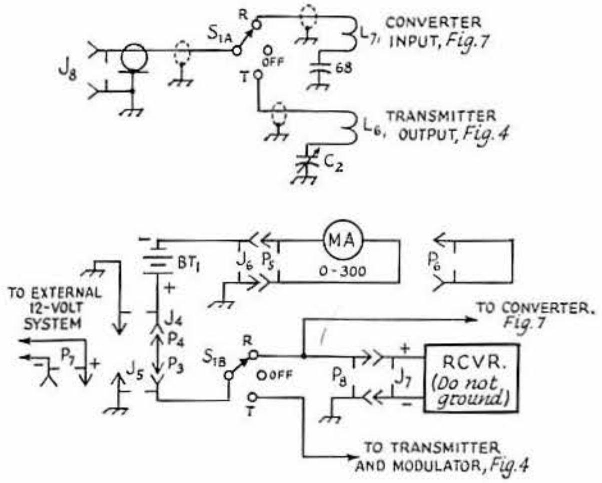


Fig. 7-13—Switching and power circuits for the transceiver.

BT_1—7 or 8 "D" cells in series.

J_4, J_5, J_6—Phono jack.

J_7—Polarized power plug on receiver battery lead (part of broadcast receiver.)

J_8—Coaxial socket, BNC type.

P_3, P_4, P_5, P_6, P_7—Phono plug.

P_8—Similar to J_7, but polarity reversed. Can be removed from top of 9-volt transistor radio battery.

S_1—2-pole 3-position wafer switch, miniature type.

holes for the transmitter and receiver adjustments, and holes for the microphone jack, transmitter crystal, and receiver audio gain control should be located according to the parts used. Jacks for metering in the negative lead, and for internal-external power selection (see Fig. 7-13) can be mounted wherever convenient on the rear wall.

The send-receive switch is a wafer type with horizontal lever action, though any small 3-position 2-pole r.f. switch will do. The crystal switch is an ordinary slide type. Antenna leads are small-size coax (RG-174/U) throughout.

The seven or eight "D" cells are wired in series with strips of metal or stiff wire. They should be piled in 4-3 layers, wrapped with electrical tape to hold them in place, and then clamped in a wrap-around metal strip that is screwed to the bottom plate. This pack has also been made up with 8 cells, for slightly-increased transmitter power.

A variety of power sources may be used. The cells shown are inexpensive by the set, and stand up very well. Transistors have a very great advantage over tubes in overall efficiency, and even smaller batteries can be used if light weight is the primary consideration. Usually it isn't, and we may be more interested in uniform performance or economy. Mercury and alkaline cells are more uniform and longer-lived than ordinary "D" cells, but because of the intermittent nature of the load, and their recuperative powers, the cheaper cells make a logical choice for most users.* Another transistor "plus" is that, with no critical filament temperature to be maintained, the efficiency of the transceiver remains constant over a wide range of battery voltage. Output drops off with fading voltage, of course, but the quality of the signal holds up until the batteries are almost dead.

The transceiver may be run from a car battery or other external power source by removing the jumper (P_3 and P_4, Fig. 7-13) and plugging P_7 into J_5. A cigarette lighter plug and cable to P_7 is handy for operation from a car battery. The car's electrical system must be negative-ground, which is the U. S. standard. Rechargable batteries intended for use with portable TV sets and other medium-drain devices are very nice for the transceiver power, where weight is not a major factor. In case you're worried about running 12 volts on a 9-volt transistor radio, this has been tried with several different types with no apparent damage resulting. If you still want protection, it's a simple matter to install a 9-volt zener regulator on the receiver line. One more possibility for power source is a simple 115-volt supply that delivers 9 to 12 volts d.c., at 300 ma. or more.

Adjustment and Use

Adjustment of the transmitter was described earlier. Monitoring of the total drain can be done with a milliammeter plugged into J_6, Fig. 7-13. If the meter is removed a phono plug with its contacts shorted (P_6) is plugged into J_6. A pilot light connected to a phono plug offers a current check of sorts also. A 150-ma. lamp will light at normal brilliance, or slightly more. A 250-ma. lamp will be bright only on audio peaks. The lamp is only a rough check and should not be left connected in normal operating, as it wastes considerable power. It is handy to have one along with the rig, however, as it tells you quickly whether or not the current drain is normal. A 150-ma. lamp is a must for a dummy load, used as described earlier.

The tuned circuits of the transmitter and converter are broad enough so that repeaking is not necessary in the course of normal use between 50 and 51 Mc., except for the retouching of C_1 in the transmitter. With the twisted-wire coupling capacitor, C_4, made as described, receiver response is nearly flat from 50 to 51 Mc. If there is a severe image problem the front-end selectivity can be improved at any one portion of the

* A review of the various types of batteries suitable for use with transistor gear is given in *QST* for September, 1967.

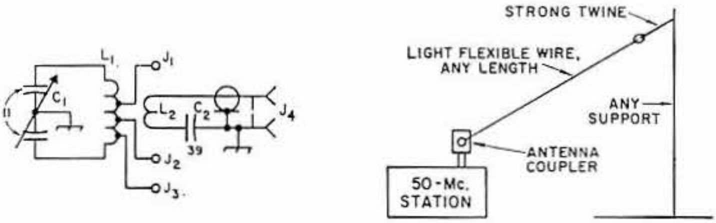


Fig. 7-14—Circuit of the antenna coupler and its application in feeding a long wire in portable work. Tip jacks J_1 and J_2 may be used for a balanced-line system. Any of the three jacks may be used for random-length long wires, merely by checking for best reception. Peak C_1 for maximum signal on receiving. Gain and directivity of the long wire will depend on length and slope.

C_1—11-pf. per section butterfly variable (Johnson 160-211 or 11MB11).
C_2—Fixed ceramic capacitor, 39 to 68 pf. Check with variable temporarily, if possible.
L_1—18 turns No. 24, ½ inch diameter, 32 t.p.i. Tap at 5 turns from each end and 1½ turns from one end (B&W No. 3004).
L_2—2 turns insulated hookup wire around center of L_1.
J_1, J_2, J_3—Tip jack.
J_4—BNC cable fitting. Connect J_4 and rotor of C_1 with copper strip.

band about 300 to 500 kc. wide by omitting this capacitive coupling, and using only the inductive coupling arising from the ⅜-inch spacing between L_9 and L_{10}. In some areas activity is concentrated below 50.4 Mc. or so, and the sharper response is no problem. It will improve the image rejection at the 500-kc. end of the broadcast set markedly, without an appreciable reduction in receiver gain or sensitivity, except in the upper part of the tuning range.

As with most receivers using bipolar transistors, it is important to use a properly tuned and matched antenna system, to avoid overloading problems from out-of-band signals. A well-matched 50-Mc. beam accomplishes the ordinarily, and something like our 50-Mc. portable job (See Chapter 9) is highly recommended. When a beam cannot be used, various "long wires" are effective, if properly tuned and matched to the transceiver input. Wire antennas and the little plastic-case antenna coupler of Fig. 7-14, will be found very superior to the collapsible whip type of antenna so often used with hand-carried equipment. Tilted wires respond to various polarizations, and they have some gain and directivity.

Various wire lengths can be plugged into the tip jacks connected to taps on L_1. A balanced line, or even an improvised V or rhombic, can be plugged into J_1 and J_2. Anything will work, but usually the longer the better. Tune in a signal on the receiver and peak the coupler for maximum signal strength.

The coupler can be connected directly to the BNC fitting on the transceiver, or a length of coax can be used. The support for the far end of the wire can be a fire tower, tree, building, or whatever happens to be handy. If there is room to maneuver, walk around (maypole fashion) until maximum signal is found. Contacts have been made at distances up to 125 miles on several occasions employing this haywire but effective approach.

A SELF-CONTAINED 2-METER STATION

A low-powered v.h.f. station complete in one package is a very desirable asset in many circumstances. It is fine for casual contacts at the home station. The man who travels can take it along easily for use in motel or hotel rooms, and it is mighty handy for portable work in contests, or for just a pleasant Sunday-afternoon outing.

The requirements for a general-purpose rig of this kind are quite different from those for the more highly-developed v.h.f. station. Usually a few watts of transmitter power will suffice, and the receiver need not be the ultimate in sensitivity and selectivity. If some compromise in these respects has to be made in exchange for easy portability, low cost, ease of construction and low power drain, it will usually be a good trade.

Perhaps the first decision to be made is the choice of power supply. Some builders may want operation from both 12 volts d.c. and 115 volts a.c. This is not difficult, nor very expensive, but a simpler approach may be the low-cost supply for a.c. input only. The latter was the choice of W1CER, builder of the package about to be described. Most of his operation was to be from a.c. lines. If field use from a car battery is wanted, the total drain of the station is well within the capabilities of inexpensive d.c.-to-a.c. inverters available for 12-volt use. One such unit tested with this rig is the Heathkit MP-10 Inverter, which makes 115 volts a.c. at up to 175 watts, continuous duty, with 12-volt input.

Transmitter power level and receiver performance are important relative considerations;

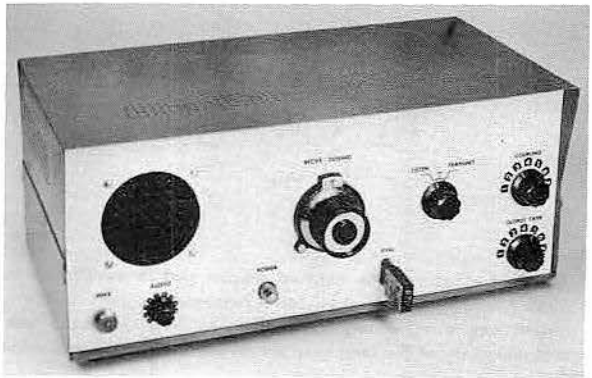


Fig. 7-15—A complete one-package station for 144-Mc. work. Only external connections required are for the antenna and microphone.

if you are satisfied with 2 or 3 watts transmitter output there is no real need for the ultimate in receiving capability. With the simplest practical transmitter and receiver designs the total power consumption, weight and overall cost can be kept fairly low. Going to a 10-watt transmitter means quite a jump in power supply requirements, and it calls for something better in the way of receiving ability.

All factors considered, there are major differences between the two power levels, yet from a practical point of view the results may not show it. The margin between 10 watts and 2 watts is only 7 db. This will not make or break a v.h.f. station, so the final choice can be made mainly on the basis of how much power supply you want to provide, and the effort and expense you want to incur in carrying the project to completion.

The objective here was a simple low-cost station, with enough power and receiver sensitivity to do interesting work on 144 Mc. Result: a station with 2 to 3 watts output, well-modulated, and a 2-tube receiver that does a pretty fair job. The audio system works on both transmitting and receiving. There are 7 inexpensive tubes in all.

Circuitry

The receiver is not unlike the simple but effective superregenerative job described in Chapter 4. A grounded-grid r.f. amplifier 6CW4, V_1, works into another 6CW4, V_2, as the detector. The main feature of the receiver is the high-Q strip-line detector tank circuit, L_5, which helps to give better selectivity and stability than is obtainable with coil-and-capacitor circuits at 144 Mc.

As may be seen from the schematic diagram, Fig. 7-17, the audio system serves as both receiver amplifier and modulator. A 12AU7 two-stage amplifier, V_3, and a 6C4 amplifier, V_4, drive a 6AQ5 amplifier-modulator, V_5.

The transmitter uses two triode-pentode 6CX8s. The pentode of V_6 is a crystal oscillator-tripler, using 8-Mc. crystals. This drives the triode portion as a tripler to 72 Mc. Then follows the triode of V_7, doubling to 144 Mc., driving the pentode, V_{7B}, as a straight-through amplifier. The circuit of the transmitter is more or less standard practice, except possibly for the neutralization. The "capacitor" C_{13} is merely a short insulated hookup wire connected to the grid pin, and bent adjacent to the plate lead to L_9.

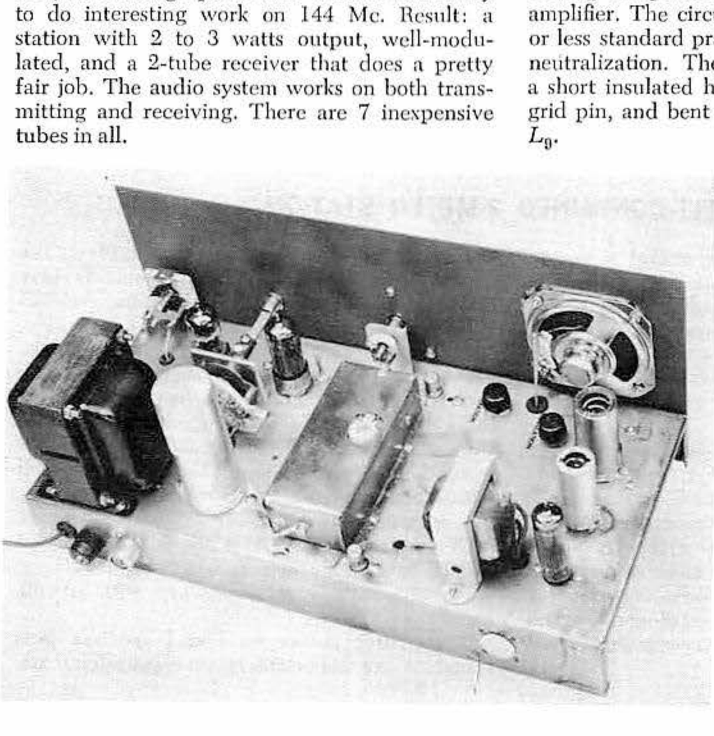


Fig. 7-16—Back-of-panel view of the 2-meter transceiver. The two transmitter tubes are at the upper left. The shielded assembly in the center is the detector tuned circuit. Audio tubes are at the right.

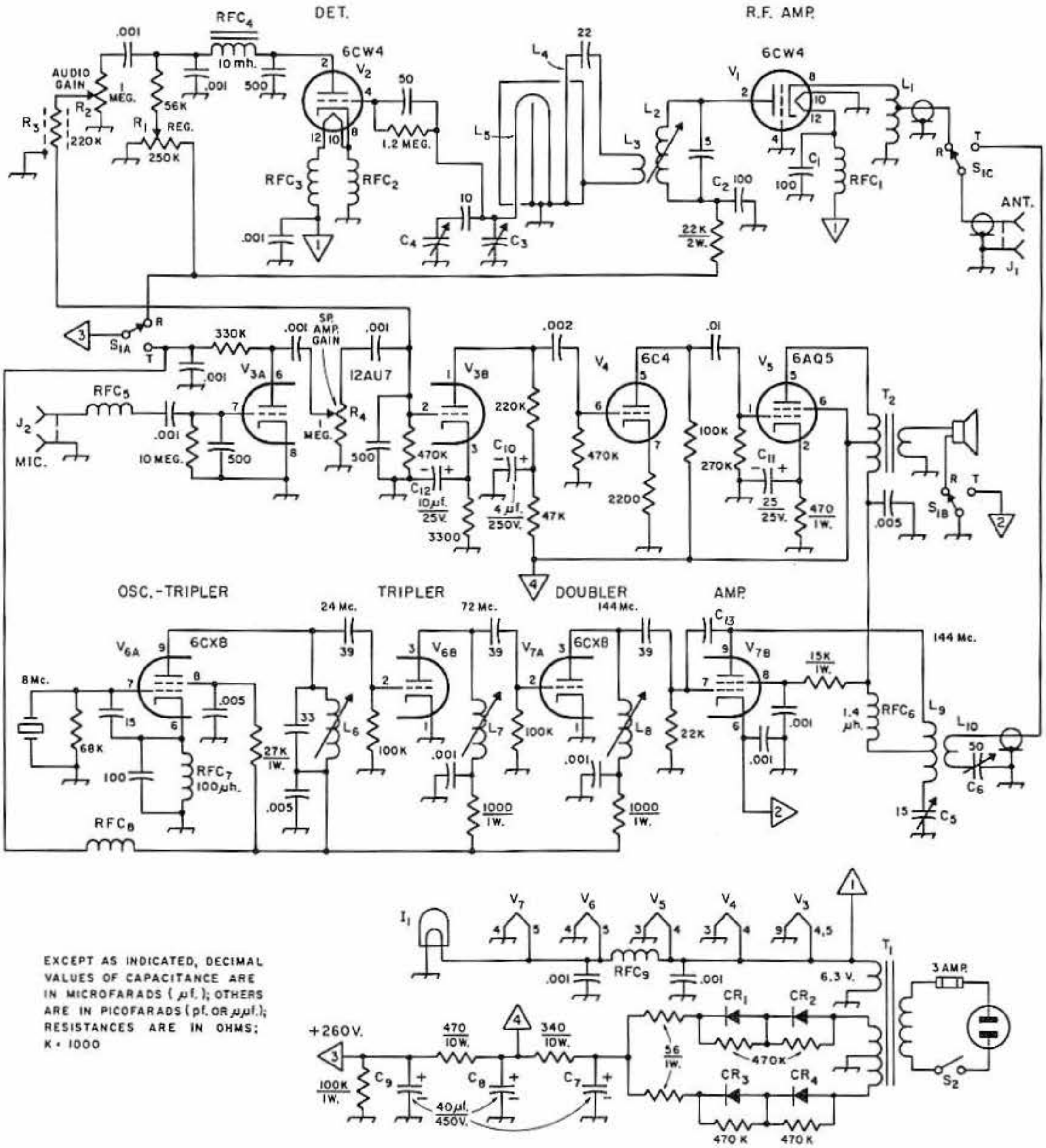


Fig. 7-17—Schematic diagram and parts information for the 2-meter station. Receiver circuits are at the top, audio portions in the middle, and transmitter and power supply below.

C_1, C_2—Series-resonant bypass, 100 pf., with ¼-inch leads. See Chapter 13 for details.

C_3—1.5 to 7-pf. ceramic trimmer.

C_4—Miniature shaft-type trimmer, 1 rotor and 2 stator plates (Johnson 160-102 or 5M11, modified).

C_5—15pf., double-spaced (Hammarlund HF-15X).

C_6—50-pf. single-spaced variable (Hammarlund HF-50).

C_7, C_8, C_9—Triple 40-uf. 450-volt electrolytic.

C_{10}—4-uf. 350-volt electrolytic.

C_{11}—25-µf. 25-volt electrolytic.

C_{12}—10-µf. 25-volt electrolytic.

C_{13}—Short insulated wire from Pin 9, adjacent to Pin 7.

CR_1, CR_2, CR_3, CR_4—400-volt p.i.v. 500-ma. diode.

J_1—Coaxial fitting, SO-239.

J_2—Microphone connector.

L_1—5 turns No. 20, 5/16-inch diam., ½-inch long, tapped 1½ turns from cathode end.

L_2—5 turns No. 22, ⅜-inch long on ¼-inch iron-slug form.

L_3—2 turns insulated hookup wire around B-plus end of L_2. Twist leads to detector assembly.

Continued on page 156

L_4—No. 16 wire inside detector assembly, ⅛ inch from L_5.
L_5—Copper strip ¼-inch wide and 9 inches long, bent into U shape. See Fig. 7-19.
L_6—12 turns No. 24 enam., closewound on 5/16-inch iron-slug form.
L_7—7 turns No. 20, ⅝ inch long, on 5/16-inch iron-slug form.
L_8—3 turns No. 20, ⅜ inch long, on ¼-inch iron-slug form.
L_9—3 turns each side of center, No. 12, ½-inch diam. Overall length 1¼ inches, with space of ¼ inch at center for L_{10}.
L_{10}—3 turns insulated hookup wire closewound, ⅜-inch diam. Insert about half diameter into space at center of L_9.
R_1—0.25-megohm control.
R_2, R_4—1-megohm control.
R_3—0.22-meg. ½-watt, encased in shield braid; see text.
RFC_1, RFC_2, RFC_3, RFC_5, RFC_6—1.4-uh r.f. choke.
RFC_4—10-mh. iron-core r.f. choke.
RFC_7—100-uh. r.f. choke.
RFC_8, RFC_9—10 turns insulated hookup wire, closewound, 5/16-inch diam.
S_1—2-position 3-section ceramic wafer switch, nonshorting.
S_2—S.p.s.t. toggle.
T_1—Power transformer, 520 volts c.t., 90 ma.: 6.3 volts, 4 amp. (Stancor PC-8420).
T_2—Push-pull pentode output transformer, 8 watts, voice coil secondary.

The power supply uses silicon diode rectifiers and an R-C filter, for light weight and low cost. The low current drain permits use of a moderately-priced power transformer. The power drain on *transmit* is somewhat over the rating for the transformer, but on receiving it is well below. Since the transmitting periods are normally a small percentage of the total running time there is no undue heating.

Construction

The transmitter, receiver and power supply are all built on a single chassis. The original is copper-plated steel, left over from a previous project, which explains unused holes that may be visible here and there. The metalwork is handmade, and of nonstandard size, but available components may be readily adapted to the job. Layout as shown is not critical, except as described hereafter, and more than the necessary amount of space was used simply because it was available. A 7 by 13-inch chassis, or perhaps one even smaller, should provide plenty of room.

Looking at the front panel, Fig. 7-15, the receiver tuning dial (a low-cost import with its knob replaced with a National HRT-M) is in the middle. At the left is the 3-inch speaker. The send-receive switch is just to the right of the dial, and at the far right are the transmitter output-loading and plate-tuning controls. Across the lower part of the panel, left to right, are the microphone jack, receiver audio gain control, pilot lamp and crystal socket.

In the interior view, Fig. 7-16, the transmitter portion is in the upper left, the two 6CX8s being mounted either side of the send-receive switch, S_1. The detector tank circuit, L_5, occupies the middle of the chassis. The detector 6CW4 is near the upper right corner of the tank circuit

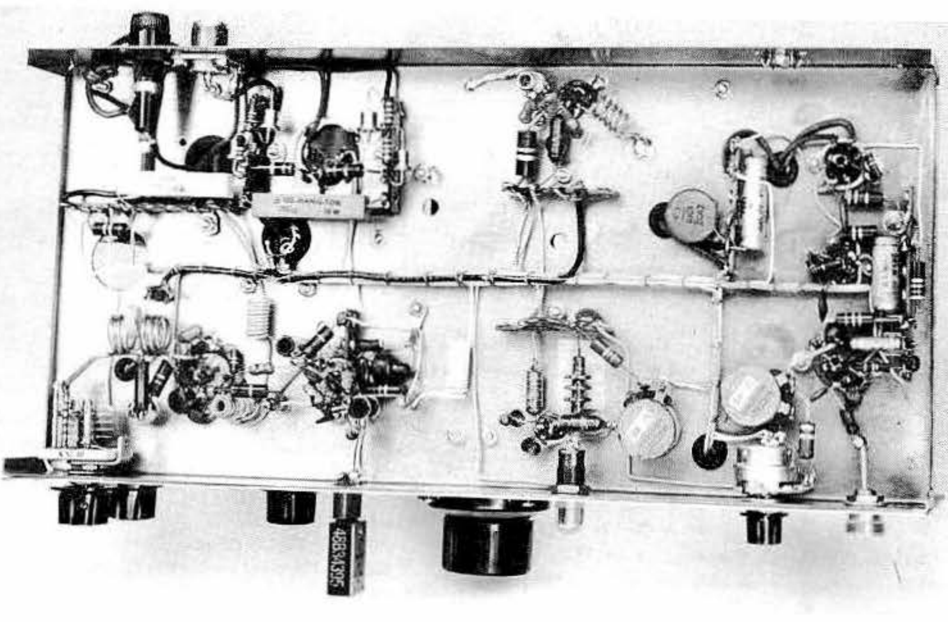


Fig. 7-18—Under-chassis view of the 2-meter transceiver. Power supply and transmitter components are at the left, receiver circuits near the center, and audio stages at the right.

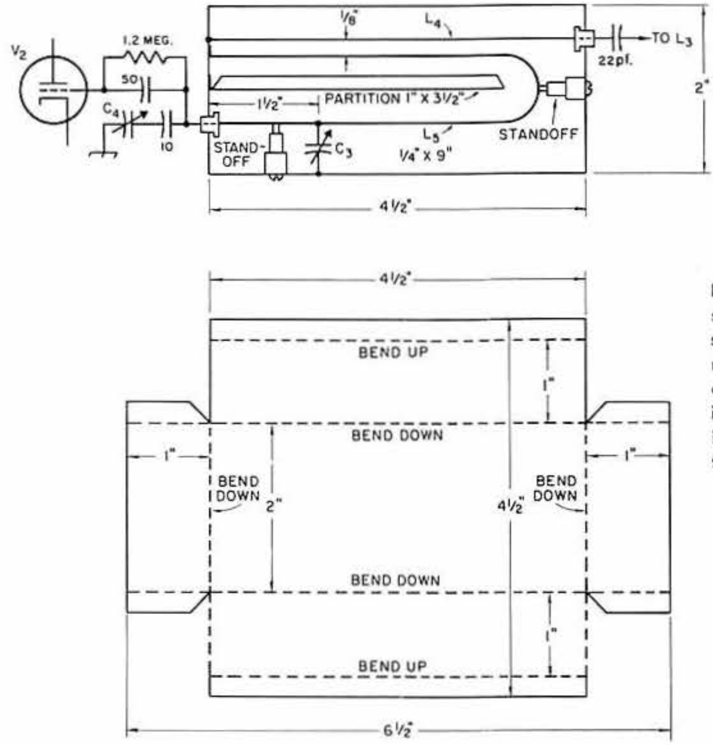


Fig. 7-19—Details of the shielded strip-line detector circuit. The assembly is shown in partial schematic form above, and dimensions of the copper sheet, before bending, are given below. The view is as it would appear looking into the assembly from the bottom.

enclosure, and the r.f. amplifier is below it, and to the right. The speech amplifier and modulator tubes are at the right end of the chassis.

In the bottom view, Fig. 7-18, the transmitter components are at the lower left, the power supply in the upper left, receiver circuits at the center, and audio stages at the far right. The speech amplifier gain and detector regeneration controls, R_4 and R_1 respectively, do not have to be reset in normal operation, so they are mounted on the chassis, in the lower right portion of this view.

Receiver Details

The shielded strip-line tank circuit, L_5, provides somewhat better selectivity and smoother detector operation than would a coil-and-capacitor circuit, because of its higher Q. The shielding also helps to keep down detector radiation. The assembly is shown in partial schematic form in Fig. 7-19, along with the principal dimensions of the copper plate from which the shield is bent.

As may be seen, a piece of copper or brass 6½ by 4½ inches is needed. This is cut and bent to form a 1 by 2 by 4½-inch box, with ¼-inch lips that bolt to the main chassis. A partition the full height of the box and 3½ inches long runs down the center. The end and lower edge are folded over, for grounding to the case. The detector inductance is a ¼-inch copper strip bent into U shape, grounded at the upper left end, and supported at two points with small standoffs. The ceramic padder, C_3, is mounted in the top of the compartment, as may be seen from the top view, Fig. 7-16. Connection to the hot end of the line is brought out through a small feed-through bushing to a 10-pf. capacitor and to the grid capacitor and grid leak. The tuning capacitor, C_4, is also connected at this point. The number of plates in C_4 determines the tuning range. As shown, with one rotor and two stator plates the coverage is just a bit more than the full 2-meter band. A piece of No. 16 wire, L_4, running the length of the box, ⅛ inch from L_5, provides coupling from the detector circuit to the plate coil of the r.f. amplifier.

The connection from L_4 to L_3 is made with twisted hookup wire, the 22-pf. series capacitor being just outside the detector circuit assembly. Miniature coaxial line could also be used, as it is for the connections from the receiver cathode input coil, L_1, and the transmitter output coil, L_{10}, to the send-receive switch. These two lines and parts of the power wiring are cabled down through the middle of the chassis. The heater and high-voltage lines to the two 6CX8s are decoupled by RFC_7 and RFC_8, which are merely a few turns of the leads in question, coiled up and laid against the chassis.

The r.f. amplifier stage gave some trouble with instability at first. This was traced to ineffective bypassing of the plate coil, L_2, and the heater circuit. The principal of series-resonant bypassing was applied to both points with excellent results. See Chapter 13 for more details.

The Transmitter

The transmitter circuits are conventional, and their design and operation have been covered thoroughly in other parts of this book. The crystal oscillator was described in Chapter 5, and other examples of it appear in Chapter 6. The output stage is a neutralized straight-through amplifier, using a series-tuned plate circuit. Other layouts might require different neutralization methods, but in this instance the small capacitance from grid to plate afforded by the small insulated wire, C_{13}, is simple and effective.

Operating conditions in the transmitter are as follows: Oscillator plate current, measured between RFC_8 and L_6—18 ma. Tripler plate current—10 ma. Doubler plate current—8 ma. Final grid current, measured between the 22,000-ohm grid resistor and ground—1.5 ma. Amplifier plate and screen current, measured between the lower end of T_2 and RFC_6—34 ma. Power output—about 2½ watts.

Results

The little rig was used for many contacts, mainly from the famous Selden Hill location in West Hartford. Using small antennas the distances worked were up to nearly 100 miles, and signal reports out to somewhat beyond the visual range have been excellent. In practical results the rig seems to be at least the equal of early models of the Communicator, both as to transmitted signal quality and strength, and receiving ability. In these categories it is well ahead of inexpensive kit transceivers for the 144-Mc. band.

As mentioned earlier, the setup was tested under mobile conditions using a Heathkit MP-10 Inverter. This resulted in some power hum on receiving, which was minimized by the shielding of resistor R_3, as shown in Fig. 7-17. The shielding is not needed for operation from a.c. lines. The power buzz had no effect on reception, and, in fact, was not noticeable except at very low audio levels. Modulation readability under low-signal conditions has been reported as adequate, and the receiving range seems to match up well with the transmitter capability.

A TWO-METER TRANSVERTER

This 6-tube package, built by Ernest P. Manly, W7LHL, and described by him in *QST* for September, 1963, will convert 144-Mc. signals down to 14 Mc., and a 14-Mc. transmitter signal up to 144 Mc. An s.s.b. transceiver for 20 will work beautifully with the transverter, or a communications receiver and a 20-meter transmitter may be used. The 20-meter band was picked for the conversion process because it was felt that there is more s.s.b. gear on 20 than on any other band that would be suitable for this purpose.

Most 20-meter transmitters and transceivers will have more than enough power to drive the transverter mixer. A way must be found to limit this power. Construction and use of a suitable step-type attenuator was described by W9ERU in *QST* for December 1959. Easily-built fixed attenuators were described by W4HJZ in May 1968, *QST*. Some transmitters have provision for taking off power at levels below that of their output amplifiers, and such a tap should be used, where available.

Output from the 6360 mixer is sufficient to drive a linear amplifier with a pair of 4X250Bs or similar tubes to several hundred watts input on s.s.b. or c.w.

The block diagram, Fig. 7-21, shows how the system works. The top line of the diagram is a typical 2-meter converter, with 6CW4 and 6AK5 r.f. stages ahead of a 6AK5 mixer. Mixer output is 14 Mc. or higher. In the middle is the oscillator-multiplier, a 6AN8 with its triode working as a crystal oscillator on 43 Mc. and the pentode tripling to 130 Mc. This is standard v.h.f. converter practice so far. The difference is that the same injection stages are used for transmitting. In the bottom line, the 130-Mc. energy is fed to a 6AK5 amplifier, and then to a 6360 transmitting mixer. The 14-Mc. signal from the 20-meter sideband rig is also fed to this mixer, and output is on 144 Mc., with the same characteristics as the 20-meter signal. Though the transverter idea is associated with s.s.b. in most amateurs' minds, it can be used with other modes as well.

Circuit Details

The first r.f. amplifier in the receiver side, V_1, Fig. 7-23, is a grounded-grid Nuvistor. Its cathode input impedance is matched by means of a quarter-wave section of 93-ohm coaxial line, L_{16}. An alternate method would be to tune the cathode coil, and tap the antenna line down on it. These matching systems give equal performance. Gain of the grounded-grid stage is about 10 db. The second stage, V_2, is a 6AK5 pentode, with a gain of 25 to 30 db. If cross-modulation is expected to be a problem, a gain control could be included readily in this stage. Both this and the 6AK5 mixer, V_3, follow conventional converter circuit practice.

The three parts of the transverter are separated by shields. This permits the injection to be adjusted to the desired level, and provides isolation, to keep down spurious responses. Output from the 6AN8, V_4, at 130 Mc., is link-coupled to the receiving mixer grid circuit, with light coupling at each end. Noise figure of the converter is under 4 db. Other tubes could be used equally well in the oscillator-multiplier

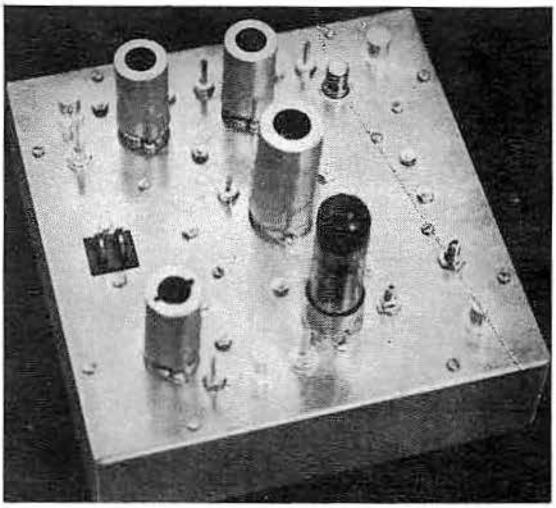


Fig. 7-20—This looks like a typical v.h.f. converter, but it is more than that: a 144-Mc. transverter which will handle both transmitting and receiving conversions from a 14-Mc. station setup for s.s.b. or other modes.

stages, and early versions of this transverter used a 12AT7 for this purpose. The triode-pentode gives somewhat more output, which is helpful in the transmitting side.

The injection stages run at low input, for good stability, so an amplifier stage is needed to build up the 130-Mc. energy for the transmitting mixer. This is done with the 6AK5 stage, V_5, which also helps to keep down the energy injected into the transmitting mixer at frequencies other than 130 Mc. Its output is link-coupled to the cathode of the 6360 mixer, V_6. Energy from the 14-Mc. transmitter is fed push-pull to the mixer grid circuit, which is tuned to 14 Mc. The mixer plate circuit is also push-pull, and tuned to 144 Mc.

A 22-volt zener diode, CR_1, is used in the mixer cathode circuit for bias, eliminating the need for an external bias supply. Operating conditions are designed to keep the tubes below their maximum dissipation rating if the crystal is removed or if it drops out of oscillation.

Construction and Adjustment

The transverter is built on a 7 × 7 × 0.032-inch brass plate. The shields are also made from 0.032-inch brass. Aluminum should work equally well. A 7 × 7 × 2-inch aluminum chassis is used to mount the transverter. Parts layout follows good v.h.f. practice, but is otherwise not particularly critical.

Tuning the oscillator, tripler and 130-Mc. amplifier is done by inserting a 50-ma. meter in the 6360 plate supply lead. The tube will draw approximately 1 ma. of plate current without drive, increasing as each stage is peaked. A grid-dip meter may be used to check each tuned circuit, to be sure that it is on the right frequency. Adjust the link between the 130-Mc. amplifier and 6360 cathode for maximum plate current, 10 to 20 ma. when 130-Mc. tuning is completed. A 14-Mc. carrier is fed in and the 6360 grid and plate circuits are tuned for maximum output at 144 Mc. The 14-Mc. carrier will drive the plate current to about 40 ma. before the 144-Mc. output starts to flatten out. Output at 144 Mc. will light a No. 47 pilot light to near full brightness.

There will be a little 130-Mc. and 116-Mc. energy appearing in the output, too. Some form of filter is desirable between the mixer output and the antenna or following linear-amplifier stages. A high-Q coaxial tank circuit with low-

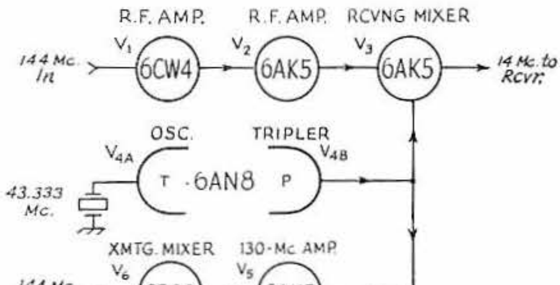


Fig. 7-21—Block diagram of the W7LHL 144-Mc. transverter. A single oscillator-multiplier section furnishes both transmitting and receiving injection, to separate mixers.

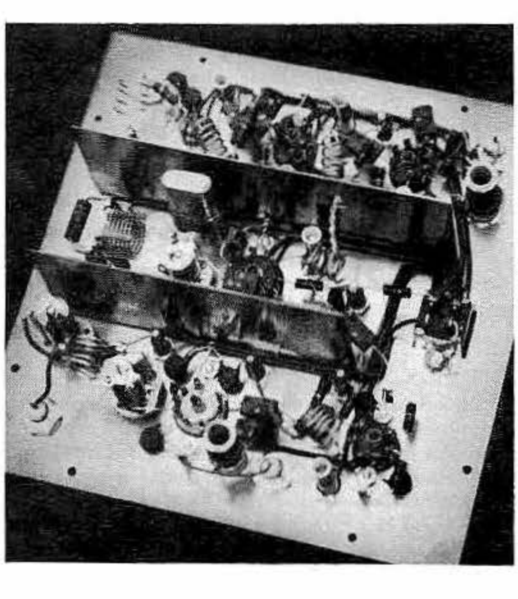


Fig. 7-22—Bottom view of the 2-meter transverter. Receiving stages are at the top, transmitting section at the bottom, and injection stages in the center section. Construction follows v.h.f. receiving converter practice.

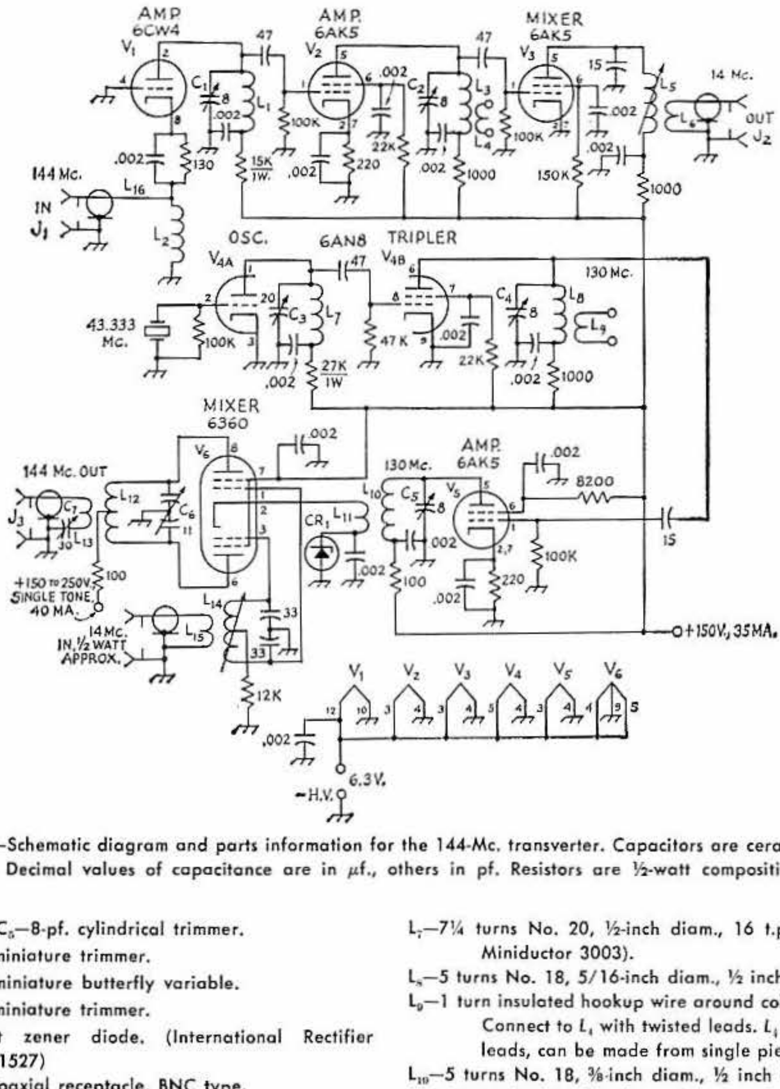


Fig. 7-23—Schematic diagram and parts information for the 144-Mc. transverter. Capacitors are ceramic unless specified. Decimal values of capacitance are in μf., others in pf. Resistors are ½-watt composition, unless specified.

C_1, C_2, C_4, C_5—8-pf. cylindrical trimmer.

C_3—20-pf. miniature trimmer.

C_6—11-pf. miniature butterfly variable.

C_7—30-pf. miniature trimmer.

CR_1—22-volt zener diode. (International Rectifier 1N1527)

J_1, J_2, J_3—Coaxial receptacle, BNC type.

L_1, L_3—5 turns No. 18, ¼-inch diam., ¾ inch long.

L_2—4 turns No. 18, ⅜-inch diam., ⅜ inch long.

L_4—1 turn insulated hookup wire, ¼-inch diam., at cold end of L_3. Connect to L_6 with twisted leads 2⅜ inch long.

L_5—35 turns No. 26 enam., close-wound on ⅜-inch diam. iron-slug form.

L_6—3 turns insulated hookup wire around cold end of L_5.

L_7—7¼ turns No. 20, ½-inch diam., 16 t.p.i. (B & W Miniductor 3003).

L_8—5 turns No. 18, 5/16-inch diam., ½ inch long.

L_9—1 turn insulated hookup wire around cold end of L_8. Connect to L_4 with twisted leads. L_4 and L_9, plus leads, can be made from single piece of wire.

L_{10}—5 turns No. 18, ⅜-inch diam., ½ inch long.

L_{11}—1 turn insulated hookup wire around cold end of L_{10}.

L_{12}—5 turns No. 16, ⅜-inch diam., ½ inch long, center-tapped.

L_{13}—1 turn insulated hookup wire at center of L_{12}.

L_{14}—33 turns No. 26 enam., close-wound on ⅜-inch diam. iron-slug form, center-tapped.

L_{15}—2 turns insulated hookup wire around center of L_{14}.

L_{16}—93-ohm coaxial matching section; RG-62/U, 16 inches long. See text.

impedance coupling in and out will serve this purpose well. Adjustment of coupling into and out of such a filter should be made for maximum attenuation of unwanted frequencies, rather than maximum transfer of 144-Mc. energy.

A Z-144 r.f. choke was used originally in the 6360 plate circuit in place of the 100-ohm resistor shown in Fig. 7-23, but this resulted in the mixer operating as a doubler from 14 to 28 Mc., because of choke resonance near the latter frequency. A Z-235 r.f. choke or the 100-ohm resistor will correct this tendency.

When using the circuit shown for matching with 93-ohm cable, Fig. 7-23, the input coil of the 6CW4 can be tuned to 144 Mc. by adjusting the turn spacing and using a grid-dip meter to indicate resonance. Disconnect the coaxial matching section when doing this. All other r.f. and mixer circuits can be peaked for maximum response in any portion of the 2-meter band.

Antennas and Feed Systems

Every radio station since the first one has had an antenna, but what a bewildering array these skywires represent! Antenna experimentation has long been a favorite pursuit for the amateur, and for good reason. Probably by no other effort can the average ham so improve his results, at so little cost, as by putting up a better antenna. But what is a better antenna? Though numerous books have been published on every aspect of antenna design, and talk about beams is heard wherever hams gather, there is probably more misinformation about antennas than about any other aspect of amateur radio.

In the following pages we will attempt to sort the wheat from the chaff, but the reader is cautioned not to look for simple all-purpose answers. There is no one antenna or feed system that "has everything." Nowhere are better antennas more needed than on the v.h.f. scene, but making an intelligent choice involves more than a perusal of performance figures found in antenna manufacturer's catalogs.

As in our equipment chapters, we will discuss principles first, presenting information that will be helpful whether the reader intends to build an antenna or buy one ready-made. This will be followed by practical examples, for the fellow who wants to build his own arrays.

OBJECTIVES

Choice of a v.h.f. antenna system begins with some decisions about the type of work we want to do best, since there is no antenna that does all things well. Does highest possible gain overrule other considerations? Is broad frequency response important? Can we live with a sharp beam pattern, or would something broader and less critical in aiming serve our needs better? How about omnidirectional coverage? Can we go all-out for size, or must there be some compromise with what simple mounts and inexpensive rotators will handle? What is the nature of nearby terrain? Are there trees, wires and buildings to be cleared? Is there a neighbor problem? Let's think about these and other points a bit before we get down to design details.

About Gain

Sad to relate though it may be, we should recognize the fact that antenna gain figures are often on the optimistic side. Even when given accurately, and with the best intentions, gain information may be in terms confusing to the average reader. True measurement of gain is difficult; few amateurs can do it accurately. On-the-air evaluation is also far from simple. It is entirely possible to get several contradictory observations in as many tries, and usually only long-term comparisons will show whether a new antenna is doing what we wanted it to do. After all, what really counts is whether or not it provides a stronger and more consistent signal at distant points than was possible before. If a new installation does not yield this result, impressive numbers have little meaning.

There is no magic about antenna gain. It is achieved only by taking radiation from some portions of the antenna's field where it may serve no useful purpose, and putting it into areas where it will do some good. A convenient way of expressing antenna gain, therefor, is to compare the field strength in the favored direction with that of an antenna that would radiate equally in all directions. Such an antenna exists in theory only, since it would have to be a point source. Called an *isotropic radiator*, it has a special appeal for the man who would have his antenna look good on paper. If it has a gain of 1, a half-wave dipole has a gain of 2.14 db. "Gain over isotropic" is a handy and legitimate way of stating antenna performance, and it is coming into more general use in antenna literature. But remember that figures so quoted are more than 2 db. higher than those for the more familiar half-wave dipole comparison. Unless otherwise stated, gains mentioned in this text are with respect to a half-wave dipole.

Frequency Response

Antenna gain is often achieved at the expense of frequency coverage, especially in arrays hav-

ing many parasitic elements. As will be shown later, the gain of even a small antenna of this kind is available over only a narrow frequency range, whereas an array of the same stated gain built with more driven elements may work well over all or most of a v.h.f. band. This point is of little concern to the 2-meter DX enthusiast who is interested only in the first few hundred kilocycles at the low end of the band, but to the fellow who wants to work effectively over a wide frequency range it could be the deciding factor in a choice between two very different antenna systems having the same advertised gain. The broad-band array will be larger and perhaps somewhat more difficult to handle and install, but with several times the bandwidth it may be a better investment, depending on one's operating habits.

Pattern Shape

An antenna with very high gain inevitably has a main lobe shaped like a cigar or a baseball bat. This can be both good and bad. In areas of high v.h.f. activity, a narrow antenna pattern is fine, if it helps to hold down the level of some of the local signals you have to work through. It may be helpful in nulling out man-made noise, if the sources of such noise lie mainly in one direction. On the other side of the ledger, a very narrow main lobe imposes stiff requirements on the rotator and direction-indicating devices, and it may keep you from hearing some choice DX that pops up a little off your line of fire. The local rag-chewer won't get his money's worth from really sharp beams.

Front-to-back ratio is allied to sharpness of pattern, but antennas can be built and adjusted to have high rejection off the back and still retain a broad frontal lobe. High front-to-back ratio is usually not obtained with the same antenna adjustment that gives maximum forward gain, however. It is another of those factors that you obtain in trade for something else. Usually the choice is between highest gain and optimum front-to-back ratio.

Omnidirectional Coverage

There are several ways of building up gain without losing a circular pattern in the horizontal plane. Compression of the vertical angle is involved here. Since radiation at angles other than those close to the horizon ordinarily does the v.h.f. man little good, trading off the high-angle power loss for gain along the ground looks almost like something for nothing. Omnidirectional antennas have many uses, and we'll be describing practical systems later on, but they, too, have their drawbacks. Gain achieved without modifying both horizontal and vertical patterns is limited to a few decibels, even when high and somewhat cumbersome stacks are involved. More important to the man in the midst of high v.h.f. activity concentration, interference problems multiply as omnidirectional gain is built up. Noise problems increase, and heterodynes and cross-talk seem to grow as if by magic.

Height Gain vs. Line Loss

In nearly every instance, the higher the v.h.f. antenna the better. Clearing obstructions in the immediate vicinity is of utmost importance. Wires, trees and buildings in the line of fire can ruin the antenna pattern, absorb power, and aggravate TVI and other interference problems. Putting the v.h.f. antenna up high enough so that its main radiation pattern is completely above nearby TV antennas may be one of the best TVI-prevention measures that can be taken.

As may be seen from the height-gain information of Chapter 2, increased height may do as much for you as putting up a much larger antenna at rooftop height. But how you get the added height is important. Particularly at 144 Mc. and higher, the added transmission-line loss may be considerable. An antenna installed at the top of a hill nearby may be a useful approach, but it probably will entail some special attention to feed methods.

And so it goes. Working with antennas is always interesting and often rewarding, but the important thing is to remember that there is more to it than a choice that promises the most decibels for the least dollars.

TYPES OF V.H.F. ANTENNAS

The simplest antenna commonly used in v.h.f. work consists of a single driven element. It may be called a dipole, a whip, a halo or some more fancy name. It may be horizontal, vertical, or something in between. It always has some gain over that theoretical isotropic antenna we spoke of earlier, but never very much. It is handy for getting on the air quickly and unobtrusively, but it will never give you a very large sphere of influence, unless perchance you live on the top of the highest mountain in your state—and maybe not even then. We'll go into dipole design later.

To build up antenna gain we do various things with dipoles. We may make several, hang them in a curtain arrangement, and feed them all in phase. This is called a *collinear* array. We'll have many practical examples in Chapter 9. Or we may line up one or more elements in the same plane with the dipole, in front or in back of it, but not connected to the feedline. These are called *parasitic* elements. They may be *directors* (one or more, shorter than the driven element, and placed ahead of it) or reflectors (usually only one, longer) placed in back. Such an antenna is called a Yagi, in honor of its Japanese co-developer.

Both systems have their uses in v.h f. communication. In either, gain is related to size. To double the gain of a collinear array (increase it by 3 db.) we must use twice as many elements, in a frontal area roughly twice as large. The size-gain relationship in a Yagi is more complex, involving length as well as number of elements, but a significant improvement always means a much larger antenna. To double the Yagi gain we must nearly double the number of elements, and more than double the length. Approximately 3 db. gain can be achieved by stacking two similar antennas side by side or one above the other, and feeding them in phase. The spacing for optimum gain increases with the length of the Yagi antenna. Pairs of antennas can be stacked, as can be pairs of pairs, and so on, but the size-doubling requirement for increasing the gain by 3 db. makes it obvious that increasing-gain projects reach the point of diminishing returns quite quickly.

YAGI ANTENNA DESIGN

Though it is possible to make a 2-element parasitic array bidirectional by adjustment of its element lengths or spacings, the objective in making a Yagi is usually a pattern that is essentially unidirectional. Whether the parasitic element operates as a reflector or a director is determined by the relative phase of the currents in the driven and parasitic elements. With element spacings commonly used (¼ wavelength or less) the current in the parasitic element will be the right phase to make it act as a reflector when it is tuned to the inductive or low-frequency side of resonance. In other words, it will be longer than the driven element. To act as a director, the element must be tuned to the high-frequency side of resonance, by making it somewhat shorter than the driven element.

Element Lengths and Spacings

The maximum gain that is theoretically possible with a 2-element parasitic array is shown in Fig. 8-1. The gain obtainable at various director and reflector spacings is also given. It is assumed that the length of the parasitic element is adjusted for each change of spacing, as it must be if it is to deliver maximum gain. The curves of Fig. 8-1 are for 2-element antennas only, and results in practice may not work out exactly as shown. To see why we look at Fig. 8-2, which shows the radiation resistance at the center of the driven element when the parasitic element is adjusted for the conditions of Fig. 8-1. Note that the radiation resistance goes very low at close element spacings.

With the low feed impedance of the close-spaced array, the r.f. current is very high, so ohmic losses go up. The bandwidth of the system goes down, and the difficulty of feeding the antenna properly rises. The result is that, for practical purposes, and especially with more than one parasitic element, some modification of these theoretically optimum spacings is necessary for best overall results. Thus we find the "wide-spaced" array in common use where three or more elements are involved.

Exact analysis of what happens in a Yagi antenna is difficult if not impossible, and it would serve little use here since optimum arrangements have been worked out experimentally by many workers. There are many ways to make a good Yagi, particularly a long one. Numerous combinations of element lengths and spacings give almost identical results. You can have an interesting time of it proving this for yourself, if you are experimentally inclined. If you aren't, you can make a v.h.f. array by following the tables given herewith, and be assured of good results, provided you have a few simple instruments to check system performance.

In practical terms, the element lengths and spacings in a Yagi array, even a long one, are by no means so critical as some harried workers would have you believe. If you can match the

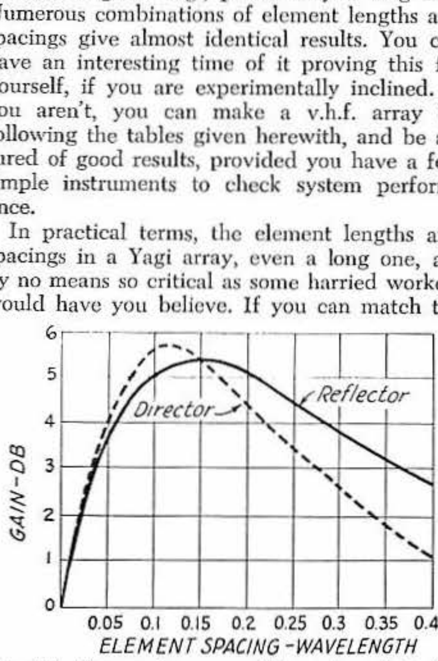


Fig. 8-1—The maximum possible gain obtainable with a parasitic element over a half-wave antenna alone, assuming that the parasitic element tuning is adjusted for greatest gain at each spacing.

antenna to its transmission line and transfer power to it efficiently, it will perform well, even if element lengths vary by as much as 1 percent and element spacings by 5 percent from the recommended values. The reflector length is more tolerant on the long side and the director lengths on the short side of the table values. A small error in these directions increases the antenna's bandwidth with only a very minor effect on gain. Error in the opposite directions can ruin Yagi performance much more quickly. This is the same as saying that frequency response of a Yagi is broader on the low side than on the high side of resonance, a fact that it is well to bear in mind in deciding on an optimum frequency for the array.

Recommended dimensions given later take this into account, but remember, if you want more bandwidth make the reflector longer and the directors shorter. If for some reason you want to try close parasitic-element spacing, make the director nearest the driven element *longer* and the reflector *shorter* than the recommended values, by 1 to 2 percent, depending on the closeness of the spacing. This is in line with the point made above that frequency response sharpens with closer spacing.

There is little point in pursuing the matter further, as these points are covered in interesting detail in all modern editions of the *A.R.R.L. Antenna Book*. A study of that manual, and some of the references it cites, is recommended to the experimentally-inclined amateur.

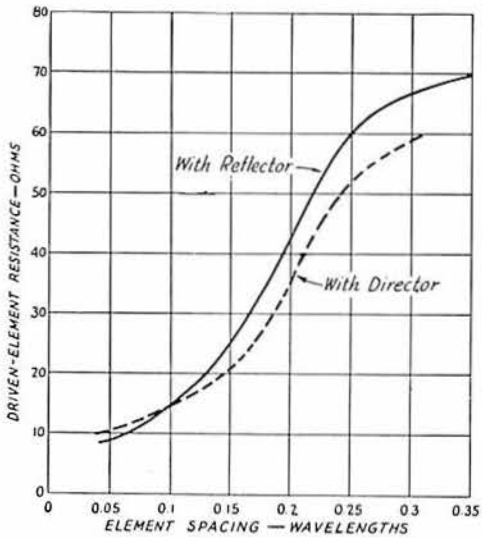


Fig. 8-2—Radiation resistance at center of driven element as a function of element spacing, when the parasitic element is adjusted for the gains given in Fig. 8-1.

Yagis—Short and Long

Element spacing of 0.15 to 0.2 wavelength is recommended for small Yagis (up to 4 elements) commonly used in 50-Mc. work. With convenient tubing and boom sizes the following figures apply for all v.h.f. parasitic arrays:

$$\textit{Driven Element Length, Inches} = \frac{5600}{\textit{Freq. (Mc.)}} \qquad (1)$$

Reflector—5 percent longer
Director—5 percent shorter
Second Director—6 percent shorter

A perceptible increase in gain can be had by spacing the second director 0.25 wavelength from the first, though the difference may not be worth the extra boom length in some instances.

Uniform element spacing can be continued for more elements, but some improvement in gain can be achieved by going to graduated spacing. Here each successive director is spaced somewhat greater than the previous one, until a spacing of about 0.4 wavelength is reached, after which all directors are spaced this much. A good rule of thumb for medium and long Yagis is:

Dipole to D_1—0.15 wavelength (λ)
D_1 *to* D_2—0.18 λ
D_2 *to* D_3—0.25 λ
D_3 *to* D_4—0.35 to 0.42. All directors beyond D_4 to be the same spacing.
Dipole to Reflector—0.2 (2)

The 5-percent-longer, 5-percent-shorter rule applies. For slightly more bandwidth, start with D1 at 4 percent shorter than the driven element and take off 0.5 percent for each succeeding director. Lengths of individual elements are sufficiently tolerant that you can change by 0.5 percent one way or the other and find only a barely measurable difference in gain. A listener at the other end of a communications circuit would not know that anything had happened. Even at 432 Mc., where half-wave elements are only about a foot long, it is possible to change *all* elements by as much as 1/16 inch without much effect, if the matching system is readjusted after the change is made.

Element Diameter

Information given in (1) applies with tubing sizes commonly used in v.h.f. antennas. Elements ½ to 1 inch in diameter are customarily used at 50 Mc. For 144 Mc. ⅛ to ⅜ inch is common. For 220 and 420 Mc. we may go as small as 1/16 inch, and anything larger than ¼ inch is seldom used. There may be exceptions however, and appreciable changes in element lengths must be made for large variations from the above practice.

The larger the element diameter the broader the frequency response, so variations on the large side may not be too critical, but when elements much smaller than standard practice are involved it is well to check the antenna performance carefully. Some idea of the practical extent to which small element diameter can

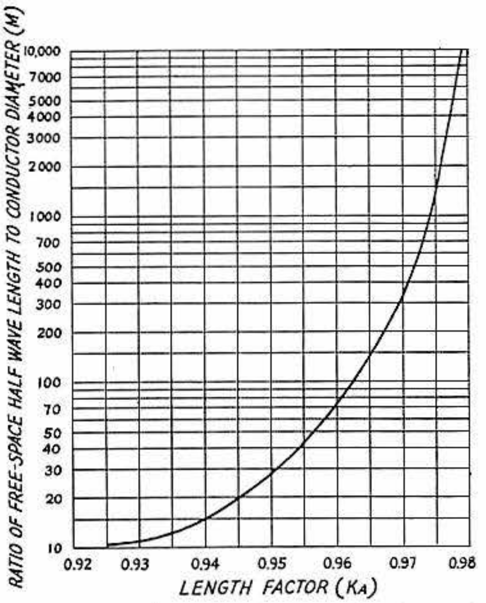


Fig. 8-3—Length factor for the range of conductor diameters used in practice. This curve applies to either quarter-wave (grounded or ground-plane antennas) or half-wave antennas.

affect element length can be seen in a 50-Mc. portable beam described in Chapter 9. In planning construction of antennas having unusual length/diameter ratios the information of Fig. 8-3 can be used to good advantage. Here the vertical scale is the free-space half wavelength divided by the conductor diameter. The horizontal is the percentage of a free-space half wavelength that should be used for a driven element.

The standard-practice range of element sizes mentioned above represent only about 1 percent change in element length, from lowest to highest *K* factor. The 50-Mc. portable array mentioned above has a factor of nearly 2000, if the diameter near the ends of the telescoping elements is used. The elements in this array had to be extended 6 inches beyond the usual lengths before this array worked properly. This is an increase even greater than that indicated by the graph of Fig. 8-3.

Gain and Size in Yagi Arrays

As mentioned earlier, there is an optimum boom length for a given number of elements in a Yagi array. This is obtained from Fig. 8-4, curve A. Note that a 4-element array is about a half wavelength long. A 6-element beam of optimum length is *twice* as long, and an 8-element array should be *four times* as long as one with 4 elements.

What these combinations should yield in gain is given by curve B. Our 4-element Yagi is capable of just under 9 db. The 6-element goes over 10, and the 8-element to about 12. There is no limit to the gain that can be achieved with ever-longer booms, as has been demonstrated many times by experimenters willing to build Yagis 50 feet or more in length for 144 Mc., but after the first 13 to 15 elements it becomes a rather dubious business.

Optimum element placement is given in Table 8-I. As may be seen from these figures, this is not particularly critical. These graphs and table and the information of (1) are all one needs to design effective Yagi arrays.

Collinear Arrays

The collinear (elements along a common line) is one of the oldest forms of directional arrays. The "Two Half-Waves in Phase," the "Extended Double-Zepp," and other simple collinears are almost as old as short-wave radio. They and their larger multielement relatives that grow on some of our best v.h.f. antenna farms are still among the most useful. Because the collinear is made up of many driven elements, with only reflectors for parasitic elements, it is much more frequency-tolerant than a Yagi of the same gain. A 2-meter collinear can

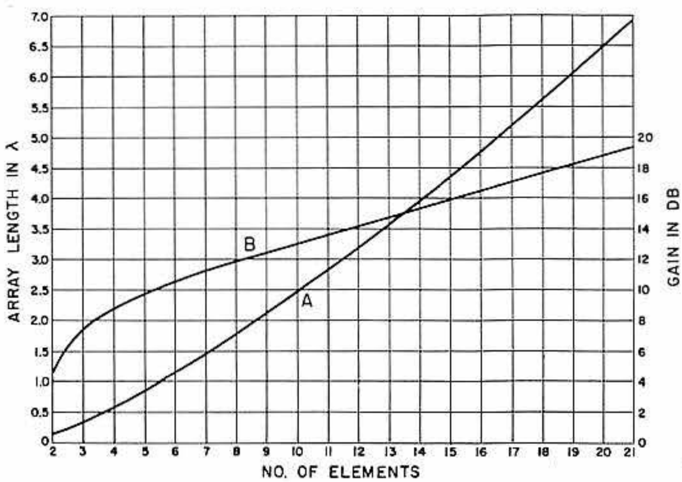


Fig. 8-4—Design information for Yagi antennas. Curve A shows the optimum boom length in wavelengths for any number of elements. Curve B shows the maximum gain that can be expected when the design information of Curve A is used.

TABLE 8-I

Optimum Element Spacings for Multielement Yagi Arrays DE—driven element; R—reflector; D—director.

No. Elements	R-DE	DE-D_1	D_1-D_2	D_2-D_3	D_3-D_4	D_4-D_5	D_5-D_6
2	0.15λ-0.2λ						
2		0.07λ-0.11λ					
3	0.16 -0.23	0.16 -0.19					
4	0.18 -0.22	0.13 -0.17	0.14λ-0.18λ				
5	0.18 -0.22	0.14 -0.17		0.17λ-0.23λ			
6	0.16 -0.20	0.14 -0.17	0.16 -0.25	0.22 -0.30	0.25λ-0.32λ		
8	0.16 -0.20	0.14 -0.16	0.18 -0.25	0.25 -0.35	0.27 -0.32	0.27λ-0.33λ	0.30λ-0.40λ
8 to N	0.16 -0.20	0.14 -0.16	0.18 -0.25	0.25 -0.35	0.27 -0.32	0.27 -0.33	0.35 -0.42

N = any number; director spacings beyond D_6 should be 0.35-0.42λ.

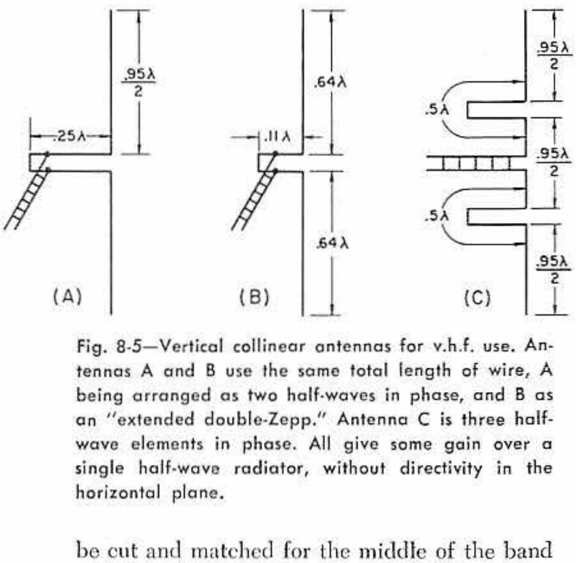


Fig. 8-5—Vertical collinear antennas for v.h.f. use. Antennas A and B use the same total length of wire, A being arranged as two half-waves in phase, and B as an "extended double-Zepp." Antenna C is three half-wave elements in phase. All give some gain over a single half-wave radiator, without directivity in the horizontal plane.

be cut and matched for the middle of the band and it will work over the entire four megacycles with only moderate variation in gain.

Properly designed, a collinear system is easy to feed with common types of transmission line. It can be strung together with sticks and wire, hauled up into a tree, and rotated by pulling on ropes—and it will work. One of the first and most renowned v.h.f. beams ever built was handled in just that way. See Chapter 1. This is not to infer that the collinear is useful mainly where ham haywire is the order of the day. Care in design and construction pays off in performance, but the collinear is tolerant of amateur methods; considerable more so than the Yagi.

Nondirectional Collinears

Simple vertical collinears like those of Fig. 8-5 provide some gain over a vertical half-wave dipole, without introducing directivity in the horizontal plane. At *A* we have two half-waves in phase, fed by means of a folded half wavelength at the center. A balanced transmission line is shown, but coax and a balun could be used equally well. Slightly increased gain and lowered radiation angle result from lengthening the radiating portions and shortening the stub, as in the extended double-Zepp, *B*. The total wire length is the same: 3 half-wavelengths. Three or more half-wave vertical elements, kept in phase by means of quarter-wave stubs, C, is another common omnidirectional vertical antenna.

Directional Arrays

Larger arrays with 4, 6, 8 or more half-wave elements stacked side-by-side and one above the other are what most v.h.f. men think of when they hear the term, "collinear." These may be driven elements only, as in Fig. 8-6, wherein a bidirectional pattern is obtained, or reflectors may be added for unidirectional characteristics, as in several examples shown in Chapter 9. Directors can be added, but this is seldom done. Large arrays with directors are better arranged and fed as combinations of Yagi bays, rather than as collinears.

Reflectors in a collinear array are usually parasitic in nature, but a reflecting metal plane can be used. This can be of sheet metal, though more often wire mesh is used, in the interest of decreased wind resistance. Spines of wire or small tubing are also usable, so long as

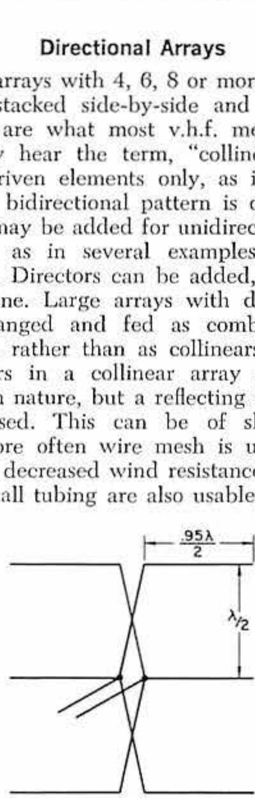


Fig. 8-6—Bidirectional collinear array using 6 half-wave elements in phase. Parasitic reflectors or a screen may be placed in back of the driven elements for increased gain and unidirectional pattern.

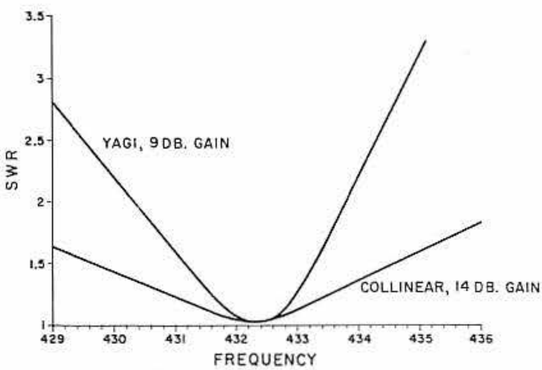


Fig. 8-7—Comparison of the frequency responses of a small Yagi antenna and a large collinear array. A Yagi of comparable gain would have a still sharper frequency response.

the spacing of the spines is well under 0.1 wavelength. To be fully effective, the plane reflector should be at least a quarter wavelength larger in both dimensions than the curtain of driven elements it backs up.

An interesting comparison between the bandwidth of a 6-element Yagi and that of a collinear array having 8 half-wave elements in phase, backed by a screen reflector is shown in Fig. 8-7. The Yagi, with a gain of about 9 db., has a much sharper frequency response than the collinear, with a gain of 14 db. Both antennas were matched carefully between 432 and 433 Mc. The collinear shows an s.w.r. under 1.8 over a range of 7 megacycles, while the Yagi exceeds this mismatch in less than 3 megacycles. Had the Yagi in question been a long one, with a gain similar to that of the collinear, its useful frequency range would have been very much sharper still. A typical 2-meter long Yagi may be expected to work well over about 1 megacycle, while a collinear of large size for this band will work nicely over at least 3 times as much frequency range.

POLARIZATION

The wave emitted by an antenna perpendicular to the earth is said to be "vertically polarized." Radiation from an antenna parallel to the earth's surface is termed "horizontal." In the space age these terms may mean nothing. Once we lose the reference of ground there is no longer any "horizontal" or "vertical," but merely what is more accurately called *plane polarization.* The radiation from any straight wire or rod is mainly plane-polarized, but it can be horizontal, vertical, or anything in between.

Much of the time it *is* something in between, so the horizontal-vertical argument that raged for years without ever being entirely settled tends to be a specious one. There is no one "best" polarization, and going along with what others in a given area are using offers the best hope for good v.h.f. coverage. Because a vertical dipole or whip has an essentially omnidirectional pattern in a horizontal plane, vertical antennas were employed for most of the early v.h.f. communication, before the days of high-gain arrays. When beams began to take over the burden of v.h.f. work it was only natural to mount them in a horizontal position. Gradually then v.h.f. men went over to horizontal antennas, except in a few areas where mobile work was a major factor. Here, verticals remained popular, for obvious reasons of simplicity and appearance.

Horizontal or Vertical?

There is no consistent large difference in coverage between horizontal and vertical, so long as the same polarization is used at both ends of the path. Reflections and the passing of the wave over intervening hills modify polarization to a marked extent. Probing with mobile antenna installations will show the polarization shifting with a car movement of a wavelength or less. Discrimination between horizontal and vertical may amount to 20 db. or more, and at 144 Mc. or higher it can be found to reverse itself at times in a matter of inches of travel. The results of this are familiar to any v.h.f. mobile worker, in the form of "mobile flutter" that is so pronounced at certain car speeds, in anything but the most wide-open terrain.

In v.h.f. mobile communication our effective working range is nearly always limited by noise, mostly ignition racket from our own car and others. Such noise tends to be vertically polarized, so in areas of appreciable motor traffic horizontal antennas yield considerably higher signal-to-noise ratios than vertical ones. This has led to adoption of halos, turnstiles and other horizontally-polarized mobile antennas, despite the concern of some family passengers who may not be sold on the esthetic virtues of these devices.

In other v.h.f. work not involving mobiles noise is still a factor, but it may or may not be predominantly vertical in nature. In general polarization is not an important consideration, as far as signal-to-noise ratio is concerned, other than with mobiles. Long experience has shown that if there is any signal-strength advantage it usually lies with horizontal polarization. This is probably because of a combination of the vertical nature of some noise and the observed tendency of polarization to roll over to horizontal in passing over hilly terrain. This is hard to pin down, however, and some v.h.f. men with extensive experience in high-mountain country insist that vertical is superior to horizontal in working with mobiles. This may

well account for the predominance of verticals in California and preference for horizontals in most other areas, where the terrain is either open or rolling in character.

Near saturation of the country with television, which employs horizontal polarization, introduces a factor not present when the move to horizontal standardization for amateur v.h.f. began in the late 1930's. Because polarization shift is slight in the immediate vicinity of the transmitting station, horizontal polarization for both home TV and amateur work does increase the possibility of TVI of the front-end-overload variety. It should not be assumed that changing back to vertical would be a cure-all for TVI problems however, for the causes and cures of TVI are much more involved than this. Furthermore, since interference that may result from matched polarization is due to receiver deficiencies and is not the fault of the amateur, he should not be required to sacrifice communications efficiency as a TVI expedient.

We had interference problems when everyone used vertical v.h.f. antennas. We still have them in areas where verticals are predominant today. TVI resulting from harmonic radiation and the all-too-common audio-rectification problem would be largely unaffected by changing the polarization of the amateur antenna. Cross-polarization has demonstrated no marked TVI cure-all properties in Great Britain, where television is vertical and amateur v.h.f. antennas are horizontal.

Space-Age Polarization

A third type of polarization is coming into widespread use in communication involving natural and artificial satellites, because of the constantly varying polarization encountered in this work. Called helical or circular polarization, this is best symbolized by a screw thread, with the wave boring through space in the manner of a bolt being turned into a threaded hole. A circularly-polarized system will accept waves of any plane polarization, as well as circularly-polarized waves, so it is useful in amateur v.h.f. communication in areas where both horizontal and vertical are in use. It suffers a 3-db. penalty for its universality, however, so it is usually not as good as matched plane polarization in such circumstances.

Two-way work involving circular polarization at both stations should be equal or superior to matched plane polarization, and it may be used on paths where there is marked polarization shift, over land as well as in space communication. There is a polarization-matching problem with circular systems also: the direction of rotation, or "sense," must be the same for both stations. A right-hand-polarization wave encounters approximately the same barrier in a left-hand-polarized antenna as does a horizontal wave at a vertical antenna, the discrimination amounting to 20 db. or more.

A further complication is introduced in work via reflection paths, as in the earth-moon-earth route: the reflection produces a reversal of polarization sense. To receive our own signal reflected from the moon, we must reverse the sense of antenna polarization between transmitting and receiving. The problem is lessened for two stations communicating by way of the moon or a reflecting satellite, since one merely needs to use right-hand and the other left.

Circular polarization is inherent in the helical antenna, Fig. 8-8, in which the driven element is a coil of wire or tubing, fed at one end and usually backed up by a screen reflector or "ground plane." Dimensions of the helix are not critical, so it is useful over a very wide frequency range. Each turn of the helix is one wavelength at the midpoint of its useful frequency range. Combinations of horizontal and vertical elements placed at right angles to one another and fed in the proper phase also produce circular polarization. The sense can be reversed by reversing the feed system in such an array.

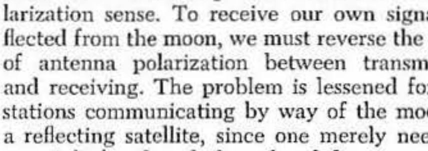


Fig. 8-8—Schematic drawing of a helical antenna. Circumference should be one wavelength, diameter 0.32 wavelength, turn spacing 0.22 wavelength, and reflector diameter 0.8 wavelength.

OFF THE BEATEN PATH

Collinear and Yagi antennas of conventional design are so universally used in v.h.f. communication that most amateurs give little consideration to other systems. To a degree, antenna principles are the same regardless of frequency, so it may be to the v.h.f. man's advantage to try methods used on other frequencies, both higher and lower than his accustomed stamping ground.

The Quad

An antenna very popular among DX men on 10, 15 and 20, but little used in v.h.f. work, the Quad has interesting possibilities. It can be built from sticks and wire, if need be, so its cost can be close to the ultimate low for beams. It is an antenna that is readily adjusted, since all elements can be stub-tuned. It has an appreciable frontal area, and it is inherently a lower-angle radiator than a Yagi of the same height above ground. A quad can be built with any number of elements, in the manner of a Yagi, or several driven elements can be arranged in sets fed in phase, much like the collinear. The same basic stacking and feeding principles apply for quads as for Yagis.

Long-Wire Antennas

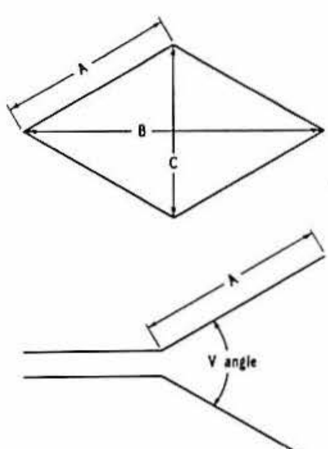
An amateur with a wire antenna for lower bands never need wait for the erection of a v.h.f. beam before operating on 50 Mc. or higher bands. Antennas for 80, 40 or 20 are not often well-adapted to v.h.f. use, but they can always be made to work, after a fashion. On rare occasions they may be outstanding, for a long wire operated on its high harmonics has interesting properties. Two-meter men the length of the Atlantic Seaboard fondly recall the booming signal put out for years by W4-CLY, using a 75-meter dipole that sloped down from the lighthouse at Cape Henry, Virginia. More often, the low-band wire will show its best properties in the least-useful directions, but there's no harm in trying.

Principles that make the V and rhombic useful on lower frequencies still apply at v.h.f. If designed for lower bands these antennas will not have dimensions that are optimum for 50 or 144 Mc., but they can be pressed into service in a pinch. With side lengths and angles adjusted for a v.h.f band they may do very well. A rhombic large enough for appreciable gain at 50 Mc. may fit on a residential lot, and if it can be aimed to take care of major activity areas it may be worth a try. Unterminated, the rhombic is bidirectional, which may help in this. A main problem with wire arrays is getting them high enough to make them really pay off in v.h.f. work. Practical V and rhombic dimensions are given in Table 8-II.

TABLE 8-II

Dimensions of V and Rhombic Antennas for V.H.F. Use. Columns 1 and 2 Are for V Designs. For Rhombics Use 1, 3 and 4.

Freq. (Mc.)	*Side Length "A" in Feet*	*V Angle*	*Over-all Length "B" in Feet*	*Width "C" in Feet*
50.5	58	60°	96.5	65.5
145	58	35°	109	39
28.7	68	70°	101.5	84
50.5	68	55°	106.5	70.6
145	68	35°	129	41
50.5	106	42°	192.5	91.5
145	106	35°	205	47.5
28.7	136	52°	237.5	133
50.5	136	37°	252.5	102



Plane and Parabolic Reflectors

Looking higher in frequency, the v.h.f. man can borrow techniques from u.h.f. practice. Plane, corner and even parabolic reflectors begin to be attractive at the upper end of the v.h.f range. Large nonresonant reflector systems offer broad frequency response, clean pattern and noncritical adjustment, but from the standpoint of gain for a given size, they are not outstanding. A corner-reflector array having a gain of 10 db., for example, is larger and more difficult to erect than a Yagi or collinear of the same gain.

The flat-plane reflector backing up collinear elements may have more potential. A light frame covered with chicken wire, window screening or hardware cloth, with sets of elements for two bands on opposite sides of the structure is a convenient way of operating on 220 and 420, or 420 and 1215 Mc. with one rotating array. Except as it affects impedance, spacing of the corner or flat-plane reflector from its driven elements is not particularly critical. The impedance of the driven element for various spacings (D) from the vertex of corner angles 45 to 180 degrees (flat plane) is given in Fig. 8-9. Gain with the flat-plane reflector remains nearly constant from 0.1 to 0.25 wavelength, so it can be seen that varying the spacing may be a convenient way of accomplishing an impedance match.

The parabolic reflector produces a very sharp and clean pattern, if it is large in terms of wavelength. A reflector diameter of about 10 wavelengths is the minimum for appreciable focusing effect, which is the basis of the system. This means about a 25-foot "dish" for 432 Mc., which may look like the hard way to develop an outstanding signal at that frequency. Where the reflector can be set up at or near ground level, as for moonbounce work, a sizeable installation is well within the capacities of the kind of workers who are apt to band together for a group project in this field.

Because of constant improvement in reflector design for military and scientific needs, some large reflectors have become available to amateurs through surplus channels. Several of the larger amateur installations have used surplus dishes, but other individuals and groups have demonstrated that construction of a suitable parabola is not beyond the realm of possibility.

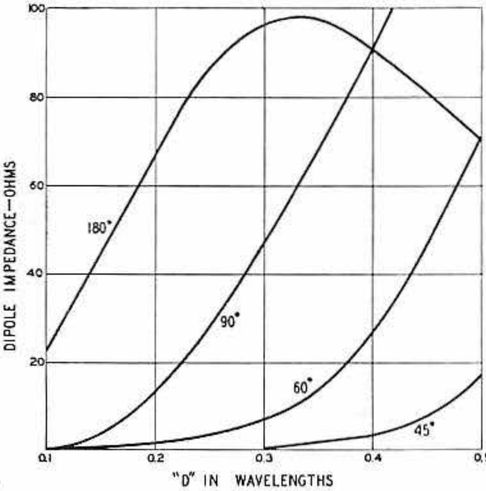


Fig. 8-9—Feed impedance of the driven element in a corner-reflector array, for various corner angles of 180 (flat sheet), 90, 60 and 45 degrees.

A reflector as small as 6 feet in diameter can be pressed into service at 420 Mc., but it will have large minor lobes and low gain. A Yagi or collinear array of equal or better performance is more practical. In general therefor the 1215-Mc. band is the dividing line above which the parabola becomes a thoroughly practical approach. For 2300 Mc. and higher it is practically standard equipment, and even a 4-footer works very well from this point on up.

STACKING PROBLEMS

In stacking horizontal Yagis one above the other on a single support, certain considerations apply whether the bays are for different bands or for the same band. As a rule of thumb, the minimum desirable spacing is one-half the boom length for two bays on the same band, or half the boom length of the higher-frequency array where two bands are involved.

In the stacked two-band array of Fig. 8-10, the 50-Mc. 4-element Yagi is going to "look like ground" to the 7-element 144-Mc. Yagi above it, if it has any effect at all. It is well known that the impedance of an antenna varies with height above ground, passing through the free-space value at a quarter wavelength and multiples thereof. At one-quarter wavelength and at the *odd* multiples thereof, ground also acts like a reflector, causing considerable radiation straight up. This effect is least at the half-wave points, where the impedance also passes through the free-space value. Preferably, then, the spacing S should be a *half* wavelength, or multiple thereof, at the frequency of the smaller antenna. The half-the-boom-length rule gives about the same answer in this example. For this length of 2-meter antenna, 40 inches would be the minimum desirable spacing, but 80 inches would be better.

The effect of spacing on the larger array is usually negligible. If spacing closer than half the boom length or a half wavelength must be used, the principal thing to watch for is variation in feed impedance of the smaller antenna. If the smaller antenna has an adjustable matching device, closer spacings can be used in a pinch, if the matching is adjusted for minimum s.w.r. Very close spacing and interlacing of elements should be avoided, unless the builder is prepared to go through an extensive program of adjustments of both element lengths and matching.

Stacking for Gain

In stacking bays for the same band fed in

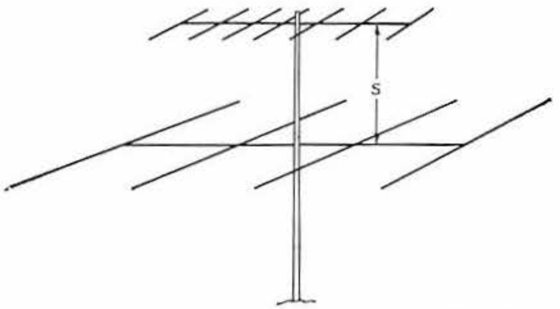


Fig. 8-10—In stacking Yagi arrays one above the other the minimum spacing between bays, S, should be about half the boom length of the smaller array. Wider spacing is desirable, in which case it should be a half wavelength, or some multiple thereof, at the frequency of the smaller array. If the beams shown are for 50 and 144 Mc., S should be 40 inches minimum, with 80 inches preferred. Similar conditions apply for stacking bays for a single band.

phase, the minimum spacing for appreciable gain is a half wavelength for Yagis of up to four elements or so. For such small Yagis, and for dipoles and omnidirectional systems such as the Big Wheel and the turnstile, a spacing of ⅝ wavelength will give appreciably more gain. This is convenient in that an electrical full wavelength of coax may be used for phasing. We'll get into phasing and feed problems later.

As bay spacing is increased in directional arrays the main lobe becomes sharper, but minor lobe content also increases. This becomes self-defeating if carried too far. Small Yagis spaced a half wavelength show a beautifully clean pattern, but only moderate gain from stacking. For Yagis up to two wavelengths long, a bay spacing of one wavelength is good, though minor lobes are quite pronounced when individual bays have 6 elements or less.

For arrays of more than two wavelengths, keep that half-the-boom-length minimum in mind, but space them wider if you can. It can be seen from this that stacking of long Yagis makes for large and ungainly structures, but gain never comes easily once you get into the upper brackets.

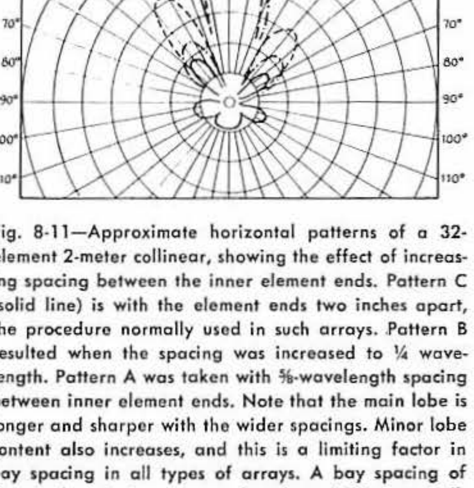


Fig. 8-11—Approximate horizontal patterns of a 32-element 2-meter collinear, showing the effect of increasing spacing between the inner element ends. Pattern C (solid line) is with the element ends two inches apart, the procedure normally used in such arrays. Pattern B resulted when the spacing was increased to ¼ wavelength. Pattern A was taken with ⅝-wavelength spacing between inner element ends. Note that the main lobe is longer and sharper with the wider spacings. Minor lobe content also increases, and this is a limiting factor in bay spacing in all types of arrays. A bay spacing of ⅝ wavelength is optimum for short Yagis, as well.

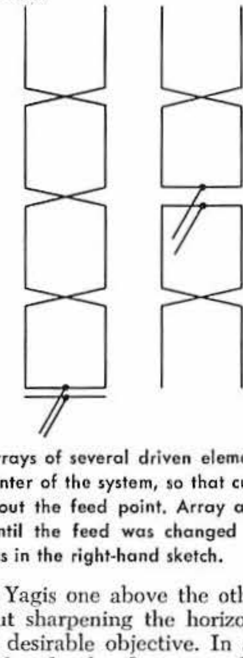


Fig. 8-12—Arrays of several driven elements should be fed at the center of the system, so that currents will be balanced about the feed point. Array at the left was ineffective until the feed was changed to the center connection, as in the right-hand sketch.

Stacking Yagis one above the other increases gain without sharpening the horizontal pattern —usually a desirable objective. In stacking another pair beside the first two, the optimum spacing depends on the length of the bays. One wavelength center-to-center is ordinarily used with booms less than one wavelength long. The half-the-boom-length minimum applies with longer ones.

Pattern effects with stacking are illustrated in Fig. 8-11, made with two 16-element collinears mounted side by side. Note that the pattern is markedly sharper with each wider spacing between halves of the array. The gain is also higher, but minor-lobe content increases rapidly at the wider spacings. Pattern A, made with ⅝ wavelength between the inner element ends, shows excellent gain, but the pattern is extremely sharp, and minor lobes are larger than for B, which was made with the bays spaced to leave ¼ wavelength between inner element ends (¾ wavelength between bay centers).

Going beyond ⅝ wavelength would result in no improvement in gain, for the minor lobes would be much larger. These would grow from here on, at the expense of the main lobe. A likely compromise between the maximum obtainable gain and the risk of large minor lobes is between ¼ and ½ wavelength between inner element ends, depending on what the builder wants most from his effort.

Also apparent from these patterns is the fact that with large collinears, as with large Yagis, it is not to our advantage to fill up an array with elements. Spacing out the inner element ends to a half wavelength probably nets as much as putting another set of elements in the space between the bays. The 48-element collinear array for 432 Mc., Fig. 9-67 is probably at least as effective as would have been a 60-element collinear of the same frontal area—and the former is much easier to build and feed properly.

Phasing and Feedpoint

Arrangement of phasing lines and the point of connection of the main transmission line are important factors in the performance of large arrays. Balance of currents about the central feedpoint is the critical point here, as the driven elements must be in phase if the system is to function properly. The author learned this the hard way years ago with a curtain array of 8 vertical half-wave elements in phase. This bidirectional system was first erected as shown at the left in Fig. 8-12. This was desirable mechanically, but the array worked very poorly. Changing the feed point to the center of the phasing system corrected the current unbalance, and turned this admittedly rather haywire arrangement into an effective v.h.f. array.

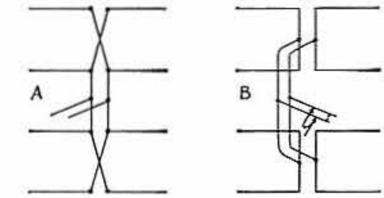


Fig. 8-13—In phasing large arrays no more than 8 elements should be connected to one line terminal, as at A. Even with 8 half waves in phase, it may be desirable to break the systems up into two parts, as at B, joining their midpoints with a phasing line. The phasing harness so used should be a half wavelength or multiple thereof each side of the main feed point. The universal stub, Fig. 8-18D, is very useful for feeding such a system.

The more driven elements there are in a phased system the more difficult it is to keep them in balance. Thus it is often desirable to break up a large driven system into several sets of elements, interconnected with phasing lines. The 48-element 432-Mc. array mentioned above is an example. Ordinarily no more than 8 elements should be in a single set, and breaking these up into two sets of 4 each may be better. See Fig. 8-13.

TRANSMISSION LINES

The best antenna is of little value if it cannot be made to accept power from the transmitter or transfer signals it intercepts to the receiver. Thus, selection of the right transmission line and an effective method of matching it to the antenna are of utmost importance. These factors are more vital to the v.h.f. man than to the occupant of lower frequencies, for even with the best lines losses run higher in v.h.f. installations than in the 80-through-10 station. It is easy to waste more than half our transmitter power in heating up the transmission line, and still more can be lost in radiation from it that should have been gone on to the antenna itself. Many 144-Mc. installations are at least this bad, and on higher bands power and received-signal losses may run up to 90 percent, with some lines that are fairly common in v.h.f. circles!

Coax, Twin-Lead or Open-Wire?

There are three principal types of transmission line commonly used in v.h.f. installations today. Each is obtainable in many styles and sizes, and each has its strong and weak points. There is no one "best" line, or we would not still be using all three. Choice of the right one begins with the line-loss information, Table 8-III, but this is by no means the whole story. These figures are for new lines, properly installed, and used in dry weather. Under average amateur-station conditions losses will almost certainly be greater than the table indicates.

Coax has relatively high loss in the tables. RG-8, perhaps the most commonly-used line, reputedly has a loss of about 2.5 db. per hundred feet at 144 Mc.—*if* the line is working perfectly. At 420 Mc., the same line, in new condition and perfectly installed, will dissipate 70 percent of your transmitter power and received-signal strength in a 100-foot run. Discouraging as these figures may seem, they are not the whole story. Transmitter power loss can be made up to some extent by increasing power, at least up to the legal limit, but in receiving the signal lost can never be recovered.

Good coax, on the other hand, is tolerant of installation. It is almost impervious to weather changes, and it can be installed anywhere. Tape it to a steel tower, or bury it; let it wrap around the tower and unwrap again as the beam is rotated—the loss will stay the same, almost regardless of conditions that adversely affect other types of lines. A prime advantage of coax that is often ignored is the fact that it permits measurement of the system performance readily,

TABLE 8-III CHARACTERISTICS OF COMMONLY-USED TRANSMISSION LINES

Type of Line	Conductor Size	Z_0 Ohms	Velocity Factor	Coax O.D., Inches	Attenuation in db./100 ft. Mc.: 50	144	220	420	Power Rating, Watts Mc.: 50	144	220	420
Open wire[1]	12	400-600	0.975		0.13	0.25	0.5	1	Over 1 kw.			
Open-Wire TV Line, ½-inch[2]	18	400	0.95		0.3	0.75	1	1.8	Over 1 kw.			
Open-Wire TV Line, 1-inch[2]	18	450	0.95		0.3	0.75	Not recommended		Over 1 kw.			
Parallel-Conductor Solid-Dielectric Twin-Lead[3]												
Standard Flat (214-056)	7/28	300	0.82		0.85	1.55	1.9	2.8	Over 1 kw. when dry			
Tubular (214-271)	7/28	300	0.82		0.85	1.55	1.9	2.8	Over 1 kw. when dry			
Tubular, Transmitting Type (214-076)	7/26	300	0.82		0.68	1.25	1.6	2.3	Over 1 kw. when dry			
Extra-Heavy Flat (Federal K-200)[2]	7/22	200	0.82		0.5	1	1.3	2	Over 1 kw.			
Coax, Solid-Dielectric												
RG-58/U[4]	20	53.5	65.9	0.195	3	6	7	15	350	175	125	90
RG-59/U	22	73	65.9	0.242	2.3	4.2	5	8	500	250	180	125
RG-8/U	7/21	52	65.9	0.405	1.5	2.5	3.5	5	1500	800	650	400
RG-11/U	7/26	75	65.9	0.405	1.55	2.8	3.7	5	1500	800	650	400
RG-17/U	0.188	52	65.9	0.87	0.5	1	1.3	2.3	4500	2300	1900	1200
Foamed RG-8A/U	7/21	50	75	0.405	1.22	2	2.75	3.9	1500	800	650	400
Aluminum-Jacket Foamflex												
⅜-Inch[5]	0.117	50	75	0.435	0.85	1.5	2	3	2200	1200	900	600
½-Inch[5]	0.162	50	75	0.60	0.65	1.2	1.5	2.3	3000	1600	1100	800
⅜-Inch[5]	0.077	70	75	0.435	0.82	1.5	1.9	2.9	similar to 50-ohm types			
½-Inch[5]	0.108	70	75	0.60	0.62	1.2	1.5	2.3				

[1] Spreaders at least 3 feet apart. Maximum spacing between conductors 1½ inches for 50 Mc., 1 inch for 144, ¾ inch for 220, ½ inch for 420. Loss Figures neglect radiation.
[2] Estimated loss, neglecting radiation.
[3] Numbers with 214 prefix are American Phenolic Corp.
[4] With all coax listed except RG-58, letter A, B or C after number signifies noncontaminating jacket. With 58, only RG-58 C/U has this type jacket.
[5] Not including vinyl jacket.

and with fairly inexpensive equipment. You can measure your s.w.r. and line loss, and the effects of any adjustments are immediately apparent. This is not easy with other types of line.

Twin-Lead is inexpensive and convenient to use. Its advertised losses look good on paper, compared with coax. The best grade of tubular Twin-Lead, transmitting type, is quoted at 1.25 db. per hundred feet at 144 Mc. and 2.3 db. at 420 Mc., but losses go up markedly in wet weather, and performance is very erratic. Flat ribbon gives the most trouble, but even the best tubular line will show fluctuating loading in heavy-rain conditions. Cheap lines with small conductors and thin insulation should be avoided entirely, unless the line is to be indoors or no more than a few feet long.

Book figures make open-wire line look best of all. If a good open line has only 0.2 db. loss at 144 Mc., why doesn't everyone use it? Even at 420 Mc., the loss per 100 feet can be under 1 db. This picture has the biggest "ifs" of all, however. Such fine results are achieved, if ever, only under the most carefully-controlled conditions. The conductors must be large, yet spaced closely so that radiation from the line will be negligible. Wire alignment must be kept constant, yet with a minimum of insulating spreaders and supports. There can be no sharp bends in the line, and it must be positioned so that it is balanced to ground.

These conditions definitely are *not* met in most amateur installations. We use TV-type lines, with too-small conductors and spacings generally too wide, at least for 420 Mc. There are spreaders every few inches. The line is often run close to a metal tower or eavethroughs, with little or no consideration of balance to ground. Nearly always there are bends of a sharpness that can be very harmful. One 220-Mc. line installed with reasonable care and using half-inch spaced open TV line showed a measured loss of 4 db. in a 125-foot run. This represents a transmitter power loss of 60 percent, yet it was probably a better-than-average amateur installation.

The potential low-loss qualities of open line can be realized in amateur work if sufficient care is taken in the construction and use of the line. Large conductors are a must; never less than No. 14, and No. 12 or larger is better. Spacing must be close in terms of wavelength; not more than 1 inch at 144 Mc. and proportionately less at higher frequencies if at all possible. Teflon is preferred for spreaders, and they should be several feet apart. If bends must be made, keep them to very obtuse angles, or in a continuous arc of large radius.

Baluns (about which more later) should be made and used with care. A 100-foot straight run of No. 12 enamelled wire, spaced ⅝ inch center-to-center with nylon spreaders every 6 feet, fed with baluns at each end, was measured for loss, including baluns, at 144, 220 and 432 Mc. It showed 1.1, 1.35 and 1.56 db., respectively, on these frequencies. By comparison a ½-inch TV line tested on 432 Mc. under identical conditions showed a loss of 2.3 db. These losses are somewhat higher than those of Table 8-III, but they represent the best that can be expected in a practical amateur installation. They also demonstrate the worth of good open-wire line, when it is used properly. If the line must be long, a good open-wire installation is probably the best way to do the job at moderate cost.

Tips on Selecting Coax

Coaxial line comes in two principal impedances: 52 and 72 ohms. There are small variations either side of these nominal figures, but they are of no significance for our purposes. Other impedances are available, but are seldom found in amateur installations. From the standpoint of overall effectiveness there is no preference between the above impedances, but practical factors tend to make the 52-ohm types the more useful. Most test equipment is set up for 52 ohms, for example. On the other hand, 72-ohm coax and a balun of the same material provides a good match to the 300-ohm balanced load that some v.h.f. antennas represent.

Losses in any line are related to conductor sizes and types of insulation. The small sizes of coax, with inner conductors of No. 22 wire or smaller, are bound to have high losses, regardless of quality or price. An inner conductor of No. 14 wire or its equivalent in stranded wire is about the minimum that should be used, except for short runs. Coax like RG-58 or 59 is convenient, but it should never be used for v.h.f. applications where the run is more than a few wavelengths. There is no easier way to waste power and lose receiving effectiveness!

Any coax costs money and good coax is quite expensive, but all things considered the best may turn out to be a good investment. Cheap coax is likely to be old, and its measured loss may be higher than figures given in the table. More important, older types of coax and some inexpensive new ones deteriorate quite rapidly when used outside. Be sure to find out whether or not the coax of your choice will stand up in outdoor service. "Non-contaminating" is the word for it. Coax guaranteed for 15 years of use, underground or otherwise exposed, is now available at moderate cost.

Coax is available in infinite variety. Worth looking for is the "polyfoam" version of standard types. These cost slightly more than solid-dielectric types, but losses are typically one-third less. Watch the velocity factor, however. Reduced density of insulation generally means a higher velocity factor. An electrical half wave-length will be a greater portion of a physical half wavelength with foam or other low-density dielectric than with solid.

Various lines are made with semiflexible sheathing, usually aluminum, and with spiral wrap or foam insulation. These are fairly costly, but they deliver excellent results and are fine for permanent installation. Flexible sections for rotation are needed with these, and a good way to handle a multiband installation is to put in a remotely-operated coaxial switch to permit the use of one line for all antennas.

About Coaxial Fittings

If you go to the expense of a good coaxial line, it is approaching the ridiculous to pinch pennies on the fittings to be used with it, particularly on 220 Mc. and higher. The so-called "UHF" fitting isn't to be trusted in the u.h.f. range, especially if you want to be able to measure antenna and feed-line performance with any degree of accuracy.

Probably the best fitting, for most of us, is the series N, a constant-impedance type that can be bought at moderate prices on the surplus market. It gives a constant impedance through the connection, and can be had in all types required. Properly installed, it is weatherproof.

Series C fittings provide constant impedance, and are weatherproof. In addition, they are quick-disconnect, and very handy on that account. However, they are not on surplus, and are quite expensive.

The BNC Series is nice, but too small for the RG-8-size line. The Type HN is a constant-impedance series, for the larger sizes of coax. Whatever series you select, be sure that the installation job is done properly. Water leaking into fittings will ruin the best system in short order. A sprayed coating of lacquer helps to prevent moisture absorption.

G-Line

Most u.h.f. amateurs are aware that there is a single-conductor transmission line, invented by Goubau, and called "G-Line" in his honor. Papers by the inventor appeared some years ago, in which seemingly fantastic claims for line loss were made; under 1 db. per hundred feet in the microwave region, for example. Especially attractive was the statement that the matching device was broad-band in nature, making it appear that a single G-Line installation might be made to serve on both 420 and 1215 Mc.

When u.h.f. TV first appeared on the scene, a G-Line kit was put on the market. Mainly because of its high cost (about $30.00, plus installation) it never sold well for home TV use, but it has since come into its own in cable TV systems. Here very long lines must be run, and losses must be held to a minimum, so the G-Line principle looks more attractive.

The basic idea is that a single conductor can be an almost lossless transmission line at ultra-high frequencies, if a suitable launching device is used. A similar launcher is placed at the other end. Basically the launcher is a cone-shaped device which is a flared extension of the coaxial feedline. In effect, the cone gets the energy accustomed to travelling on the inner conductor, as the outer conductor is gradually removed. The inner conductor should be large and heavily insulated. No. 14, vinyl covered, is supplied with the kit.

Since the kit was designed for home TV use the small end of the horn launcher has a balun of sorts for conversion from unbalanced to balanced line. This can be removed for amateur purposes, and the system fed directly with 72-ohm coax. The G-Line is very sensitive to bends. If any must be made, they should be in the form of an arc of large radius, this being preferable to even an obtuse-angle change in the direction of run. The line must be kept several inches away from any metal, and should be supported with as few insulators as possible.

A 100-foot run using direct 72-ohm feed to the launchers measured for loss at 432 Mc. showed 2.7 db., which may have been mainly in the launchers, since they were much too short to be really effective at this frequency. Theory states that the cones should be at least 3 wavelengths long, and the kit type is less than one wavelength at 432 Mc. Since loss in the line itself is presumably very low when properly installed, the G-Line idea should be useful where very long runs are required in u.h.f. and microwave work.

Practical Line Installations

It is one thing to quote losses for a straight Twin-Lead or open-wire line, without bends or insulating supports, and well away from metal or semiconducting objects such as trees, roofs and walls. It is quite another to put up a practical installation for an amateur station, where a line must be run from inside the house, be fastened to a tower part way up, and then allowed to swing free as the antenna is rotated. The inevitable losses and mechanical troubles that result from compromises inherent in the amateur approach, particularly with rotatable arrays, make a strong case for coax. But with any line these problems must be dealt with, and how we handle them can make a good many decibels difference in our signal reports, sending and receiving.

If coax is used it is best to support it frequently throughout the run and not depend on strain to keep it up out of harm's way. Burying coax is fine, provided that it is the noncontaminating variety. Lines of this type have a letter following the number. Example: RG8A/U is the noncontaminating version of RG8/U. The letter may be A, B or C, depending on other characteristics. Most coax made today is noncontaminating, but the buyer should watch this point in picking up "bargains."

Coax can be taped to a tower, so long as there is no abrasion to cut through the insulation. Sharp bends are best avoided, but only for mechanical rather than electrical reasons. Where coax must swing free, as in the portion that will rotate with the antenna, be sure that enough

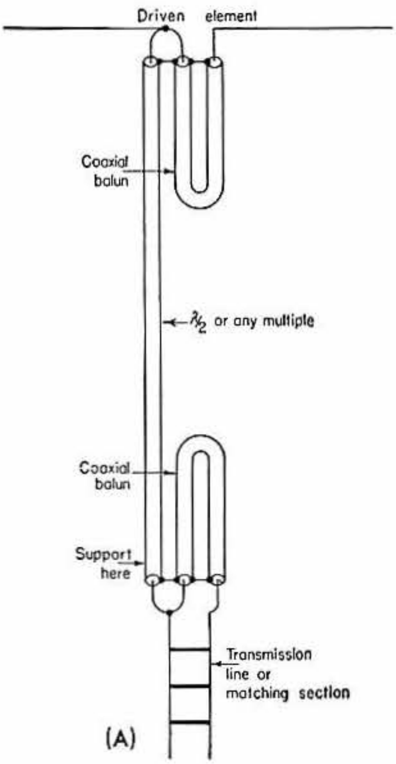


Fig. 8-14—Flexible sections for rotatable arrays. Coax may be used, as at A. If the coax section is any multiple of a half-wave length, the antenna impedance will be repeated at the bottom end. Twin-Lead may be used either as a Q section or as an impedance repeater, as shown in B.

slack is left to assure free rotation without additional strain. An extra turn or two around the tower, near the point of attachment to the beam, is usually desirable. Make all supports extra strong, to take care of extra loads imposed by ice and wind.

Properly handled, coax makes the best available rotating section for antennas that are fed with other types of line that may be more critical as to proximity to metal. Open-wire lines are particularly susceptible to breakage or shorting out unless special precautions are taken. Usually some form of insulated flexible line is connected between the antenna proper and a stationary support at the top of the tower or mast on which the antenna is mounted.

Such a flexible section can take several forms, and it can be made to do double duty as a matching device. Probably the most satisfactory method for arrays that are not to be fed directly with coax, is to use a flexible section of coax with baluns at each end, as shown in Fig. 8-14A. If the flexible section is made any multiple of a half wavelength electrically the impedance of the array will be repeated at the bottom of the flexible section.

A similar method is to use Twin-Lead for the rotating section, as shown in Fig. 8-14B. The 300-ohm tubular transmitting-type line is recommended. Here again, halfwave sections repeat the antenna impedance at the bottom end. Such a rotating section can also be made any odd multiple of a quarter wavelength, to act as a *Q* section, giving a step-down between a 450-ohm open line and a 200-ohm antenna impedance. More on these applications will be found in the text relating to matching devices that follows.

IMPEDANCE MATCHING

We know, or can determine, the impedance of the transmission line we want to use. If we knew the impedance of the antenna with equal certainty, matching one to the other would be a simple matter and one of our major v.h.f. antenna problems would be solved forever. Unfortunately, the actual impedance of an antenna is subject to so many variations that it is seldom possible to put a precise value on the impedance the transmission line must work into. Some kind of adjustable matching device is, therefore, a very useful tool.

Matching systems are many and varied, but all perform one basic function: that of impedance transformation, so that the feedline will "see" an impedance similar to its own, regardless of the actual antenna impedance. Matching may be combined with other functions, such as conver-

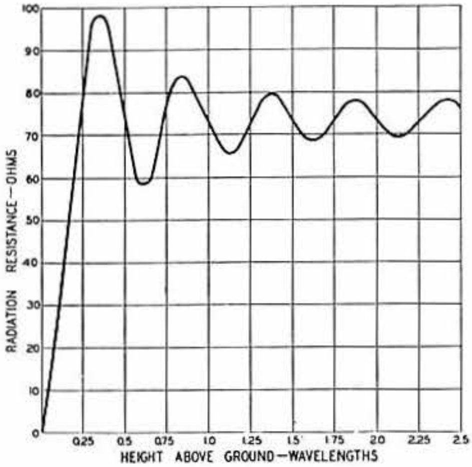


Fig. 8-15—Variation in radiation resistance of a horizontal half-wave antenna with height above perfectly conducting ground.

sion from an unbalanced line (coax) to a balanced load (center-fed antenna element). Matching may be included in the phasing lines connecting the bays of stacked arrays. The matching element may also be used to tune the system to resonance. We'll get to examples of all these methods shortly, but first a bit more about what they are going to be called upon to do.

About Antenna Impedance

This was discussed briefly earlier, but to review, a half-wave dipole in free space has an impedance of about 72 ohms. When the dipole is close to ground, or objects that simulate ground, its impedance changes. In the first half wavelength from the ground up, the impedance swings from a few ohms near ground, through the free-space value near 0.25 wavelength to as much as 100 ohms at 0.3 wavelength, and then back to 72 ohms at the half-wave point. Beyond here it drops off to 60 ohms and rises through 72 ohms again to nearly 85 ohms, then drops back to 72 ohms again at one wavelength. The effect of ground on impedance becomes relatively insignificant beyond two wavelengths, as shown in Fig. 8-15, but it can be seen that in situations most hams encounter in putting up antennas the impedance of a dipole is anything but a sure thing.

Ground is only one factor. Adding parasitic elements drops the impedance, but how much is anyone's guess, especially in arrays with both reflector and director elements. Length, diameter and spacing of these elements can effect great changes in the impedance of the driven element, to the point where it is almost impossible to predict what the feed impedance of a Yagi array will be. The best course, then, is to make the antenna first, determine its impedance by estimate or experiment, and then make a matching device to fit the requirements. If we can make a reasonable guess at the impedance, we can make an adjustable matching device of small range that will do the job.

If our antenna is just a half-wave dipole, Fig. 8-16A and B, we can assume 72 ohms, knowing from the curve of Fig. 8-15 that it cannot vary much more than 30 ohms either way. Adding a reflector will bring the impedance down—to 40 or 50 ohms, on the average. Putting on directors will lower it further, to something around 20 ohms. All these are for the fed point of the split dipole, A. At the center of a dipole that is unbroken, Fig. 8-16B, the r.f. voltage between the element and ground is zero. This point can thus be grounded, as in all-metal arrays, and the impedance matched by tapping the line out on the element in various ways.

R.f voltage and impedance at the ends of half-wave elements are very high. So is the feed impedance of two dipoles fed in phase at their inner element ends, Fig. 8-16C, the simplest collinear array. The feed impedance of an "H" array of four half-waves in phase is somewhere around 600 ohms. The popular v.h.f. collinear 16-element array (8 halfwaves in phase as in Fig. 8-13, but with reflectors) gets down to around 200 ohms—*maybe!* Remember that there are modifying factors, including that of coupling between elements, but 200 ohms is a good starting point for setting up a matching system for this type of array.

All these assumptions are valid approximations only for the frequency at which the system is resonant. If the array is out of tune all bets are off. We then must have some means of tuning the system before we can match it.

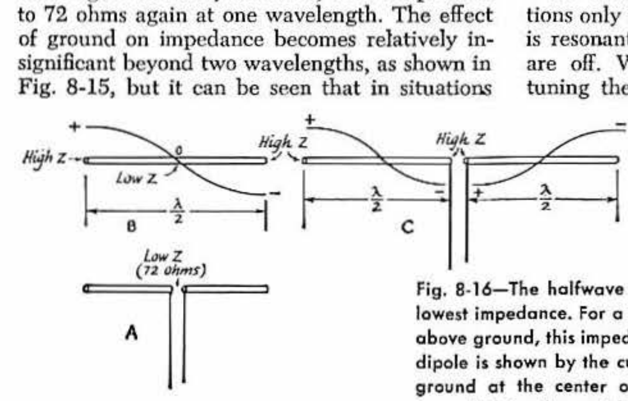


Fig. 8-16—The halfwave dipole, A, is fed at its center, the point of lowest impedance. For a dipole in free space, and at certain heights above ground, this impedance is 72 ohms. R.f. voltage on a halfwave dipole is shown by the curved line in B. Since there is no voltage to ground at the center of an unbroken dipole, this point can be grounded to the metal support. R.f. voltage and impedance are high at the ends of two collinear dipoles in phase, as at C.

COMMON MATCHING METHODS

We will not describe all kinds of matching systems, but will consider only those commonly used in v.h.f. work, or those that should get more attention.

The Delta

First there is the *delta* or *Y-match*, Fig. 8-17A. Here the transmission line is fanned out and tapped onto the driven element at points equidistant from the center. The taps can be adjusted until an impedance match is achieved, and then fastened permanently in place. One of the first impedance-matching devices ever employed, it still has its merits, not the least of which is simplicity. Chief fault is the likelihood of radiation from the fanned-out portion, especially when not properly proportioned. It is also quite frequency-sensitive.

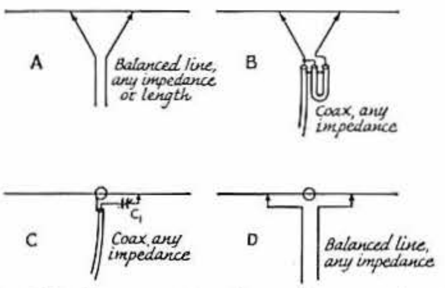


Fig. 8-17—The transmission line and antenna impedances may be matched by tapping the feedline out on the dipole in various ways. The delta or Y-match is shown at A. A variation for coaxial feed, using a balun, is given at B. The gamma match, C, is popular where coax feed is used. The T-match, D, may be fed with balanced line, or through a balun as in the case of B.

The delta works well with a balun made of coax, or an antenna coupler of some kind. A coaxial balun connected at the base of the delta is shown at B. If this is made of 72-ohm coax there could be a 300-ohm line of any convenient length between the balun and the delta. Adjustment is very easy when the delta is combined with coax feed. You merely insert an s.w.r. bridge in the coaxial line near the balun and adjust the delta side length and spread for zero reflected power. If the balun or balanced line is connected directly to the delta as shown in Fig. 8-17A and B, the lines can be of any impedances commonly available. More on baluns below.

Gamma and T-Match

Variations of the tapping-out idea are seen in the *gamma* and *T-match*, *C* and *D* of Fig. 8-17. The gamma is fine for coaxial feed, while the T is most often used with balanced line. A balun and coaxial feed could be used with the T, of course, just as with the delta. The series capacitor, C_1, is used to tune out the inductive reactance of the gamma arm. Without it the gamma system cannot be made to work perfectly, as a slight unbalance is always present. The gamma arm is usually made of tubing of about the size of the driven element, and a sliding clip is used between the two, to facilitate adjustment. The capacitor can be at either end of the arm.

Once the proper value is found for C_1 it can be removed and a fixed capacitor substituted. An assumed value for your line can be taken, and only the point of connection of the arm made adjustable. Suitable fixed values for 50 ohms are as follows: 50 Mc.—65 pf., 144 Mc. —20 pf., 220 Mc.—15 pf., 432 Mc.—8 pf.

Strictly speaking, series capacitors should be used with the T system too, but since omitting them does not upset the balance of the dipole, as it would with the one-sided gamma, they are not always used.

Folded Dipole

One of the most commonly-used matching devices is the *folded dipole*, shown in various forms in Fig. 8-18. When a single conductor is bent around as shown at A, the impedance seen by the transmission line is quadrupled. Thus a folded dipole made from one size of conductor throughout has an impedance of 4×72, or 288 ohms, and it can be fed with 300-ohm line, or with a balun and 72-ohm coax, without appreciable mismatch. The dipole element can be made from a piece of Twin-Lead, with each outer end shorted and one conductor broken at the midpoint, for connecting the transmission line. This

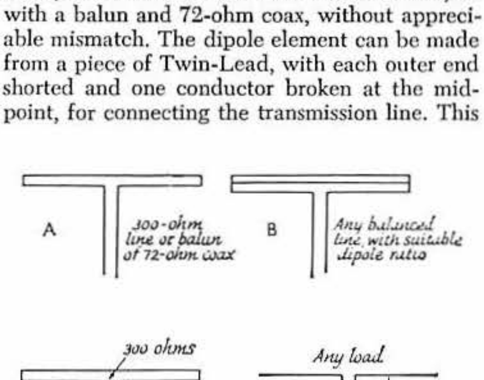


Fig. 8-18—A single conductor may be bent as at A to form a folded dipole, giving an impedance four times that of a simple split dipole. It may thus be fed with 300-ohm balanced line, or 72-ohm coax and a balun. Higher impedance step-up can be achieved by making the unbroken portion of the dipole of a larger conductor, as at B. A quarter-wavelength matching transformer, or Q section, is shown at C. A matching device that is useful for any balanced load is the universal stub, D. The transmission line can be coax or balanced line, any impedance.

is a convenient arrangement for temporary or indoor use.

Additional impedance step-up can be obtained by making the unbroken portion of the dipole of larger cross-section than the fed portion, as shown in Fig. 8-18B. This is widely used in parasitic arrays, where the feed impedance is nearly always much lower than 72 ohms. Impedance step-up depends on the ratio of the conductor sizes, and on the spacing between them. If the approximate impedance of the antenna is known, a suitable element can be made for 50 or 144 Mc. by using the nomogram, Fig. 8-19.

Where the spacing between the portions of this type of dipole is an appreciable portion of a wavelength, as it must be at 220 or 420 Mc., the information of Fig. 8-19 is no longer reliable. A better method of matching arrays for these frequencies is to use the universal stub, Fig. 8-18D, or the Q section, Fig. 8-18C. For more on matching Yagis for 220 and 432 Mc., see practical examples in Chapter 9.

A problem with folded dipoles is that one must know the impedance to be matched in order to design one to do the job. Educated guesses may come close enough for most practical purposes. For example, if we assume the feed impedance of a Yagi array to be 20 ohms, we can use a folded dipole with a 15-to-1 step-up as the driven element, and feed the array

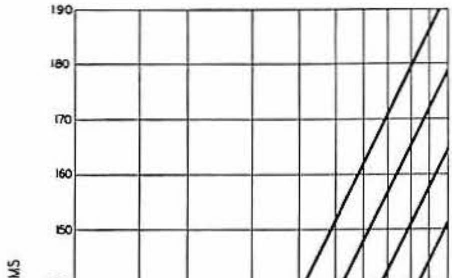


Fig. 8-20—Characteristic impedance of typical air-insulated coaxial lines.

with 300-ohm Twin-Lead. The mismatch will be slight, even if the dipole impedance turns out to be 15 ohms, or 25 ohms, instead of 20. The s.w.r. will be only about 1.2 to 1 in either case. We could use a 10-to-1 dipole and a 50-ohm balun equally well.

The folded dipole is easy to make, and it is somewhat more frequency tolerant than some other matching systems. It is very useful in stacked-Yagi arrays having open-wire phasing lines. A fairly high value of dipole impedance is desirable here, but the exact value is not important, as matching will be taken care of where the main transmission line connects to the phasing section.

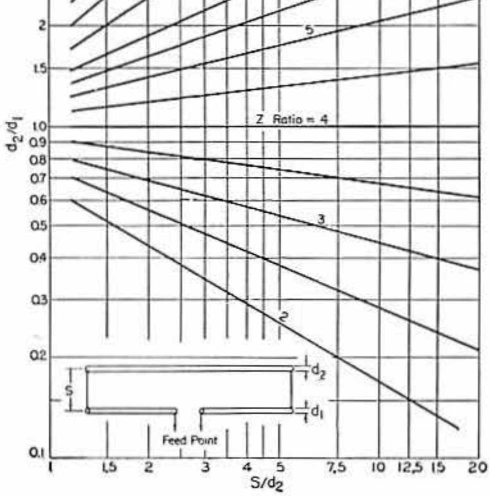


Fig. 8-19—Impedance step-up ratio for the two-conductor folded dipole, as a function of conductor diameters and spacing. Dimensions d_1, d_2 and S are shown on the inset drawing. This information is not reliable for use on amateur bands above 148 Mc.

The Q Section

A quarter wavelength of transmission line has the property of acting as a matching transformer between two different impedances. Such a transformer is called a Q section, and an example is given in Fig. 8-18C. Here a 300-ohm dipole is matched to a 500-ohm line by using a Q section whose impedance is equal to the square root of the product of the two impedances to be matched. A 375-ohm section is required here, but the principle may be applied to many v.h.f. matching problems. The impedance obtainable with various conductor sizes and spacings is obtainable from Fig. 8-20 for coax, and 8-21 for balanced lines. Our 375-ohm Q section could be two No. 10 wires spaced 1¼ inches apart, or two ¼-inch rods 2¾ inches apart, to show two examples.

An adjustable Q section is a convenient way of matching two impedances that are known only approximately. Two ¼-inch rods can be made to provide impedances of about 210 to 400 ohms, by varying their spacing from ¾ to 3 inches. The system can be used to step up as well as down, and it works with coaxial or parallel conductors. We'll have examples later.

The Corrective Stub

Probably the most useful matching device of all is the universal stub of Fig. 8-18D. Because the matching stub must be a half wavelength or more to start, it is cumbersome at 50 Mc. or lower, but it is ideal for 144 Mc. and higher bands. No impedances need be known to utilize it, and within limits the system to be matched does not have to be resonant. The short on the line is adjusted to resonate the system to be fed, and then the transmission line is tapped onto the stub at the matching point. The load can be any impedance, the stub can be any convenient wire or tubing size, and any spacing. The feedline can be coaxial or balanced, any impedance. A balun is used with coax as indicated in the sketch. The shorting bar can be grounded, and the unused portion of the stub cut off, once adjustment is completed.

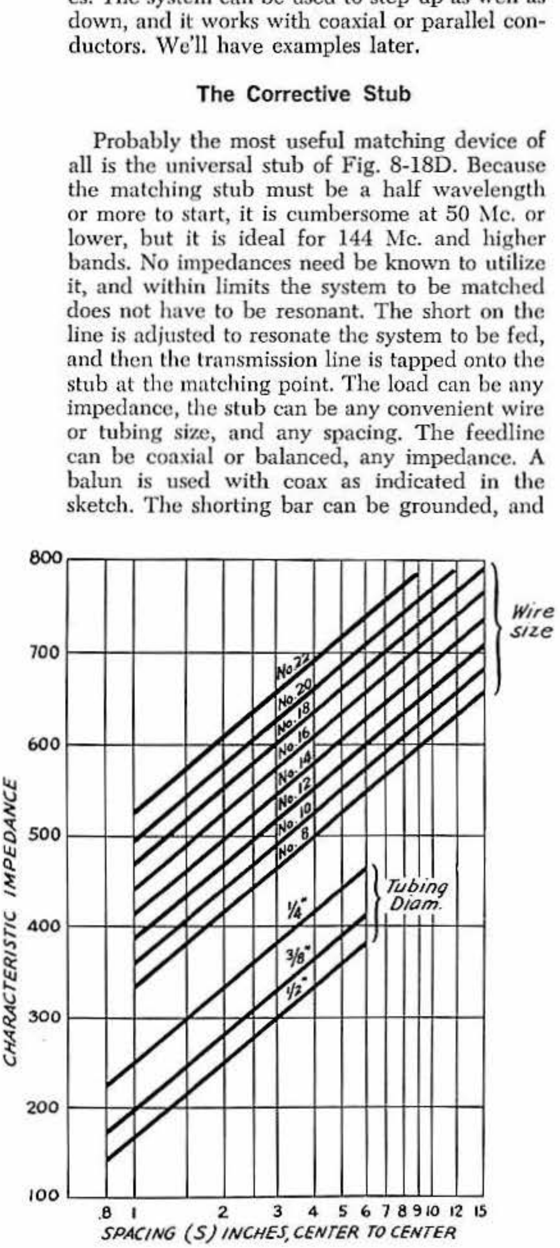


Fig. 8-21—Characteristic impedance vs. conductor size and spacing for parallel-conductor lines.

Two variables are involved, which complicates the adjustment procedure a bit, but with a standing-wave bridge in the line the job is quite simple. You merely move the position of the short and the point of connection of the transmission line until zero reflected power is indicated on the s.w.r. bridge. Coupling *at the transmitter* is then adjusted for the desired loading.

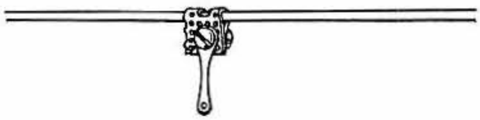


Fig. 8-22—Clip for use in adjusting the point of connection of a balun, or the adjustable short of Fig. 8-18D, made from a piece of perforated aluminum. Balun leads are soldered to the lug. When the adjustment process is completed, the clip may be removed and the connection soldered permanently to the line.

Where the point of connection of a balun or shorting bar must be made adjustable, a small clip of perforated aluminum, Fig. 8-22, is handy for a temporary connector. The holes are already made, and with some tension on the clip the edges of the aluminum bite into the conductor slightly, assuring good contact. Small Fahnstock clips are also useful. When adjustment is completed, remove the clip and solder the connection permanently, using the same overall lead length.

Making and Using Baluns

As its composite name implies, a balun is a device for working between an unbalanced line (coax) and a balanced line or load. It can take several forms, some of which also include the function of impedance matching along with the unbalanced-to-balanced conversion.

The Antenna Coupler Probably the most versatile of baluns, the antenna coupler, Fig. 8-23, can be made to work from any impedance of coaxial line at J_1 to any impedance balanced load at J_2. The low-impedance input circuit, L_1C_1, is series resonant at the operating frequency, and inductively coupled to the balanced circuit, L_2C_2. The balanced output, connected to J_2, is tapped down on L_2 an equal amount from each end.

Component values in the antenna coupler are not critical, and it will handle a wide range of impedance combinations merely by adjusting the capacitors. Changing the tap positions on L_2 extends the range of impedances still further. The values of L_1 and C_1 should be roughly those that have inductive and capacitive reactance equal to the value of the coaxial line impedance. Since the value of capacitance is the more readily estimated, it is customary to aim for this and

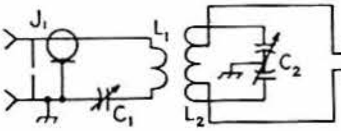


Fig. 8-23—Circuit and parts information for the v.h.f. antenna couplers.

C_1—100-pf. variable for 50 MC., 50-pf. for 144 Mc. (Hammarlund MC-100 and MC-50).

C_2—35-pf. per-section split-stator variable, 0.07-inch spacing (Hammarlund MCD-35SX). Reduce to 4 stator and 4 rotor plates in each section in 144-Mc. coupler for easier tuning; see text.

J_1—Coaxial fitting, female.

J_2—Two-post terminal assembly (National FWH), or crystal socket.

L_1—50 Mc.: 4 turns No. 18 tinned, 1-inch diameter, ⅛-inch spacing (Air-Dux No. 808T) inside L_2.
144 Mc.: 1½ turns No. 14 enam., 1 inch diameter, ⅛-inch spacing. Slip over L_2 before mounting.

L_2—50 Mc.: 7 turns No. 14 tinned, 1½ inch diameter, ¼ inch spacing (Air Dux No. 1204). Tap 1½ turns from each end.
144 Mc.: 5 turns No. 12 tinned, ½ inch diameter, ⅞ inch long. Tap 1½ turns from each end.

adjust the size of L_1 to resonate with it. Approximate values for the various bands are as follows: 50 Mc.–65 pf., 144 Mc.–25 pf., 220 Mc.–15 pf., and 420 Mc.–10 pf. A variable capacitor used for C_1 should be chosen so that these values can be reached with some to spare. Often a fixed capacitor of approximately the above value will suffice, adjustment then being made entirely with C_2.

For adjustment of the coupler an s.w.r. bridge should be connected in the coaxial line between the antenna changeover relay and J_1. The two capacitors are then adjusted for zero reflected power, as indicated on the bridge. If this results in unsatisfactory transfer of power from the transmitter, the loading control in the transmitter should be readjusted for maximum forward power on the bridge meter. Do *not* adjust the antenna coupler for maximum forward power reading; always set it for zero reflected. This applies in any matching adjustment.

Connected as described, the antenna coupler will aid in reception, reducing the strength of any out-of-band signals before they reach the receiver, where they might otherwise cause overloading and other spurious responses. The coupler is also an effective filter, attenuating any unwanted frequencies present in the transmitter output, before they reach the antenna.

The coupler can be connected at any point between the transmitter and the antenna where the conversion between the unbalanced and balanced lines is desired. Because of the need to retune the coupler for appreciable frequency excursions, it is usually mounted within easy reach of the operating position. If rack-mounted, as shown in the 50- and 144-Mc. units of Fig. 8-24, they can be included in the transmitter assembly if it is convenient to do so.

The couplers are identical circuit-wise, and are mounted inside a standard 3 by 4 by 17-inch chassis. A bottom plate, not shown, is added to complete the shielding. The 50-Mc. coils are made of commercially available stock, though they can be made by hand if desired. The coupling winding, L_1, is inserted inside the tuned circuit. The polyethylene strips on which the coils are wound keep the two coils from making electrical contact, so no support other than the wire leads is needed.

Leads to L_1 are brought out between the turns of L_2, and are insulated from them by two sleeves of spaghetti, one inside the other. Do not use the soft vinyl type of sleeving, as it will melt too readily if, through an accident to the antenna system, the coil should run hot. In the 144-Mc. coupler the positions of the coils are

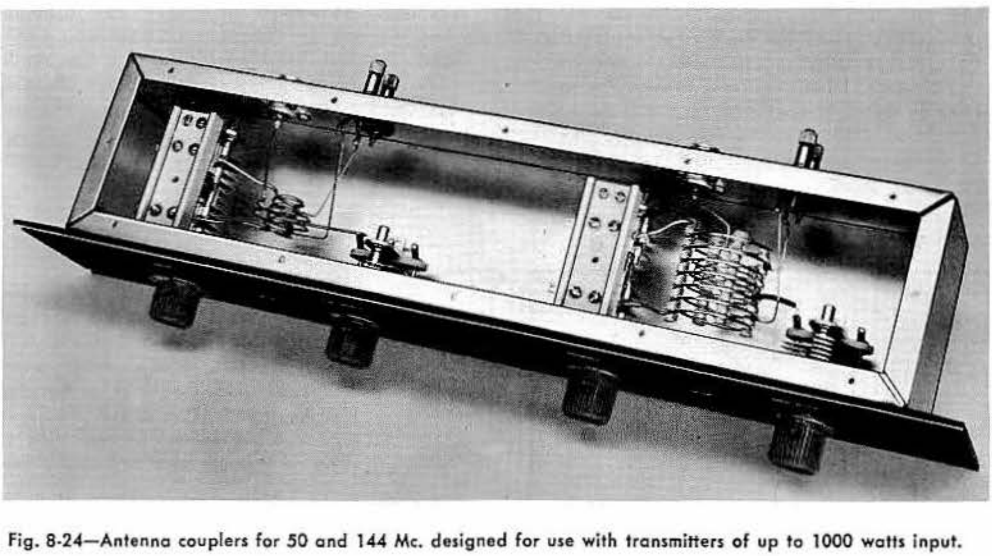


Fig. 8-24—Antenna couplers for 50 and 144 Mc. designed for use with transmitters of up to 1000 watts input.

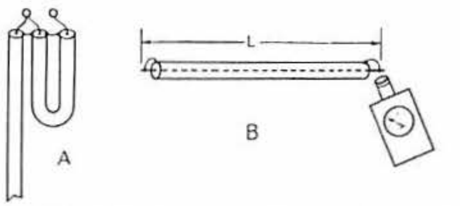


Fig. 8-25—A balun for working from coaxial to balanced line is shown at A. Impedance at the balanced end, top, is four times that of the coaxial line used. The loop is an electrical half wavelength. Its resonant frequency may be checked with a dip meter as shown at B.

reversed, with the tuned circuit, L_2, at the center, and the coupling coil outside it.

Similar tuning capacitors are used in both couplers, but some of the plates are removed from the one in the 144-Mc. circuit. This provides easier tuning, though it has little effect on the minimum capacitance, and therefore on the size of the coil.

4-to-1 Baluns Broad-band baluns of several types are readily constructed. Bifilar-wound coils can be used in the same manner as on lower frequencies, but this method is seldom used above 30 Mc. The most common balun for v.h.f. service is made from an *electrical* half-wavelength of coax, usually the same type as used for the main transmission line, folded back on itself and connected to the main line and the antenna as shown in Fig. 8-25. This balun provides an impedance step-up of 4 to 1, while handling the unbalanced-to-balanced conversion.

The physical length of the balun loop will vary with different types of coaxial lines. With solid-dielectric coax the loop will be about 65 percent of the free-space value for a half wavelength. Less dense insulation such as foamed polyethylene may increase this to as much as 80 percent of the free-space value, so it is well to check the loop for resonance. Using the length of leads that will be involved in the eventual connections, short the ends as shown in Fig. 8-25B, and couple the dip meter to one end.

The coaxial balun is cumbersome for use below 50 Mc., and it ceases to be practical above 450 Mc. At the lower frequency the loop can be rolled into a coil if desired. (See Fig. 9-6.) Usually it is taped to the main transmission in U shape. It is best to run the balun perpendicular to the load, which would mean dropping it vertically from the boom of a horizontal Yagi. The main line can then be looped back to the boom and taped in place, if this is desirable mechanically. A permanent position for the balun with respect to its load is particularly important at 220 or 420 Mc. The loop for the latter frequency is only about 8 inches overall, so mechanical variations can throw the balance the loop is supposed to provide quite a bit off.

Small coax such as RG-58 or 59 is not recommended for baluns. Soldering weakens and distorts the insulation, making shorts likely, and the small conductors break very easily. Losses in small-coax baluns often run prohibitively high. RG-8 or similar sizes are much better, if the balun is made and mounted with care.

The impedance at the end of the balun is 4 times that of the coax used. A balun with 52-ohm coax will match a 200-ohm load, and 72-ohm coax and a balun match a 300-ohm load.

1-to-1 Baluns The unbalanced to balanced conversion can also be made without an impedance change, using the balun shown schematically in Fig. 8-26A. Here a split dipole or other balanced load is fed directly with coax. This would make for unbalance and r.f. flow on the outer conductor of the line, but for the detuning sleeve (or *bazooka*) that has been added to the last quarter-wavelength of the line. Being open at the top and shorted to the outer conductor at the bottom, this sleeve presents an infinite impedance to r.f. at the resonant frequency, effectively choking off current flow, and preventing radiation from the line.

A similar effect can be achieved with the bazooka in B, wherein a quarter-wave line section shorted at the bottom is formed by connecting an additional piece of coax or tubing as shown. This is less effective than the sleeve method however, and is seldom used above the 50-Mc. band. It is used occasionally for v.h.f. mobile antennas wherein it may be convenient for feeding a split driven element directly with coax. In this application the bazooka is usually a piece of small coax similar to the main line. Taping the two pieces together leaves the issue of true electrical length somewhat in doubt, and other feed

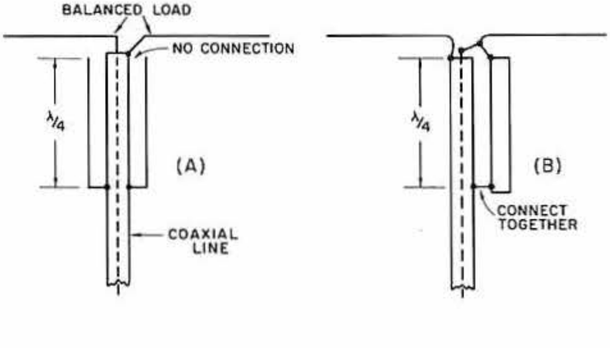


Fig. 8-26—Two types of baluns, for conversion from a coaxial line to a balanced load. The coaxial sleeve, A, is the preferred type, and it also can be used as a matching device, as described in the text. Both serve the same main purpose: prevention of current flow on the outer conducter of the coaxial line.

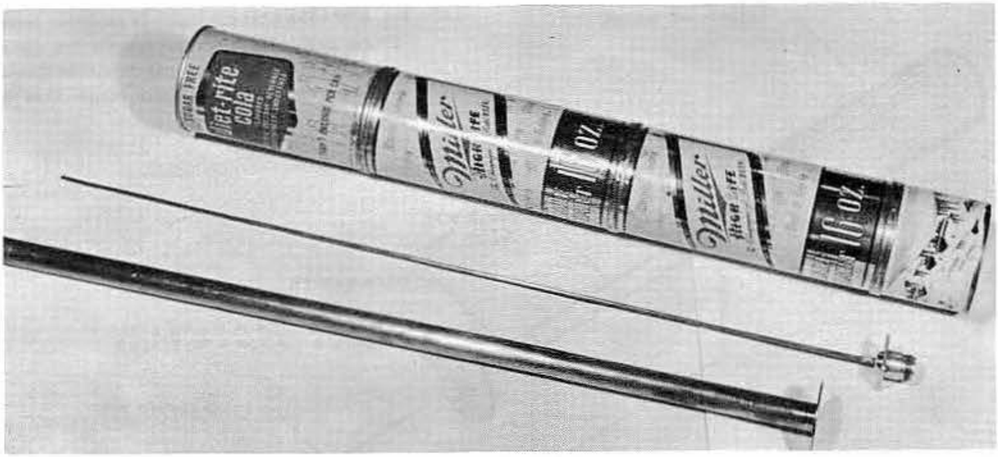


Fig. 8-27—Coaxial-sleeve balun for 144 Mc., showing the parts that make up the air-dielectric matching section.

TABLE 8-IV

Lengths for the decoupling sleeve, A, and copper pipe outer conductor, B, for 144, 220 and 432 Mc.

	A	*B*
144 Mc.	$19\frac{3}{4}''$	$20\frac{5}{16}''$
220 Mc.	$12\frac{1}{2}''$	$12\frac{29}{32}''$
432 Mc.	$6\frac{1}{4}''$	$6\frac{31}{32}''$

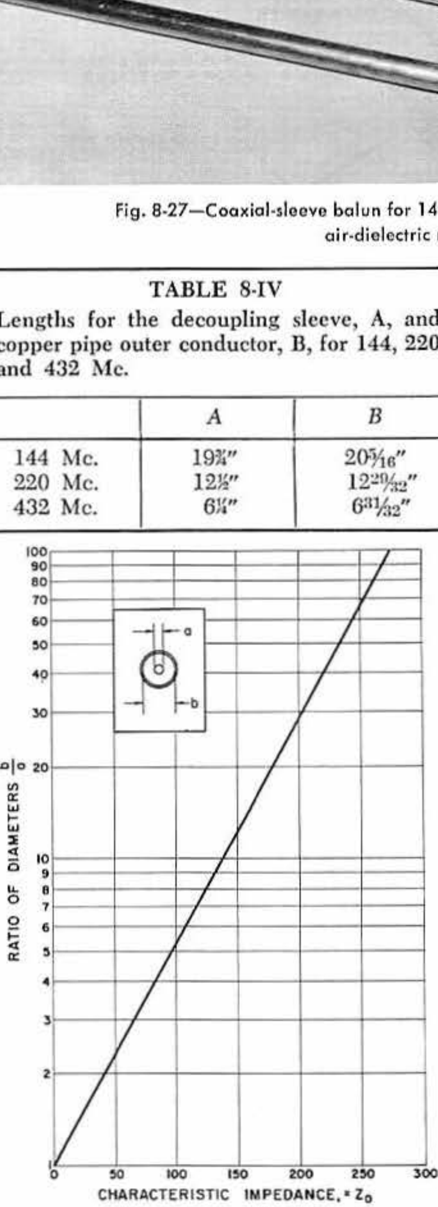


Fig. 8-28—Characteristic impedance of coaxial matching sections for various conductor diameter ratios. The outside diameter of the inner conductor and the inside diameter of the outer conductor are used.

methods are generally preferable.

Impedance-Matching Balun Since the sleeve assembly of Fig. 8-26A is a quarter-wavelength long, it is a simple matter to make it serve as a *Q* section for impedance matching, as well as a balun. Examples are the beer-can baluns of Fig. 8-27 and 29, made by K6HCP and WA6GYD. These are assemblies to which the main coaxial line run is attached by means of standard coaxial fittings. By making the inner portion of the balun of the right combination of conductor sizes it is made to act as a coaxial *Q* section. Construction of the balun is detailed in Fig. 8-29. Lengths of the sleeve and coaxial section for 144, 220 and 432 Mc. are given in Table 8-IV.

The inner coaxial portion of the balun can be made to continue the line impedance, or transform it to other impedances. If standard wire sizes are used for the inner conductor, load impedances from 70 to 450 ohms can be matched by using 9/16-inch i.d. copper pipe for the outer conductor. The matching combinations for various wire sizes are given in Table 8-V. You can

TABLE 8-V

Inner conductor wire sizes to be used with $\frac{9}{16}$-inch i.d. copper pipe outer conductors, for various impedance matching jobs commonly encountered in v.h.f. work. The impedance of the main coaxial transmission line, Z_s, is given in the left column. Next is the balanced load, Z_r, to be matched.

Z_s, *ohms*	Z_r, *ohms*	*Wire Size, A.W.G.*
50	72	4
50	200	10
50	300	12
50	450	18
75	200	12
75	300	18
75	450	24

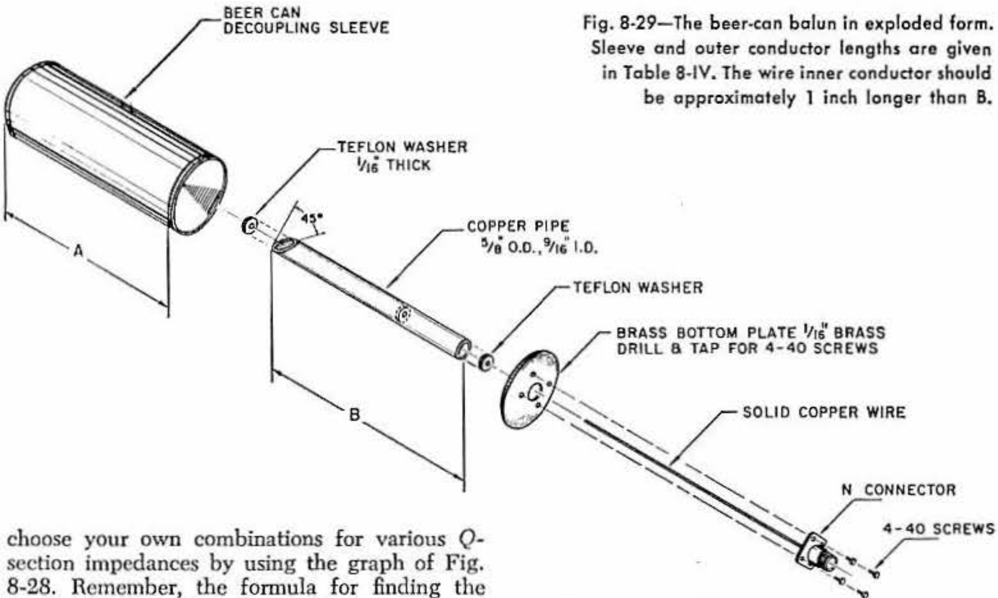


Fig. 8-29—The beer-can balun in exploded form. Sleeve and outer conductor lengths are given in Table 8-IV. The wire inner conductor should be approximately 1 inch longer than B.

choose your own combinations for various *Q*-section impedances by using the graph of Fig. 8-28. Remember, the formula for finding the needed *Q*-section impedance, Z_o, is

$$Z_o = \sqrt{Z_s Z_r} \qquad (3)$$

Where Z_s is the impedance of the main transmission line, and Z_r is the impedance to be matched.

The load to be matched can be either balanced or unbalanced, but if it is the latter the outer sleeve is not needed. The diameter of the detuning sleeve is not critical; it just happens that beer cans work out conveniently. Let's say we want to match 50 to 300 ohms, a frequently-encountered situation that cannot be handled with a flexible balun. From formula (3) we find that we need a *Q* section with an impedance of 122 ohms. Fig. 8-28 tells us that a *b/a* ratio of 7.5 is needed. With standard 9/16-inch i.d. plumbing copper pipe and a No. 12 wire we can take care of this job nicely. Many other usable combinations can be worked out, using pipe and wire sizes that are readily available.

FEEDING STACKED AND PHASED ARRAYS

If individual bays of a stacked array are properly designed they will look like noninductive resistors to the phasing system that connects them. The impedances involved can thus be treated the same as resistances in parallel, if the phasing lines are a half wavelength or multiple thereof. The latter point is important because the impedance at the end of a transmission line is repeated at every half wavelength along it.

In Fig. 8-30 we have three sets of stacked dipoles. Whether these are merely dipoles or the driven elements of Yagi bays makes little difference for the purpose of these examples. Two 300-ohm antennas at A are one wavelength apart, resulting in a feed impedance of approximately 150 ohms at the center. (It will be slightly less than 150 ohms, because of coupling between bays, but we can neglect this for practical purposes.) This value holds regardless of the impedance of the phasing line. Thus, we can use any convenient type of line for phasing, so long as the *electrical* length is right.

The velocity factor of the line must be taken into account. As with coax, this is subject to so much variation that it is well to make a resonance check if there is any doubt. The method is the same as for coax, Fig. 8-25. A half wavelength of line is resonant both open and shorted, but the shorted condition (both ends) is usually the more readily checked.

The impedance-transforming quality of a quarter-wavelength of line can be employed in combination matching and phasing lines, as shown in B and C of Fig. 8-30. In B, two bays spaced a half-wavelength are phased and matched by a 400-ohm line, acting as a double *Q* section, so that a 300-ohm main transmission line is matched to two 300-ohm bays. The two halves of this phasing line could each be 3 or 5 quarter-wavelengths long equally well, if these lengths serve any useful purpose. An example would be the stacking of two Yagis, where the desirable spacing is more than one-half wavelength.

A double *Q* section of coaxial line is illustrated in Fig. 8-30C. This is useful for feeding stacked bays which were originally set up for 52-ohm feed. A spacing of ⅝ wavelength is optimum for small Yagis, and this is the equivalent of an electrical full wavelength of solid-dielectric coax, such as RG-11/U. If our phasing line is made one quarter-wavelength on one side of the feed and three quarters on the other, one driven element should be reversed with respect to the other to keep the r.f. currents in phase. If the number of quarter-wavelengths is the same on either side of the feedpoint the two elements should be in the same position, not reversed as shown in C.

One marked advantage of coaxial phasing lines is that they can be wrapped around the vertical support, taped or grounded to it, or arranged in any way that is convenient mechanically. The spacing between bays can be set at the most desirable value, and the phasing line placed anywhere necessary to use up the required electrical length.

Making adjustments

Wherever adjustable matching devices are used, any really effective adjustment procedure must be carried out either with the antenna in the position in which it will eventually be used, or under conditions simulating the eventual installation. The thought of making adjustments at the top of a tower is often a bit staggering to the budding big-antenna enthusiast, but fortunately such a high-wire act is not really necessary. There are right and wrong ways to do the job at ground level, however. From preceding discussion of the effect of ground on antenna impedance it is easy to see that matching adjustments made with an array close to ground could be quite a bit off when the array is hoisted to its eventual resting place 50 feet or more in the air. Furthermore, even if there were no impedance change from the effect of ground, objects quite some distance out in front of the array may reflect enough power back to it so that an appreciable reflected-power reading is observed from effects other than actual mismatch. The bigger and sharper the array, the more troublesome these reflections become.

The solution to this problem is obvious, but not too many antenna workers seem to think of it: aim the beam straight up, with the reflectors close to ground. If the front-to-back ratio is 20 db., the amount of power that will be radiated downward with the beam in the straight up position is negligible, and so is the effect of ground on the antenna impedance. This lazy-man approach has been used many times, on bands from 144 through 432 Mc., and on each occasion it has resulted in very close to optimum matching when the array was finally installed in the tower position. Very much better results are possible in this way than with the array's line of fire parallel to and near the ground.

How Important is Matching?

Due mainly to over-exposure to the term, a good many hams tend to worship perfect matching. To have a 1-to-1 s.w.r. is the ultimate achievement, for them. But is it so very important? Not necessarily! It depends on what you're going to do. As may be determined from Fig. 8-31, a 100-foot line of RG-8 coax at 144 Mc. will have its loss increased by less than 0.5 decibel with a 2:1 s.w.r. compared to a perfectly-matched line. If the loading on the transmitter is adjusted properly and the line is trimmed for length, if necessary, a listener at a distant point would not be able to tell the difference. Note that this line trimming is to achieve a resonant condition and proper loading. *It does not affect the s.w.r.!*

Mismatch is important in some ways, and it can tell you things about your antenna system. Make a frequency run, measuring s.w.r. at 144, 144.5, 145, 145.5, 146 and so on. If your s.w.r. dips to near 1:1 at 147 Mc., and is 3:1 at 144, you need some work on your array. You're sure to be getting less than top performance at the low end. But if 2:1 is as low as you can get, and it is around the frequency you work most often, you don't need to worry too much if the transmitter loads satisfactorily.

With high power a high s.w.r. runs you into the danger of flash-over of the line, but this

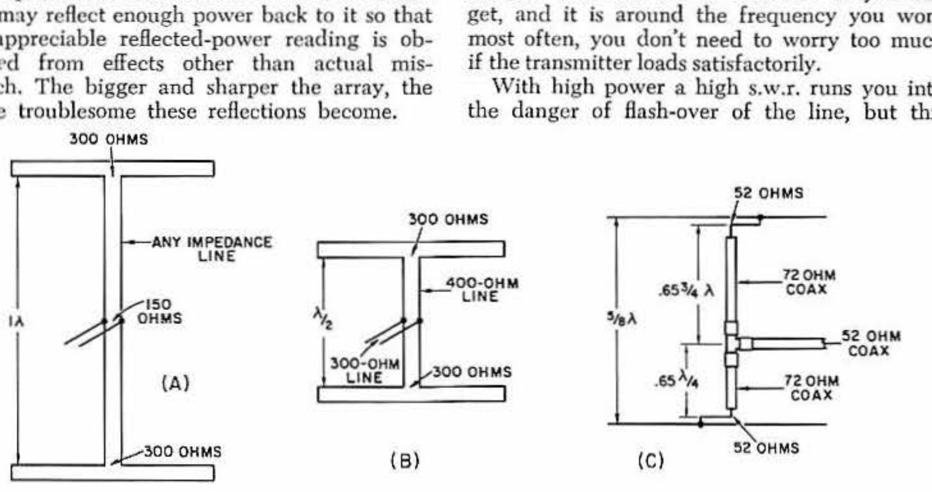


Fig. 8-30—Three methods of feeding stacked v.h.f. arrays. A and B are for bays having balanced driven elements, where a balanced phasing line is desired. Array C has an all-coaxial matching and phasing system.

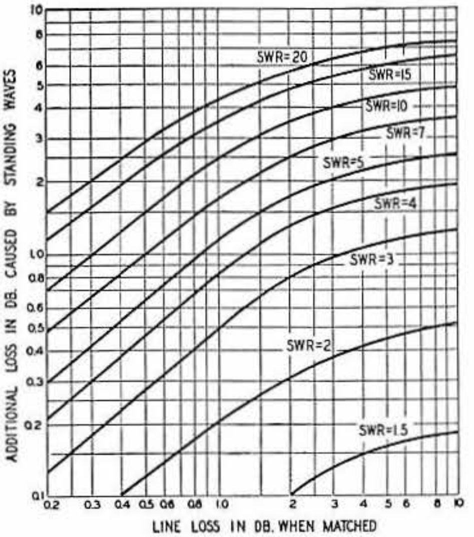


Fig. 8-31—Increase in line loss because of standing waves. To determine the total loss in decibels in a line having an s.w.r. greater than 1, first determine the loss for the particular type of line, length and frequency, on the assumption that the line is perfectly matched (Table 8-III). Locate this point on the horizontal axis and move up to the curve corresponding to the actual s.w.r. The corresponding value on the vertical axis gives the *additional* loss in decibels caused by the standing waves.

doesn't happen very often in v.h.f. circles, at least with any coax worth using.

Exact matching *is* important in making measurements of antenna performance. If you would learn anything from attempted gain measurements you have to know *exactly* how much power you're putting into the antenna, or at least you have to know that you're using the same power every time. Forward-power readings with the usual s.w.r. bridge are useless for antenna evaluation purposes, unless the system is perfectly matched. This means adjusting for *zero* reflected power, every time a comparison or measurement is made.

Much of the conflicting evidence reported in articles on antennas over the years has resulted from a lack of understanding of the importance of this precaution. Just putting up a field-strength meter and then pruning the elements or adjusting their spacing for maximum meter reading may result in your having a fairly good antenna, but it is a wholly unreliable way to make measurements. If you find the element lengths and spacings recommended in much of the literature on antennas confusing, failure to keep the radiated power constant, or inability to determine it accurately, may well be at the bottom of most of the inconsistencies.

Using the S.W.R. Bridge

Coaxial feed is recommended, if only for the reason that it permits easy monitoring of the matching process. You merely connect a standing-wave bridge in the coaxial line and adjust the matching device for lowest possible reflected power. This should be zero, or very close to it. All that is left then to make your antenna radiate effectively is to adjust the coupling at the transmitter for maximum forward power on the bridge meter. Note that you do *not* adjust the matching device for maximum forward power; you adjust for *zero* reflected.

Where the bridge is inserted in the line is important. Many hams are happy about their antenna systems because a bridge connected in the line at the transmitter output shows zero reflected power, but they may be in a fool's paradise. If the transmission line is long in terms of wavelength, and lossy (all coaxial lines are lossy enough to throw us off) the line may, in effect, be self-terminating. That is to say you can have the world's worst mismatch at the end of a 100-foot run of RG-8 on 432 Mc. and you'll never know it if the bridge is connected at the transmitter. Try a direct short on the end of your line, or disconnect the antenna entirely, and see how little difference it makes on *your* line. Remember these are the ultimate extremes of mismatch! The bridge must be connected at or near the antenna, when making matching adjustments.

There is no way to adjust an antenna properly without a bridge. Repeat—*no way!* Don't try to do without one, for it is probably the most important instrument you can own. It need not be fancy or "commercial." A very simple unit is described in Chapter 11. It works well on 50 through 450 Mc., and it costs only a few dollars to make. Its meter is connected so that it can be used for other transmitter test jobs as well.

If you have not already done so, get the bridge habit. It will be the best step you've ever taken in the v.h.f. antenna field.

Building and Using V.H.F. Antennas

To some extent an antenna is an antenna, regardless of frequency. Certain basic principles apply all across the r.f. spectrum, but the wavelength factor makes for a very large difference in practical problems encountered in building and erecting antennas, even within the v.h.f. range. Mainly for this reason, the explanatory material of Chapter 8 may not be enough for many v.h.f. enthusiasts who would like to try their hands at building their own beams.

Arguments in favor of building rather than buying are not greatly different for antennas than for other equipment we need for communication, except perhaps that fabrication of antennas may be more within the capabilities of the home craftsman than other equipment phases of the game. The hardest part of the job, the erection of the antenna, has to be done by the amateur in any case, so he is more likely to go the whole way and build the skyhook himself. Any able-bodied ham with a few simple tools can build and erect his own antennas, and usually he will enjoy the work, and learn much from it. Very likely he will stretch his dollars somewhat further too, for good antennas come rather high these days.

Material for the construction of arrays may be costly, depending on where you do your shopping, but there are many ways for the ingenious ham to adapt inexpensive items to his purposes. Serviceable beams have been made by coating wooden dowels with conducting paint, or even by wrapping them with aluminum foil. Neither of these techniques is recommended, but they are examples of what can be done in a pinch. Salvage should not be overlooked, if costs are really important. Lumber yards; electrical, welding, or plumbing-supply houses; metal-smelting companies; junkyards and surplus lots—these are just a few places in addition to the usual channels where we may come upon metal products that can be adapted to our needs.

Almost anything that is strong enough can be used for booms and supporting frames, whether it is insulating or conducting material, and there is no law requiring that elements be round in cross-section, so long as they are of a metal that is a reasonably good conductor. And though electrical rotation systems have become almost standard equipment in amateur antenna practice, there is much to be said for simple "armstrong system" rotating devices.

Rotating provision of some kind *is* important, however. In earlier days at least, much of the magic ascribed to a ham's first beam was actually the result of its having been the first antenna he ever put up high and in the clear, and equipped with some form of rotator. Even a simple dipole with these attributes is not a bad antenna, but the bigger and better a v.h.f. array is, the more it needs a rotator and some means of telling where the antenna is actually headed. If these requirements can be handled adequately by pulling on ropes and looking out the window, then there is no reason to be ashamed of doing it that way.

Because the band for which they are designed makes such a difference in the size and structural details of v.h.f. antennas, our practical constructional information is given by bands. It should be stressed that the following are

TABLE 9-I

Dimensions for V.H.F. Arrays in Inches

*Freq. (Mc.)**	*50**	*144**	*220**	*432**
Driven Element	111	38⅜	25 7/16	13
Change per Mc.*	2	¼	⅛	1/32
Reflector	116½	40½	26¾	13½
1st Director	105½	36⅝	24⅜	12 11/32
2nd Director	103½	36⅜	24	12 9/32
3rd Director	101½	36⅛	23⅞	12 7/32
1.0 Wavelength	236	81½	53⅜	27¼
0.625 Wavelength	149	51	33½	17
0.5 Wavelength	118	40¾	26 13/16	13⅝
0.25 Wavelength	59	20⅜	13⅜	6 13/16
0.2 Wavelength	47¾	16¼	10¾	5 7/16
0.15 Wavelength	35½	12¼	8	4

*Dimensions are for the *most-used* section of each band: 50 to 50.6 Mc., 144 to 145.5 Mc., 220 to 222 Mc., and 432 to 434 Mc. The element lengths should be adjusted for each megacycle difference in frequency by the amount given in the third line of the table. Example: if optimum performance is wanted much above 145 Mc., shorten all elements by about ¼ inch. For above 146 Mc., shorten by ½ inch. See text.

Element spacings are not critical, and table figures may be used, regardless of element lengths chosen. Parasitic element lengths are optimum for collinear arrays and small Yagis, having 0.2-wavelength spacing.

examples; they by no means cover the range of possibilities. Nor should it be inferred that, because a particular antenna is shown for only one band, it cannot be used, in principle at least, for others. These are ideas, to be adapted as the reader may see fit. The true ham will "take it from there."

To aid those who like to work strictly on their own, as far as materials and mechanical construction are concerned, Table 9-I gives the principal dimensions needed in building antennas for 50 through 450 Mc. Note that the most-used portion of each band is used for this information. Line 2 of the table is a change factor to be applied to table element lengths when other parts of a band are to be emphasized. Only element lengths are ordinarily this critical. Element spacings and phasing-line lengths can be left as given.

In the practical construction examples the dimensions of the original are given. Where the array is one that will be highly frequency-sensitive, as in a long Yagi, the portion of the band where the antenna works best is stated. The change factor of Table 9-I can be applied if some other band segment is to be favored. When in doubt, check back through Chapter 8, for basic information.

ANTENNAS FOR 50 MC.

Nearly everyone, at one time or another, has need for something simple and/or inexpensive for an antenna. This means a half-wave dipole of some sort, ordinarily. It can be any of the arrangements described in Chapter 8, so we will consider here only those that are most commonly employed in 50-Mc. work.

The Folded Dipole

Probably the most universally useful 50-Mc. dipole, all things considered, is the folded variety. It is broad in frequency response and not critical as to construction or adjustment. It can be made of a wide range of conductor sizes and materials, and it is adaptable to various mounting arrangements. It can be fed directly with 300-ohm balanced line (Twin-Lead or open-wire) or coax and a balun.

A folded dipole can be suspended from rope or wires or supported on a mast, depending on how the element is made. The center of the element can be grounded for lightning protection, or left floating electrically. The dipole of Fig. 9-1A can be made entirely of Twin-Lead, ready-made open-wire line, or of any wire you may have on hand. When made of Twin-Lead, it is occasionally tacked or taped to a wall, when a temporary and unobtrusive antenna is the principal requirement.

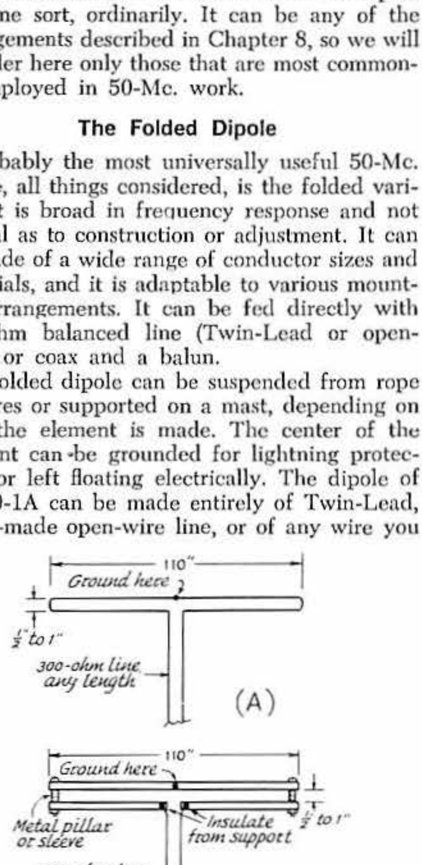


Fig. 9-1—Folded dipoles for 50 Mc. Either may be fed with 300-ohm line, or 72-ohm coax and a balun. Dipole A is made of wire or Twin-Lead; dipole B of any convenient size tubing. Either can be grounded at the center of unbroken portion.

Where it is to be mounted on a support such as a rotating mast, the dipole of 9-1B is preferable. The conductor size is not critical, except that both the broken and unbroken halves should be the same size. The unbroken portion can be attached to or run through a metal pipe or tubing support, in which case only a small cross-arm, or perhaps none at all, will be needed. The wire dipole, A, can be supported on a wooden "T", using vinyl-insulated screw eyes of the type sold for TV installations. The inner ends of the broken portion of either dipole, where the feed line connects, should be insulated from the support. The upper and lower parts of dipole B are connected at the outer ends by means of metal pillars or sleeves. If the tubing is sufficiently flexible it can be bent around as in dipole A, but this is usually done only in dipoles for the higher bands, where small conductors are stiff enough to be self-supporting.

The feed line can be Twin-Lead or open-wire. If the latter, the half-inch-spaced type is preferable to the 1-inch, as it is closer to 300 ohms impedance. Coax and a balun can also be used. If this is done, 72-ohm coax will give a better match than 52-ohm, though even with the latter, the s.w.r. will not be more than about 1.5 to 1, which is not serious.

The "J" Antenna

Center-fed systems like the folded dipole are well adapted to horizontal polarization, but the need for running the feed line perpendicular to the dipole for some distance makes them cumbersome for most vertical applications. The "J" system of Fig. 9-2 is more useful for vertical polarization. This is a vertical dipole with the matching arrangement at the bottom end, for convenience. It may be fed in various ways.

Antenna A has a balanced-line feed. This can be any impedance, as the point of connection is moved along the stub portion until a match is achieved. Antenna B is fed at the bottom end with coaxial line. This is a good approximation, if the antenna is to be erected and used without attention to matching. Though it may not be a perfect match, it will be close enough for practical purposes. Antenna C is for use where adjustment for match is desired. As in A, the coaxial line and balun are moved along the stub until an s.w.r. bridge in the line shows zero reflected power. The bottom end of the system can be grounded for lightning protection in either A or C. In B, the bottom of the stub portion can be grounded.

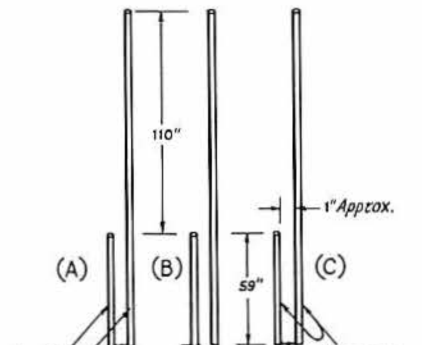


Fig. 9-2—Three versions of the "J" antenna, with dimensions for 50-Mc. operation. Grounding for lightning protection may be done as indicated by the ground symbol.

The basic idea of the "J" is that the stub should not radiate, but in actual practice it does, to some extent. This radiation interferes with that from the main portion of the antenna, and may result in raising the effective radiation angle. For this reason other matching methods are preferable, though the "J" has the virtue of simplicity. It is occasionally used as the driven element about which parasitic elements are rotated in a simple vertical Yagi. Except in B, it is preferable to run the feed line as nearly perpendicular to the stub as possible, for at least a quarter wavelength.

Coaxial and Ground-Plane Antennas

Particularly where the supporting structure is metal, the coaxial antenna, Fig. 9-3, and the ground-plane, Fig. 9-4, are superior to the "J" for omnidirectional vertical use. Both are intended primarily for coaxial feed, and the line can be run up inside the pipe mast, if one is used. The upper element is, in effect, an extension of the inner conductor of the coax. The radials of the ground-plane and the skirt or lower element of the coaxial are connected to the outer conductor, and to the support, if desired. The skirt of the coaxial antenna should be so connected only at the top; the rest of it must be insulated from the coax and the support. Element and skirt lengths are not critical in the coaxial antenna. About 54 inches should be suitable for work across the whole 50-Mc. band.

The ground-plane is perhaps the best all-purpose vertical antenna for coaxial feed. As the name implies, the horizontal radials simulate ground; consequently the impedance of the antenna is little affected by variation in height above actual ground and the nature of the supporting structure.

A simple and often-used version would be as shown schematically in Fig. 9-4A, except fed entirely with 52-ohm coax, without a matching section. The feed impedance of a ground-plane is low; of the order of 30 ohms, so there will be some mismatch when it is fed with 52-ohm line. The s.w.r. is under 2 to 1, however, and performance should be satisfactory.

Matching can be achieved in several ways. A simple method is shown at A, but it requires that the main line be 72-ohm coax. The quarter-wave *Q* section of 52-ohm line makes an almost perfect match, and it can be connected very simply. The lower end of the *Q* section and the upper end of the main line can be fitted with coaxial connectors, and a coaxial junction used between them. The length of 38 inches for the matching section is for solid-dielectric coax with a velocity factor of 0.65. Foam and other low-density insulaton will make the matching section longer. Means for checking resonant lengths are outlined in Chapter 8. See Fig. 8-25.

Another method for matching with 52-ohm line is to shorten the radiating element slightly, and then tune out the reactance so introduced by connecting a closed-end stub in parallel with the antenna. If the antenna assembly includes a coaxial T fitting at the base of the driven element, the stub can be connected at will, and

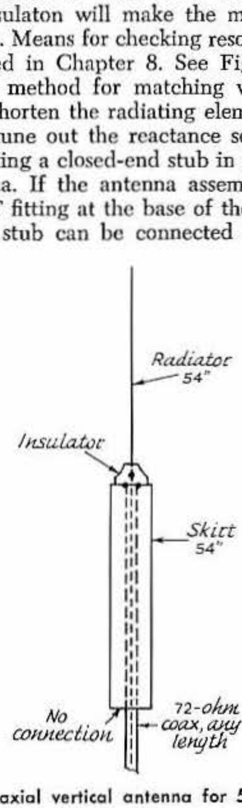


Fig. 9-3—Coaxial vertical antenna for 50 Mc. A supporting pipe mast can run up inside the skirt portion.

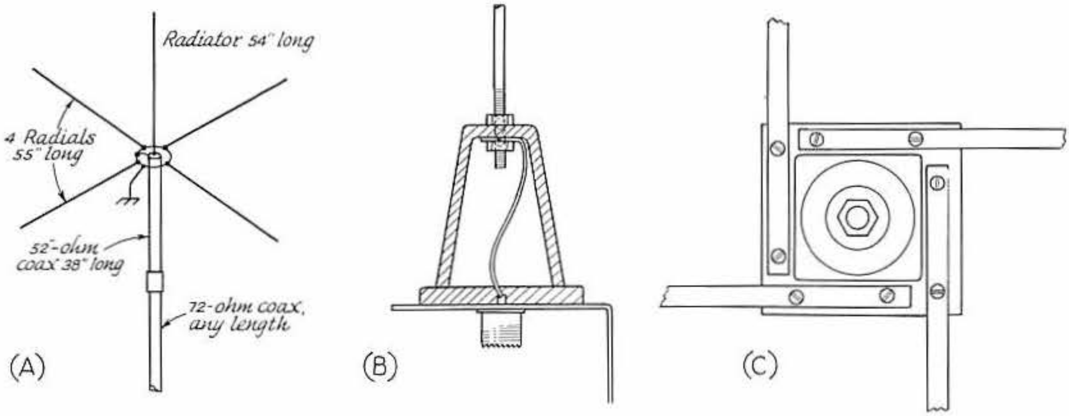


Fig. 9-4—The ground-plane antenna, shown with Q matching section, A. One method of making the antenna, with a metal mounting bracket and ceramic insulator is shown at B. Radials, omitted from B in the interest of clarity, are shown attached to the metal mounting bracket in C.

trimmed for length until a perfect match is achieved. With a 53-inch radiator the stub should be about 21 inches, if made with 52-ohm coax, but it is well to start with one somewhat longer and trim for match. Remember to short the far end of the stub each time a check is made.

Another matching trick with the ground-plane is to droop the radials downward, adjusting their angle below the horizontal until the antenna feed impedance becomes 52 ohms. This usually occurs at about a 45-degree angle. The antenna ceases to be a true ground-plane under these circumstances, but the method is often a satisfactory compromise. There will be some radiation from the radials in the drooping position, but this is not necessarily bad. Mixed polarization could be a "mixed blessing" under some propagation conditions.

Ground-planes can be made in many ways. One is shown in Fig. 9-4B. The vertical radiator is ¼-inch rod, threaded at the bottom end, held in the top of a ceramic standoff insulator with nuts above and below the top of the cone. Before the insulator is bolted to the angle bracket that serves as a mounting, a wire or flexible strip of copper is fastened under the nut. It is left long enough so that it can be soldered to the coaxial fitting mounted on the angle bracket, before the insulator is bolted in place.

Radials are omitted from B in the interest of clarity. They can be fastened to the angle bracket, as shown in 9-4C. The angle bracket can be fastened to any vertical support, of wood or metal. A metal support preferably should be grounded independently of the outer conductor of the coax, for lightning protection.

Omnidirectional Horizontals

Often it is desirable to maintain uniform field strength in all directions about the station when horizontal polarization is used. In 50-Mc. work this is usually accomplished with some version of the halo antenna. Basically, the halo is a half-wave dipole, bent around into a circle or some other shape that will give it fairly uniform radiation in the horizontal plane. Any of the common matching systems described in Chapter 8 can be used with the halo.

Where the halo is used for 50-Mc. mobile work the total length of the ring is usually reduced by capacitive loading between the ends, as shown in Fig. 9-5. The circumference of the ring so formed is usually 60 to 70 inches.

The gamma matching system is convenient for mobile halos, as it permits matching to coaxial line and the use of an unbroken driven element. The step-up type of folded dipole is also used in halos, though the mechanical work involved has limited this type of feed mainly to manufactured antennas. A three-ring model of this kind has long been a fixture on the 6-meter mobile scene.

The capacitively loaded halo is a high-Q device, and it must be tuned with care or it will be all but useless. Usually some provision is made for varying the end-to-end capacitance, C_1. If a gamma matching system is used, the series capacitor, C_2, and point of connection of the gamma arm to the element should be made variable, at least temporarily. Adjust these and the tuning capacitor, C_1, for minimum reflected power in an s.w.r. bridge connected in the transmission line. This is made easier if the element is first resonated roughly at the middle of the desired operating range with C_1, checking resonance with a grid-dip meter. The best point for coupling the dip-meter coil is near the feed point, just to the left of the coaxial line ground point in Fig. 9-5. Variation of the effective capacitance of C_1 is usually done by mounting a small disk on an adjusting screw, equipped with lock nuts, to one of the plates, and then adjusting its position with respect to the other plate.

Halos can be stacked one above the other in a vertical line, to lower radiation angle and build up gain. This is done occasionally for fixed stations where omnidirectional coverage and something better than a single halo are required. Matching and feeding can be done in the same ways as with any two 50-ohm antennas. See Chapter 8, and below. The optimum spacing for stacked halos is ½ to ⅝ wavelength.

Two 50-Mc. halos adjusted individually for 50-ohm feed can be stacked physically ⅝ wavelength apart by connecting them with half-wave (77-inch) sections of 52-ohm coax, with a T fitting. A 38-inch 52-ohm Q section at this point will then match a 72-ohm transmission line. Specific halo designs follow in the mobile section, later in this chapter. For more stacking details see Fig. 9-46.

The turnstile antenna, shown for 144-Mc. use in Fig. 9-47, can be adapted readily to 50-Mc. service. It is larger physically than the halo, and it should provide slightly more gain and considerably broader frequency response. A stacked turnstile system makes a very good omnidirectional antenna for home-station use. Stacking methods similar to those outlined for halos and other coax-fed antennas can be employed with turnstiles.

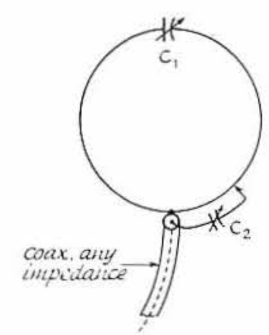


Fig. 9-5—The halo antenna is a half-wave radiator bent into circular shape for nearly uniform radiation pattern. Capacity plates, C_1, permit use of a small radiator for 50 Mc. Gamma match and series capacitor, C_2, are for coaxial feed.

YAGI ARRAYS FOR 50 MC.

The Yagi antenna is almost ideally suited to 50-Mc. operation. Usually only a relatively small portion of the band is covered so the Yagi's limited frequency response presents no problems, and arrays having gains of up to 10 db. are easily built and erected. Except under the most severe weather conditions, rotation of 50-Mc. Yagis of up to at least 6 elements can be handled with inexpensive TV-type rotators, provided means are taken to prevent the entire weight of the structure from bearing on the rotator driving mechanism. Some rotators have thrust bearings available as accessories, for this purpose.

3-Element Lightweight Array

The 3-element 50-Mc. array of Fig. 9-6 weighs only 5 pounds. It uses the closest spacing that is practical for v.h.f. applications, in order to make an antenna that could be used individually or stacked in pairs without requiring a cumbersome support. The elements are half-inch aluminum tubing of $\frac{1}{16}$-inch wall thickness, attached to the 1¼-inch dural boom with aluminum castings. The mounting method of Fig. 9-16 is also usable. By limiting the element spacing to 0.15 wavelength, the boom is only 6 feet long. Two booms for a stacked array can thus be cut from a single 12-foot length of tubing.

The folded-dipole driven element has No. 12 wire for the fed portion. The wire is mounted on ¾-inch cone standoff insulators and joined to the outer ends of the main portion by means of metal pillars and $\frac{6}{32}$ screws and nuts. When the two halves are pulled up tightly and wrapped around the screws, solder should be sweated over the nuts and screw ends to seal the whole against weather corrosion. The same treatment should be used at each standoff. Mount a soldering lug on the ceramic cone and wrap the end of the lug around the wire and solder the whole assembly together. These joints and other portions of the array may be sprayed with clear lacquer as an additional protection.

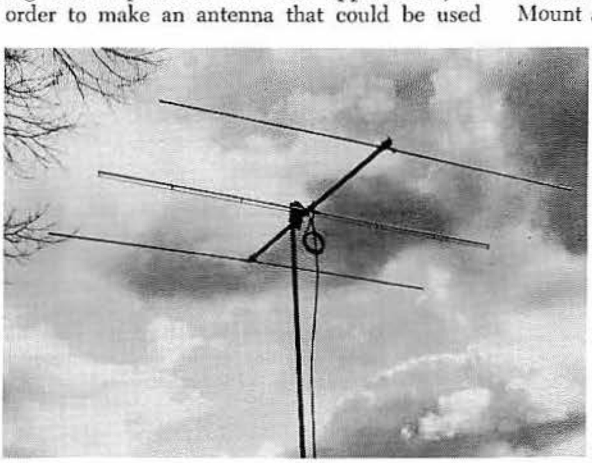


Fig. 9-6—A lightweight 3-element 50-Mc. array. Feeder is 52-ohm coax, with a balun for connection to the folded dipole driven element. Balun loop may be coiled as shown, or taped to the supporting pole.

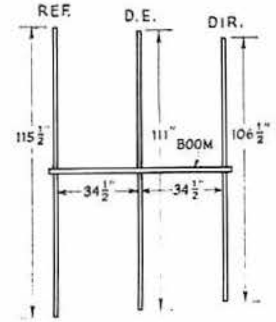


Fig. 9-7—Dimensions of the 3-element array of Fig. 9-6, for working in the lower portion of the 50-Mc. band. Driven elements are ½-inch aluminum tubing. The folded dipole driven element uses No. 12 wire for the fed portion.

The inner ends of the fed section are 1½ inches apart. Slip the dipole into its aluminum casting, and then drill through both element and casting with a No. 36 drill, and tap with 6/32 thread. Suitable inserts for mounting the standoffs can be made by cutting the heads off 6/32 screws. Taper the cut end of the screw slightly with a file and it will screw into the standoff readily.

Cut the element according to Fig. 9-7 for operation in the first megacycle of the band. Shorten all elements by 2 inches for each higher megacycle. The reflector and director are approximately 4 per cent longer and shorter, respectively, than the driven element. The closer spacing of the parasitic elements (0.15 wavelength) makes this deviation from the usual 5 per cent desirable.

The folded dipole gives the single 3-element array a feed impedance of about 200 ohms at its resonant frequency. Thus it may be fed with a balun of the type shown in Fig. 8-25, using 52-ohm coax. A gamma-matched dipole may also be used, suggested construction being as shown in Fig. 9-9. If the gamma match and 72-ohm coax are used, a balun will convert to 300-ohm balanced feed, and Twin-Lead or 300-ohm open-wire TV line may be used for the main transmission line. The dimensions of Fig. 9-7 are for optimum performance at 50.5 Mc. The array will show good performance and a fairly low standing-wave ratio over the range from 50 to 51.5 Mc.

A closeup of a mounting method for this or any other array using a round boom is shown in Fig. 9-8. Four TV-type U bolts clamp the horizontal and veritcal members together. The metal plate is about 6 inches square. If ¼-inch sheet aluminum is available it may be used alone, though the photograph shows a sheet of 1/16-inch stock backed up by a piece of wood of the same size for stiffening. Tempered Masonite is preferable to wood, as it will stand up better in weather, particularly if lacquer is sprayed on all surfaces.

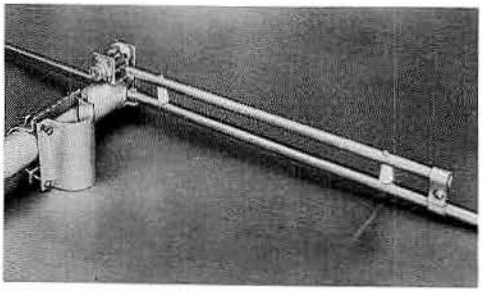


Fig. 9-9—Typical gamma match construction. The variable capacitor in series with the matching arm should be mounted in an inverted plastic cup, or otherwise protected against moisture. The arm should be about 14 inches long for 50 Mc., 6 inches for 144 Mc.

High-Performance 4-Element Array

The 4-element array of Fig. 9-10 was designed for maximum forward gain, and for direct feed with 300-ohm balanced transmission line. The parasitic elements may be any diameter from ½ to 1 inch, but the driven element should be made as shown in the sketch. For a ½-inch driven element use the information for the antenna of Fig. 9-6. The spacing between

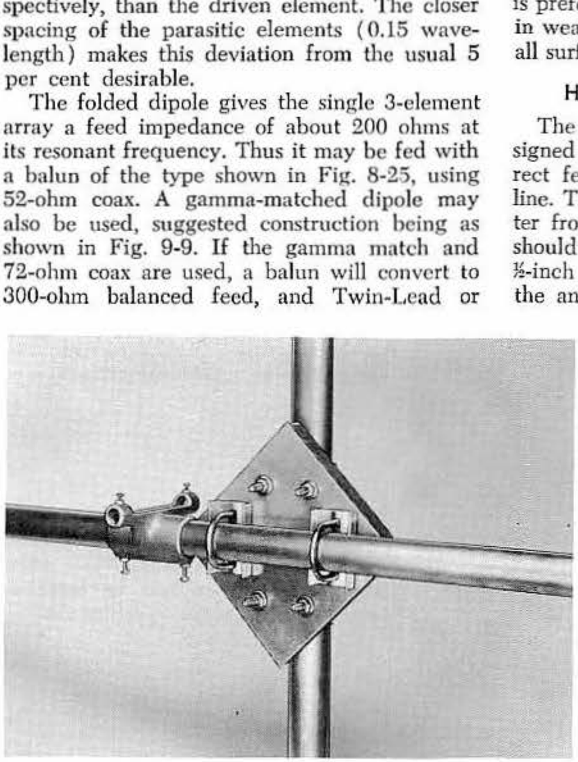


Fig. 9-8—Closeup view of the boom mounting for 50-Mc. arrays. A plate of alumiunm about 6 inches square is backed up by wood or Masonite. TV-type U clamps hold the boom and vertical support together at right angles. At the left of the mounting assembly is one of the aluminum castings for mounting the beam elements.

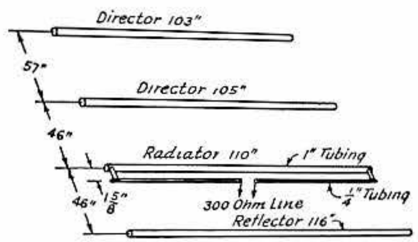


Fig. 9-10—Dimensions of a 4-element 50-Mc. array having maximum forward gain at 50.5 Mc. The folded dipole details are for 300-ohm balanced feed.

driven element and reflector, and between driven element and first director, is 0.2 wavelength. Between the first and second directors the spacing is 0.25 wavelength.

The same general arrangement may be used for a 3-element array, except that the solid portion of the dipole should be ¾-inch tubing instead of 1-inch. The boom length would then be about 8 feet.

With the element lengths given, the array will give nearly uniform response from 50 to 51.5 Mc., and usable gain to about 52 Mc.

If a shorter boom is desired, the reflector spacing can be reduced to 0.15 wavelength and both directors spaced 0.2 wavelength, with only a slight reduction in forward gain and bandwidth. A slight modification for mounting on a 12-foot wooden boom is shown in Fig. 9-11. Also included is a method of attaching the boom to a pipe support.

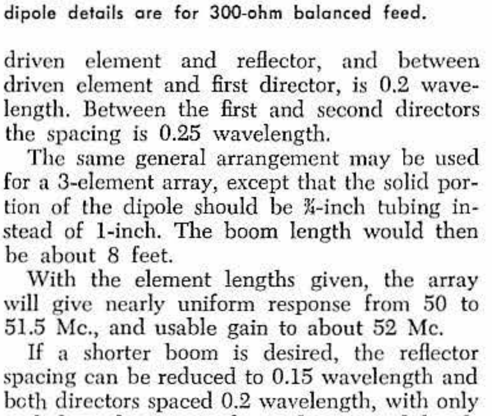


Fig. 9-11—Suggested construction for an inexpensive 4-element array for 50 Mc. using a wooden boom. Dimensions are a slight modification of the optimum given in Fig. 9-10, to fit a 12-foot boom. Mounting arrangement is for clamping to a pipe mast.

5-Element 50-Mc. Array

As aluminum or dural tubing is often sold in 12-foot lengths, this dimension may impose a practical limitation on the construction of a 50-Mc. beam. A 5-element array that makes optimum use of a 12-foot boom may be built according to Fig. 9-12. If the aluminum-clamp method of mounting elements shown in Fig. 9-16 is employed, the weight of a 5-element beam can be held to under 10 pounds.

The gamma match and coaxial line are recommended for feeding such an array. A folded dipole similar to that used in the 3-element array will provide an approximate match to 52-ohm coax and a balun, but the gamma system is preferable, as it permits adjustment for exact match at the favored frequency range.

Gain and bandwidth of this compromise design are both slightly below optimum, but it represents effective use of a 12-foot boom. It will have a bit more gain than a 12-foot 4-element array, but will be somewhat more critical in frequency response.

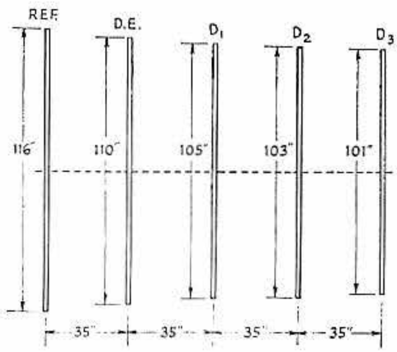


Fig. 9-12—Five-element 50-Mc. Yagi for a 12-foot boom. Dimensions are for working over the lower 2 megacycles of the 50-Mc. band.

Long Yagis for 50 Mc.

Once we go beyond 12-foot booms we become concerned mainly with getting the best possible performance. Graduated element spacing, discussed in Chapter 8, is generally employed, and boom lengths become considerable for 50-Mc. arrays. Yagis of 6 elements or more can still be light in weight and relatively easy to handle, however, if we design with these points in mind.

The 6-element arrays of Figs. 9-13 and 9-20 are examples. The 20-foot boom can be made of light aluminum TV masting. This comes in 10-foot lengths which telescope one inside the other for about 6 inches. The joint can be held firm with self-tapping screws. Any light boom more than about 12 feet long should be braced to keep it in alignment, preferably from above the boom, as shown in Figs. 9-13 and 9-16.

Dimensions in Fig. 9-14 are for operation in the first megacycle of the band. Set up in this way, the array can be adjusted for perfect match at 50.3 and it will show an s.w.r. under 1.7 to 1 from 50 to 50.6 Mc. Slightly more bandwidth can be achieved by making the directors one-half inch shorter than shown. Additional directors should be spaced 70 inches

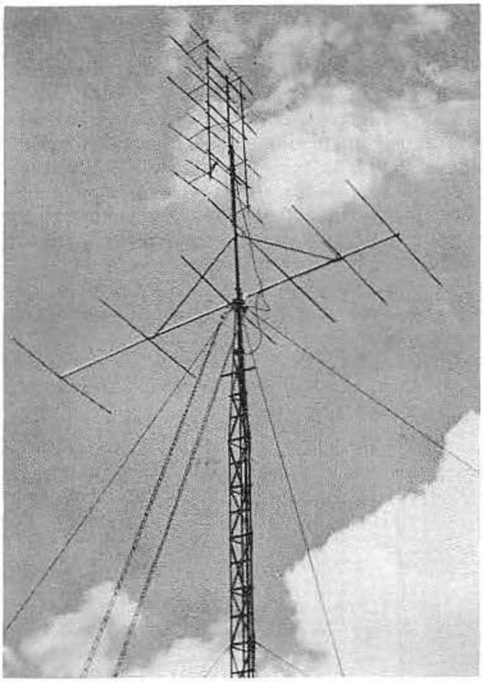


Fig. 9-13—A 6-element Yagi for 50 Mc., with optimum spacing for forward gain given in Fig. 9-14. Boom is 20 feet long. Antenna at the top is 16-element all-metal collinear array for 144 Mc.

apart. They may be similar to D4, or each one inch shorter for greater bandwidth.

Elements are mounted in the same way as for the 3-element job, using aluminum castings or sheet-metal clamps. The center support is the same as in Fig. 9-8. Elements can be run through the boom and clamped, as shown in Fig. 9-16, but the aluminum-casting method of mounting is stronger. If suitable castings can be found, they are recommended for light boom materials. Matching is by means of a coaxial gamma arrangement shown in Fig. 9-15. Once adjustment is completed the open end of the gamma arm can be wrapped with plastic tape, and it and the joint between the sliding and stationary parts of the gamma capacitor sprayed with clear lacquer. If anyone doubts the ability of this arrangement to withstand weather, the original was used for over seven years, and was working nicely when replaced by that in Fig. 9-20.

The main gamma arm is cut from the same material as the elements. It is suspended parallel to the driven element by means of two 1-inch ceramic standoffs and 4 sheet aluminum clips, as shown in the photograph. The ½-inch tube is 15 inches long. Its inner end is connected to the inner conductor of a coaxial fitting, which is mounted on a small bracket screwed to the boom casting. Holes are drilled and tapped in the casting to take two 6/32 machine screws for mounting the bracket.

The sliding arm that is the movable element of the coaxial capacitor is made of ¼-inch tubing or rod, about 14 inches long. It is maintained coaxial with the main arm by means of two polystyrene bushings. One is force-fitted to the end of the rod that goes inside the main arm. The other is fitted tightly into the far end of the main arm, but reamed out to permit the movable rod to slide freely in and out. These bushings can be made from ⅝-inch polystyrene rod, or they can be fashioned easily from small polystyrene coil forms. The National PRC-1 form is ideal for the purpose. It fits tightly over the ¼-inch rod and slides freely inside the ½-inch arm. The bearing at the end of the arm where the adjusting rod projects was made by cutting the bottom off one of the PRC-1 forms and drilling out the inside so that it would pass the rod freely. It is shimmed up with plastic tape to a sufficient thickness to make it a tight fit inside the main arm. It is slipped over the rod and then pressed into place in the end of the arm.

A clip of sheet aluminum makes contact between the driven element and the sliding rod. Be sure that all surfaces at the points of contact are completely clean, as solid low-resistance electrical contact is of utmost importance here.

Proper adjustment of the gamma match requires an s.w.r. bridge. If the work cannot be done with the beam in the position in which it is to operate, set it up temporarily with the boom vertical and the reflector close to the ground. Insert the bridge in the line near or at the antenna. If more than a few watts of power is present in the line, adjustment of the gamma will best be done with gloves on the hands, to

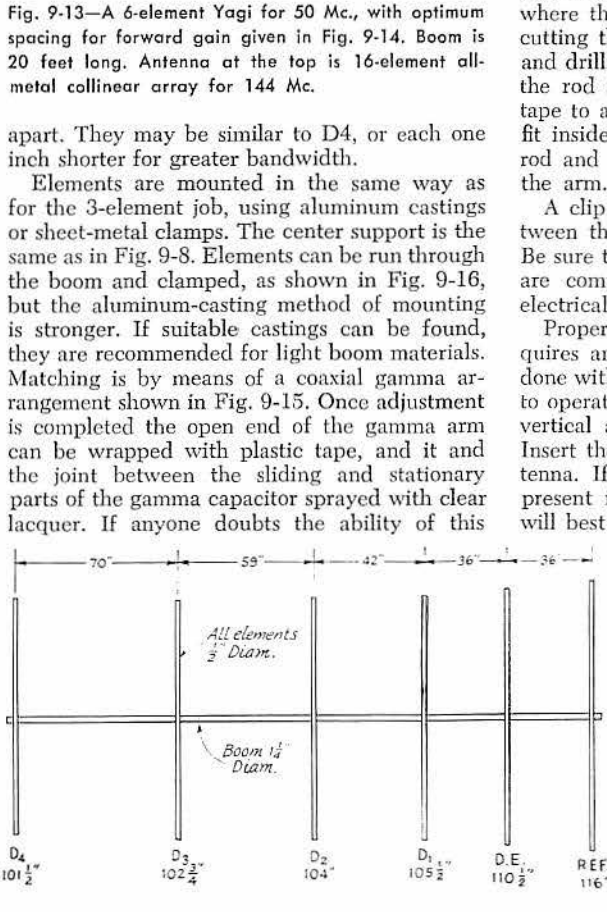


Fig. 9-14—Dimensions of the 6-element 50-Mc. array. If the boom length is limited to exactly 20 feet, reduce spacing of D_3 and D_4 by 2 inches each. Dimensions are for the low megacycle of the band.

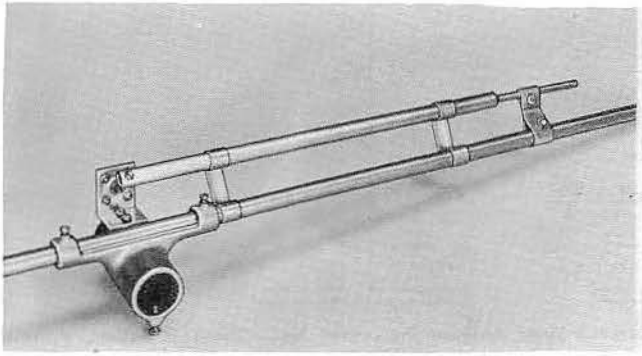


Fig. 9-15—Gamma matching section for the 6-element array, using a coaxial variable capacitor. The sheet aluminum clip at the right and the length of the small rod protruding from the arm are adjusted for minimum reflected power. The movable rod element of the capacitor is about 14 inches long, and is insulated from the fixed portion, of ½-inch tubing.

prevent r.f. burns. The operation is twofold; both the point of connection and the value of the series capacitor must be adjusted. Start with the clip set about 16 inches out from the boom, with enough tension on the clip to insure a good electrical connection. Adjust the capacitance and the point of connection for zero reflected power at the midpoint of the frequency range you want to work over effectively. If you are interested only in the first 600 kc. or so of the band, use 50.3 for the adjustment frequency. For good coverage of 50 to 51 Mc., use 50.5. In the latter case, the s.w.r. should be below 2 to 1 over the 1-megacycle range.

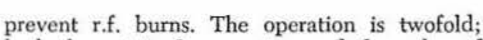


Fig. 9-16—Method of mounting elements through a metal boom, left, and of bracing the boom to the vertical support, right. Suspension bracing is recommended for long booms. Shorter ones can be braced below the boom. A one-piece wrap-around clamp on the boom takes the angle brace.

Stacking 50-Mc. Yagis

Lowering of the radiation angle effected by stacking two 50-Mc. Yagis one above the other can result in quite marked improvement in signal level, particularly on long paths. The gain is achieved without appreciable sharpening of either frequency response or horizontal pattern. The same gain obtained through adding elements in a single bay makes for a sharper horizontal pattern and some sacrifice in bandwidth. Choice between the two methods thus becomes a matter of determining what kind of work one wishes to do best. Local activity levels and distribution may have some bearing on the decision.

3-Over-3 The bays of the 3-over-3 of Fig. 9-17 are like the array of Fig. 9-6. Spacing is ⅝ wavelength, or about 12 feet. This is convenient (see Figs. 8-30C and 9-18) since it allows use of an electrical one-wavelength of coax for phasing. This can be handled in several ways. In Fig. 9-18A the main line is to be 52-ohm coax. This is matched to the two 52-ohm bays by ¼-wave and ¾-wave Q sections of 72-ohm coax, with baluns of the same material. Note that with this off-center feed the driven elements are connected in opposite polarity.

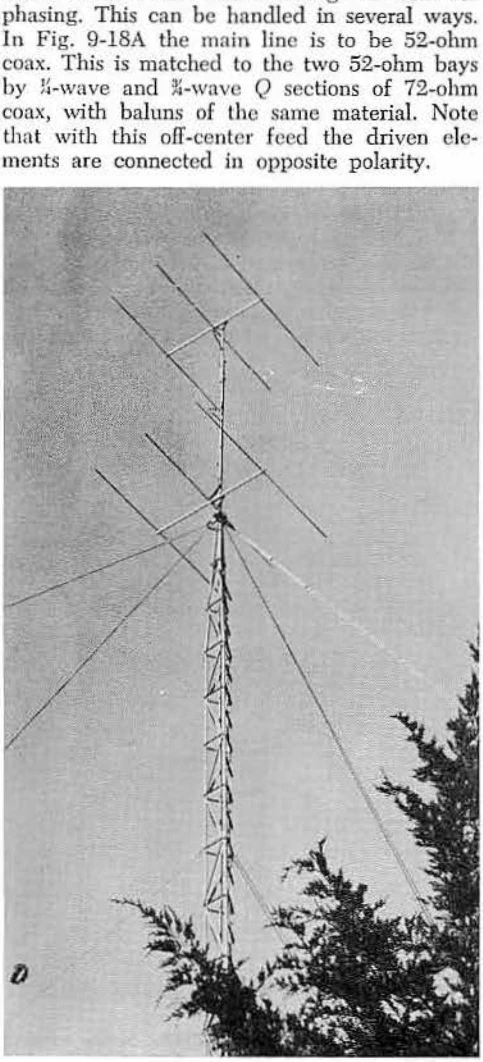


Fig. 9-17—Stacked array for 50 Mc., using two bays like Fig. 9-6. Phasing line and rotatable section of the main transmission line are of coax. Various methods of feed are described in the text.

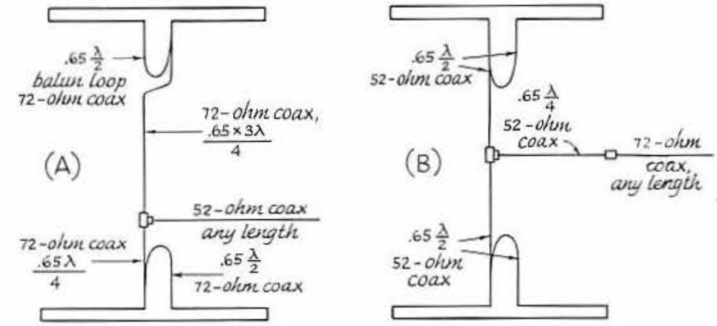


Fig. 9-18—Two methods of feeding stacked Yagis designed for coaxial feed. Off-center method, A, requires that the bays be connected in opposite polarity, to keep currents in phase. Feeding at center, B, allows both driven elements to be connected in the same manner. Method A is for 52-ohm main feed, B for 72-ohm. Sections in A can be any odd multiple of a quarter wavelength; in B any of a half wavelength.

The arrangement at B uses two half-wave sections of 52-ohm coax and baluns, with the T fittings at the midpoint. A Q section of 52-ohm line is then used to match the low center impedance to a main line of 72-ohm coax. A slight mismatch is involved here, but it is not enough to affect operation of the system to any practical degree.

In the original version a one-wavelength section of coax terminated in a balun at the low end was used as the rotating portion of the feed line with the phasing as in Fig. 9-18B. The balun then worked into a balanced Q section made of ¼-inch tubing 59 inches long, spaced ¾ inch center-to-center, and then into 450-ohm open-wire line for the main run to the station.

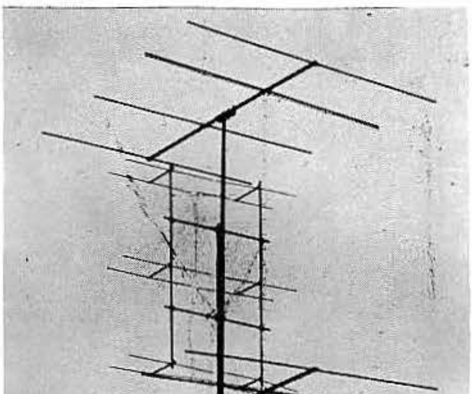


Fig. 9-19—A 4-over-4 stacked 50-Mc. array using a balanced phasing line. Matching is similar to Fig. 8-30B, in which the phasing line operates as a double Q section, matching a 300-ohm main line to two phased 300-ohm bays. Antenna between is a 12-element 144-Mc. collinear of all-metal construction.

4-Over-4 Two 50-Mc. arrays designed for 300-ohm feed can be stacked conveniently a half wavelength apart. This does not give quite the gain obtainable with wider spacings, but the pattern is very clean and the feed system is simple. The balanced phasing line between the two bays should have an impedance of about 400 ohms, if the main line run is to be 300. Two No. 12 wires spaced one inch, or half-inch-spaced TV line, may be used. In Fig. 9-19, a 300-ohm main line of tubular Twin-Lead was connected at the midpoint of the phasing line. The transmission line could also be 72-ohm coax and a balun. If 52-ohm coax is to be used with a balun, the phasing line should, theoretically at least, have slightly closer spacing, to give an impedance of about 330 ohms. In practice, ½-inch-spaced TV line will come close enough to providing an adequate match.

The 4-element bays that make up the array of Fig. 9-19 are similar to Fig. 9-10, except that both directors are spaced 0.2 wavelength (46 inches) and the reflector 0.15 wavelength (36 inches). Booms can be either metal or wood. This array was used very successfully for several years of intensive 50-Mc. DX work.

144-Over-50 Stacking of arrays for 50 and 144 Mc., and a method of using a single low-loss transmission line, are illustrated in the two-band system of Fig. 9-20. The 4-bay 144-Mc. array will be described in the next section. The 6-element 50-Mc. Yagi is very similar to the one already described, except for the simplified gamma matching system and special attention to lightweight element design.

The 50-Mc. elements have center sections of half-inch aluminum tubing making up about half their total length. Thin-walled fuel-line tubing inserts in each end keep the total element weight down, and provide a means of adjustment of length, if the builder wishes to experiment with tuning. The entire element can be made of the heavier tubing, though the arrangement described is lighter, and has a bit lower

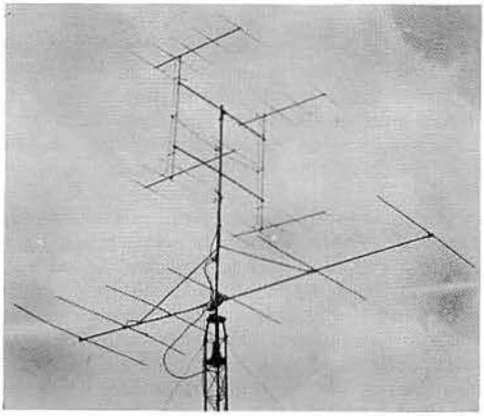


Fig. 9-20—All metal arrays for 50 and 144 Mc. All parts of both beams can be assembled readily with ordinary hand tools. In this installation the two beams are fed from a single feed line, with a waterproofed coaxial switch at the top of the tower permitting selection of the desired array from the operating position.

wind resistance. Some shopping around in surplus houses, aluminum smelting places and hardware and plumbing supply stores, as well as the usual aluminum tubing sources, will turn up several tubing size combinations that can be used in this way. The ends of the larger tubing can be slitted with a hacksaw to a depth of about three inches, and then tightened onto the smaller material with a wrap-around clamp. The elements are clamped to the boom as pictured in Fig. 9-21. Details of clamps used in the 50- and 144-Mc. arrays are given in Fig. 9-22.

Metal castings for mounting elements are fine if available, but they are getting hard to find. The sheet aluminum clamps will do, and they're not hard to make. Elements can be run through the boom, and held in place as shown in Fig. 9-16. These were made of 3/64-inch sheet aluminum, which can be bent easily by hand. Any heavier stock is good, if you have suitable bending facilities. Use of self-tapping screws to hold

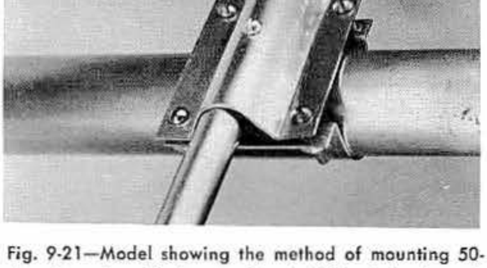


Fig. 9-21—Model showing the method of mounting 50-Mc. elements on the boom without drilling holes through the latter. For strongest permanent assembly, self-tapping screws should fasten the sheet-metal clamps in position.

components in alignment, as shown in Fig. 9-21, is recommended with thin clamp stock.

The lips of the clamps should be bent upward at right angles first. Forming the "U" is started by placing the tubing in a vise in a vertical position, and bending the clamp around it. The actual U shape is achieved by opening the vise to slightly more than the width of tubing-plus-clamp, placing the clamp U-down loosely in the vise with the tubing lying in it, and then tapping the tubing lightly with a hammer. Alignment of the holes in the clamps is not fussy, and if they are drilled slightly larger than needed to pass the screws there will be no assembly problem. We used a No. 22 drill and 6-32 screws. The nuts should be pulled up only tightly enough to hold the assembly firmly together. Check the nuts after the array has been in use for a few days and tighten as necessary.

The gamma method is about as simple as you can get: the coax is merely brought along the boom to the driven element, bent at right angles, and run out far enough to match the antenna

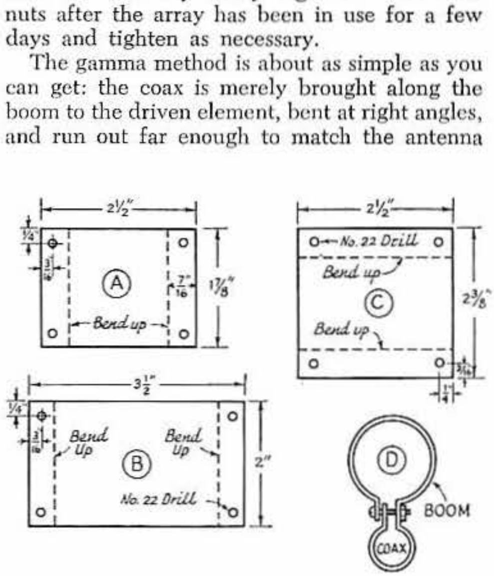


Fig. 9-22—Dimensions of aluminum plates used to make the assembly clamps in the v.h.f. arrays. Sheet metal should be 3/64 inch or thicker. Two A-type clamps are needed for joining ¾-inch tubes at right angles for assembling the frame of the 144-Mc. array. One B and one C are needed to mount a ½-inch element on a 1¼-inch boom, as in Fig. 9-21. The "figure 8" clamp, D, made from a ½-inch wide strip approximately 6 inches long, is used to ground the coax to the 50-Mc. boom.

impedance when fed through a 100-pf. fixed capacitor. The point of connection was found experimentally, though 20 inches and 100 pf. should do. Put an s.w.r. bridge in the line and move the connection along the element, for minimum reflected power. The outer conductor of the coax should be grounded to the boom at about 54 inches from the capacitor end. Strip a narrow band of the outer covering off, and fasten it to the boom with the "figure 8" clamp of Fig. 9-22D. Waterproof by wrapping with plastic tape and spraying with Krylon. Treat the capacitor similarly. It must stand high r.f. current. The Centralab 8505-100N is adequate. A variable may be used if mounted in a weatherproof box.

144-MC. ANTENNAS

Though some information in the 50-Mc. section may be useful to the builder of 144-Mc. arrays, the roughly 3-to-1 difference in size tends to make construction ideas for the two bands mutually exclusive. The 2-meter antenna nearly always has more elements than the one for 6, for two good reasons: size permits it and performance demands it.

Though gain over a dipole comes progressively easier as we go to higher frequencies and smaller elements, overall communications efficiency does not. Regardless of frequency, the *physical size* of the antenna is what really counts in determining how well a station will work out. A 5-element Yagi, properly designed, will have the same gain whether built for 50 or 144 Mc. It is accepted as quite a good antenna on 6, but if everyone used nothing larger on 2 our results on that band would be dismal indeed. Thus we find long Yagis and large collinear arrays in common use on 144 Mc. and higher bands, but relatively rare on 50 Mc.

Frequency response may be important on 144, too. Most 50-Mc. activity is concentrated near the low end of the band, but we spread out more in the 144-Mc. band. Its ability to work over a relatively wide frequency range makes the collinear a good choice for many 2-meter men, while those who want optimum performance in one narrow segment probably will go for the long Yagi.

COLLINEAR ARRAYS

Two collinear systems for 144 Mc. and higher bands are shown in Figs. 9-23 and 9-24. Either can be fed directly with 300-ohm balanced line, or with coaxial line and a balun. The actual feed impedance depends on many factors, but if no means of matching is used the s.w.r. on the main line will never be very high. For precise matching, use of the universal stub (Fig. 8-18D) is recommended.

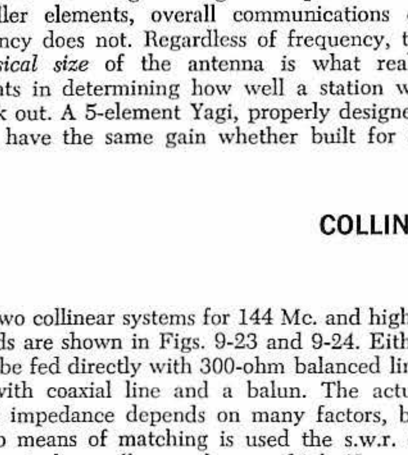


Fig. 9-23—Schematic drawing of a 12-element collinear array for 144 Mc. that may be fed with coaxial line and a balun, or 300-ohm balanced line. The supporting frame of Fig. 9-25 was intended for use with this type of array.

Feed impedance can be controlled to some extent by varying the spacing between the driven elements and the reflectors. In the 16-element array, Fig. 9-24, the impedance tends to be on the low side of 300 ohms, so the reflectors are spaced 0.2 to 0.25 wavelength behind the driven elements. In the 12-element system, Fig. 9-23, the impedance is higher, due to the lesser number of elements and connection of the main line at the inner ends of the middle pair instead of at the midpoint between two pairs, so the reflectors are spaced 0.15 wavelength to bring the impedance down. There is little gain difference with reflector spacings from 0.15 to 0.25 wavelength.

A supporting frame may be made of wood or metal, if elements are supported at their centers, and no insulating mounts will be needed. It is best to keep the supporting structure in back of the plane of the elements insofar as possible, and to avoid use of insulating material near the element ends. All-metal construction is illustrated in Figs. 9-25 through 9-27. All-wood

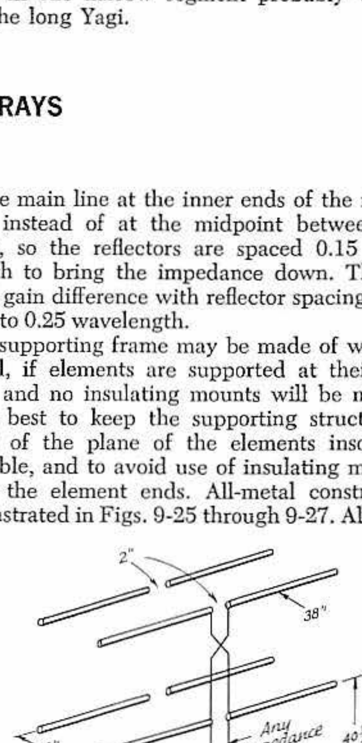


Fig. 9-24—A 16-element collinear array similar to the 12-element of Fig. 9-23, except that the feedpoint is midway between the middle pairs of elements. Wider spacing of the reflectors results in nearly the same feed impedance. An example of this array is shown above a 50-Mc. Yagi in Fig. 9-13.

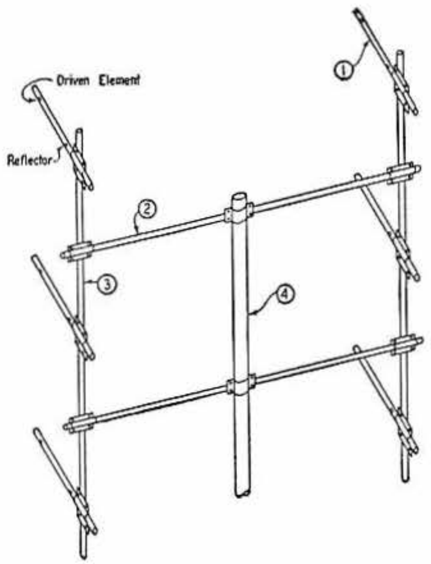


Fig. 9-25—Supporting framework for a 12-element 144-Mc. array of all-metal design. Dimensions are as follows: element supports (1) ¾ by 16 inches; horizontal members (2) ¾ by 46 inches; vertical members (3) ¾ by 86 inches; vertical support (4) 1½-inch diameter, length as required; reflector-to-driven-element spacing 12 inches. Parts not shown in sketch: driven elements ¼ by 38 inches; reflectors ¼ by 40 inches; phasing lines No. 18 spaced 1 inch, 80 inches long, fanned out to 3½ inches at driven elements (transpose each half-wave section). The elements and phasing lines are arranged as shown in Fig. 9-23.

construction is used in 432-Mc. arrays described later.

Elements should be rigid enough so that light tension on the phasing lines will not bend them appreciably. Aluminum or dural tubing ¼ to ⅜ inch in diameter is commonly used, and ¼-inch aluminum rod is good. Frequency response

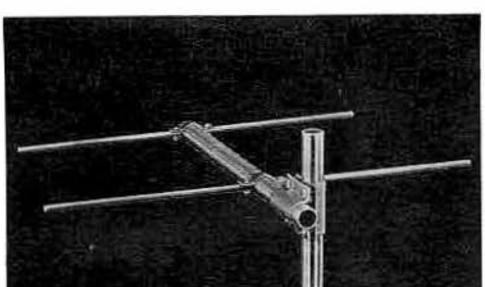


Fig. 9-26—Model showing method of assembling all-metal arrays, using clamps detailed in Fig. 9-27.

of the collinear array is broad enough so that element lengths and other dimensions can be taken from the figures of Table 9-I.

Large collinear arrays should be kept to a maximum of 8 driven elements per set of phasing lines. See Fig. 8-13 and the 432-Mc. collinear of Fig. 9-67 for recommended methods of feeding arrays of more than 8 driven elements. Two 16-element 2-meter arrays side by side are shown in Fig. 9-28. This is a duplicate of the 32-element setup used to collect the data of Fig. 8-11, on the effects of spacing between bays. Increasing the bay spacing makes a large array that is hard to handle mechanically, but a very worthwhile improvement in gain and pattern sharpness results.

The feed impedance at the center of the phasing line between two such bays is roughly half that of one bay alone. A Q section with variable spacing, or the universal stub, Fig. 8-18C and D, will provide for matching.

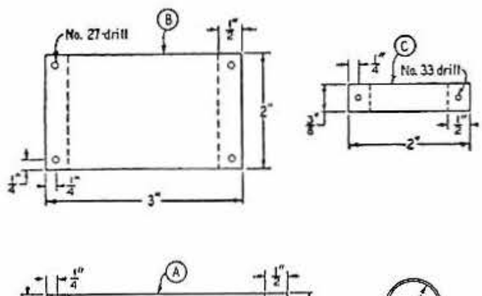


Fig. 9-27—Clamps used for assembling all-metal collinear arrays. A, B and C are before bending into U shape. Right-angle bends should be made first, along dotted lines, then the plates may be bent around pipe of the proper diameter. Sheet stock should be 1/16-inch or thicker aluminum.

All-Metal Construction

Collinear arrays of all-metal design can be very light in weight, yet rugged enough to withstand extreme weather hazards. The 16-element array of Fig. 9-13, built according to Figs. 9-24 through 27, survived four severe winters, mounted 70 feet in the air on a windswept New England hilltop, yet it was in good condition when taken down. It weighed less than 10 pounds and was relatively easy to handle. The entire frame, except for the 1½-inch vertical support, was made of ¾-inch aluminum tubing. Elements are ¼-inch tubing, center-mounted. Phasing is done with open-wire TV line soldered to lugs bolted to the inner element ends.

The rotating portion of the transmission line was 300-ohm tubular Twin-Lead, brought to an

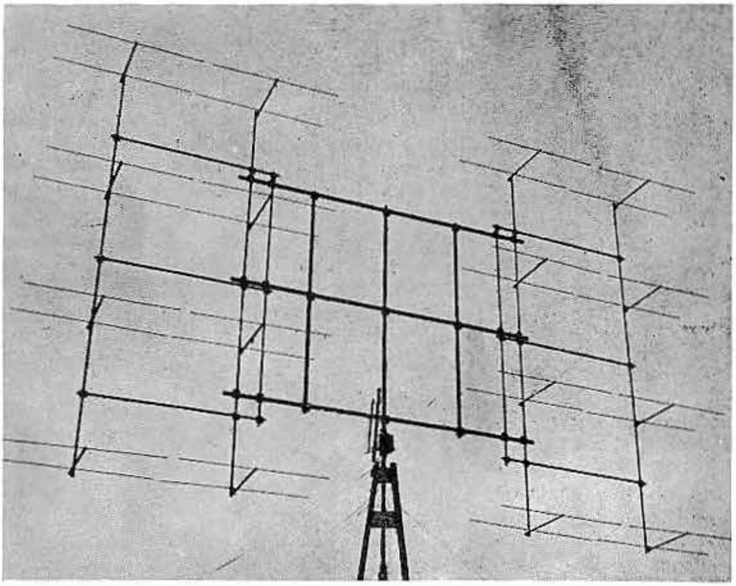


Fig. 9-28—Two 16-element all-metal arrays, arranged for adjustable spacing between bays. The patterns of Fig. 8-11 were made with a duplicate of this array.

insulating support just below the tower bearing. Here it joined the main transmission line, which was open-wire pulled up tight on strain insulators at each end. As pointed out in Chapter 8, straight unsupported runs are recommended for low radiation loss with open line. This installation had 125 feet, yet it performed in outstanding fashion.

YAGIS FOR 144 MC.

For a small antenna with appreciable gain, a Yagi is the usual choice at 144 Mc., as well as 50. Yagis for 144 Mc. and higher seldom have less than 5 elements, and more is certainly desirable. Such antennas for 144 Mc. are so much smaller than for 50 that quite different structural methods are usually employed, though the principles of operation remain the same. With a little effort and shopping around, the materials for a pretty fair 2-meter antenna can be picked up for almost nothing. The wood-boom Yagis that follow are good examples.

Broomstick Beam for $1.50

Something close to the ultimate in decibels per dollar in 2-meter antennas is shown in Fig. 9-29. It was made by W1ICP after a search for the lowest-cost materials available. The beam works just as well as one of the same number of elements costing 20 times as much, though it might not last as long in rooftop service. It can be dismantled readily for portable use.

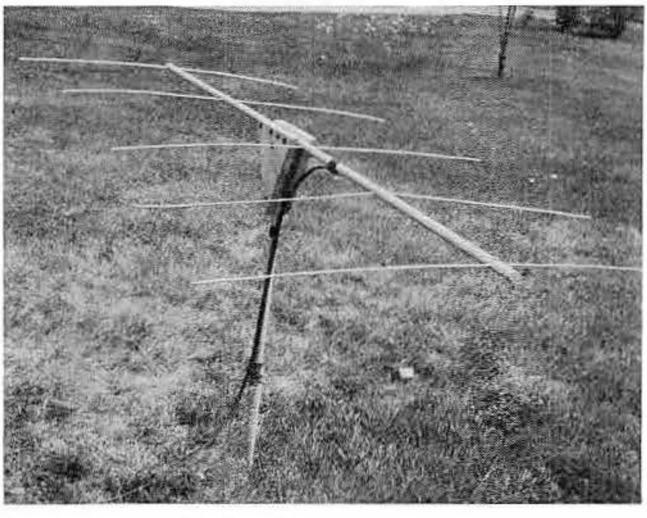


Fig. 9-29—Low-cost 5-element beam for 144 Mc. Note feedline taped to the boom and support.

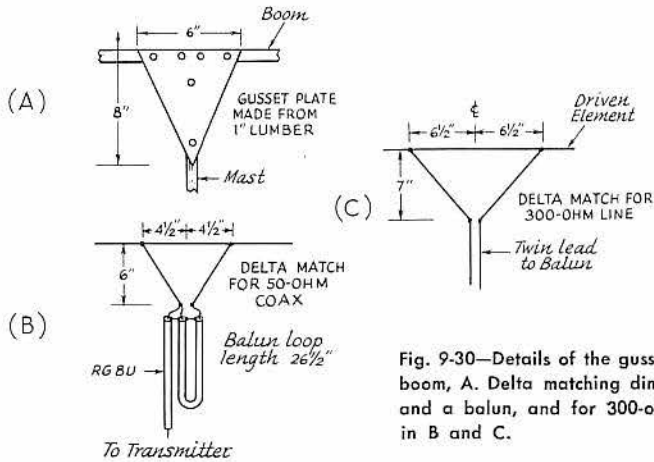


Fig. 9-30—Details of the gusset-plate mounting for the boom, A. Delta matching dimensions for 50-ohm coax and a balun, and for 300-ohm Twin-Lead, are given in B and C.

The boom is two 3-foot lengths of ¾-inch wood dowelling, available at any lumber yard and most hardware stores, for around 25 cents each. These are fastened to another length, for the vertical support, using 6 × 8-inch gusset plates, as shown in Fig. 9-30A. These can be cut from any wood scraps. Thin outdoor plywood or tempered Masonite will be light and strong.

Elements are cut from hard aluminum "picket wire," found in most hardware and garden-supply stores. It sells for around 75 cents for 25-foot rolls. Other element materials are available, though they may cost more than the wire. Aluminum rod of various diameters and tubing up to about ¼ inch are fine. Welding rod is cheap and plentiful, though it may be sold in lengths less than the 38 and 40 inches needed for the driven element and reflector.

First, drill the dowels for the elements, as shown in Fig. 9-31. Perfect alignment is important only from an appearance standpoint. Exact spacing is not important either. Preferably the holes should be of such size that the elements will be a press fit. Wood screws run down from the top of the boom bite into the elements slightly, to hold them in place. Drill holes for these screws somewhat smaller than the threaded portion, to prevent splitting the boom when the screws are inserted. The dowels and gusset plates are best assembled with 6/32 screws, with washers under the heads and nuts.

The feed system uses a delta match, and either 300-ohm Twin-Lead or coaxial line and a balun. Fig. 9-30C gives dimensions for Twin-Lead or 72-ohm coax and a balun. Connection of a balun is shown at B, with dimensions to be used when the coax is the 52-ohm variety. In the method at C, a run of Twin-Lead can be used and 72-ohm coax and a balun connected at the end near the transmitter, if desired. Some mismatch will result if 52-ohm coax and a balun are used in this way, though it is not enough to be harmful. Do not use small coax under any circumstances, except for runs of 25 feet or less.

The delta or Y can be made of Twin-Lead, merely by slitting it for the necessary length with a sharp knife, and then fanning it out Connections to the driven element are made with small clips cut from any convenient metal. Be sure that contact is clean and tight. The wire elements won't stand much tension, so the Twin-Lead or coaxial line and balun should be taped to the wood boom.

If a Twin-Lead line is used, the match will be close enough if the dimensions are followed. With coax you can adjust the points of connection on the driven element for a perfect match, using an s.w.r. bridge connected in the coaxial line. Be sure that tight contact is made to the driven element each time a reading is taken, and that the distance out from the center is the same on each side.

The wood support can be inserted in a TV rotator, or fitted into 1-inch water pipe or ¾-inch electricians' conduit, to extend the length. The beam makes a fine setup for portable work, as it can be taken apart in a few minutes and carried in a small bundle.

Intermediate-Length Yagis

The low-budget 5-element Yagi just described demonstrates the fact that gain up to

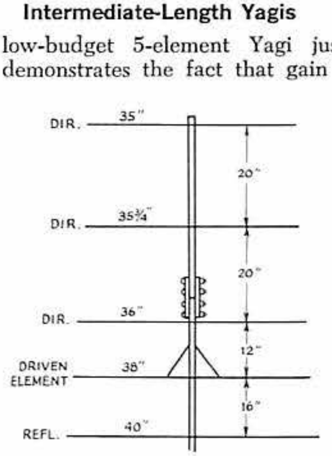


Fig. 9-31—Element lengths and spacings for the 5-element Yagi.

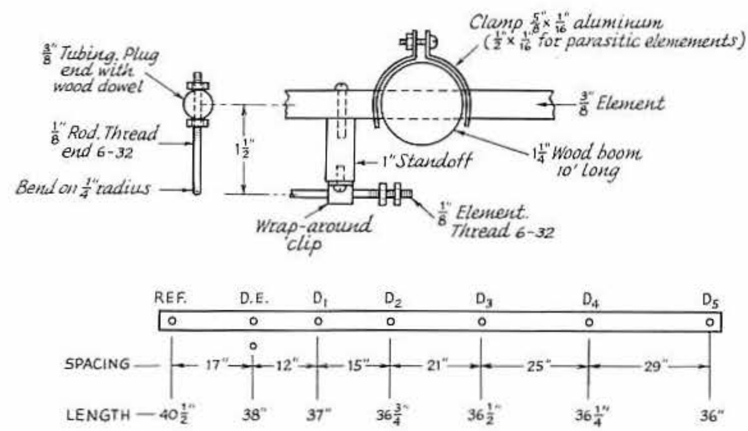


Fig. 9-32—Dimensions and structural details of a 10-foot wood-boom Yagi for the 144-Mc. band. The end view of the folded-dipole driven element is shown at the upper left, and the method of mounting it in the boom, at the right. Dimensions are for feeding with 52-ohm coax and a balun. Element lengths are for optimum performance between 145 and 146 Mc. Detail of the driven element is shown for one side only, in the interest of clarity.

9 or 10 db. comes quite easily in 2-meter Yagis. If we look back to Fig. 8-4, we see that after the first few elements the gain per element tapers off, and the boom length increases markedly. Despite this, Yagis of intermediate length may be useful in many circumstances. They can be built readily and at moderate cost, if wood is used for the boom and element materials are shopped for with low price in mind. A 10-foot Yagi of respectable performance, that can be erected with no more fuss than a good TV antenna, is detailed in Fig. 9-32.

A good material for medium-length booms is round wooden stock available in most lumberyards. Commonly called rug or closet poles, they come in various lengths and diameters. The boom shown is 10 feet long, and 1¼ inches in diameter. It should hold alignment without bracing, if properly treated to prevent moisture absorption. Select stock that is thoroughly dry, and free of knots. When holes have been drilled, spray or brush on clear lacquer, dry, and then brush on outside paint. When all mechanical and electrical work has been completed, the assembled antenna can be sprayed with clear lacquer to prevent corrosion of metal parts.

Parasitic elements can be hard-drawn aluminum wire or welding rod, ⅛ inch or larger, or tubing up to ⅜-inch diameter. The smaller stock is preferable. The 10-foot boom is probably about the longest that will hold up well without bracing, but if a longer beam is wanted it can be braced as shown in Fig. 9-16. Additional directors should be spaced 29 inches apart, and each made progressively ¼ inch shorter. Element spacings are not particularly critical. The reflector can be anywhere from 12 to 20 inches in back of the driven element, with only a slight effect on performance, provided that the matching system is adjusted to take care of varying feed impedance. Director spacing can be varied plus or minus an inch or so with no noticeable change in characteristics. Element lengths can vary plus or minus ⅛ inch without any change that could be observed except by the most careful check on frequency response.

With the ratio-type dipole shown in Fig. 9-32, the antenna can be fed directly with 50-ohm coax and a half-wave balun. A small range of adjustment can be had by bending the ⅛-inch portion of the folded dipole nearer to or farther from the ⅜-inch portion. The curve of Fig. 9-33 shows the standing wave ratio of the array with the dimensions given, using this method of feed. This is a good setup for the fellow interested mainly in operating above 145 Mc. If optimum performance is wanted near the low end of the band the elements can all be made ¼ to ½ inch longer.

In mounting the elements the boom should be drilled just large enough so that the elements fit tightly into the holes. They can be held in place by wood screws run into the boom and bearing firmly against them. These screws can be bonded together with a wire running down the boom, and this can be grounded, for lightning protection. A better method of holding the elements in place is shown at the upper right of Fig. 9-32. This clamp arrangement works equally well with any round boom, regardless of material used.

Driven-Element Construction

Construction of the ratio-type dipole is shown in Fig. 9-32. The unbroken portion is ⅜-inch tubing, the ends of which are plugged with wood dowels to permit tightening nuts against it to hold the fed portion of the dipole in position. The latter is ⅛-inch wire or rod, with the ends threaded for 6-32 nuts. An alternative

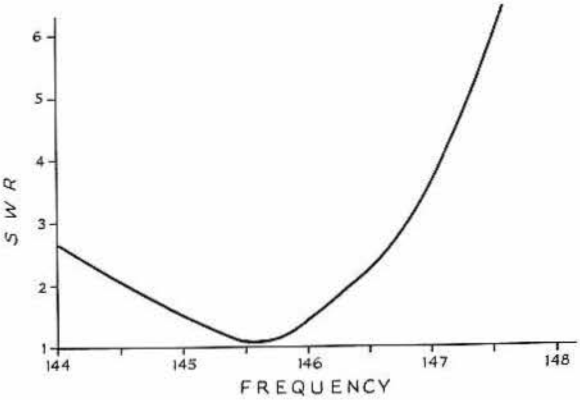


Fig. 9-33—Curve of standing-wave ratio taken with the 10-foot Yagi of Fig. 9-32. For optimum matching in the first megacycle, make all elements ¼ to ½ inch longer, leaving all other dimensions as shown. Note that mismatch rises more rapidly on the high-frequency side.

method of making end connections, if you do not have a threading die, is to hammer the rod end flat and then drill it to pass a 6-32 screw. The outer end can be bolted to the ⅜-inch portion, and the inner to soldering lugs attached to the ends of the coaxial line and balun.

The two portions of the dipole are held in alignment by means of 1-inch ceramic or teflon standoffs, one on each side of the boom and about one inch out from it. A 6-32 screw running through the ⅜-inch upper portion and a wrap-around clamp of thin metal, also with a 6-32 screw, hold this assembly together as shown in the sketch.

An alternative driven-element design that allows a wide range of adjustment, and use of any type of transmission line, is shown in Fig. 9-34. This is made from a single piece of stiff wire 160 inches long, bent as shown to include both the dipole and the universal matching stub. The sliding short on the stub, and the point of connection of the line or balun, are adjusted for zero reflected power, using a frequency in the middle of the range where optimum performance is desired. Readjusting for various frequencies will extend the useful range of this array beyond that in the s.w.r. curve of Fig. 9-33. The stub matching method can be used to extend the range of the ratio dipole, as well, if desired.

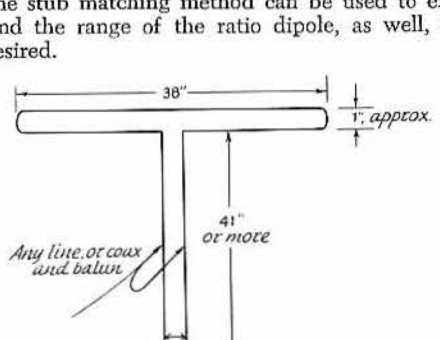


Fig. 9-34—One-piece folded dipole and universal matching stub, for substitution in place of the ratio-type dipole. Adjustment of the stub length and point of connection of the feedline can be made for optimum matching anywhere in the lower half of the band, with the element lengths given in Fig. 9-32.

Stacking

Where two of these Yagis are to be stacked the type of dipole used is unimportant, for matching will be taken care of at the central feed point, preferably with the universal stub. If a ratio-type dipole is used in each bay, with open-wire line for the phasing section, the s.w.r. on this line will be lower than if a uniform-conductor dipole is used, but this is not an important consideration, in view of the short run of phasing line.

The bays should be a full wavelength (81 inches) or more apart, using phasing and matching information already presented. Two of these 10-foot Yagis stacked should nearly equal the single 24-footer of Fig. 9-35 in gain. The stacked pair will have broader frequency response, and in some circumstances may be easier to install.

Long Yagis for 144 Mc.

Though there is no theoretical limit to the amount of gain that can be achieved in making ever longer arrays, the practical limit is reached in 144-Mc. Yagis at somewhere around 24 to 30 feet for most of us. If the limited frequency response of arrays this long is not a severe handicap, they are an attractive means of developing outstanding antenna performance.

Optimum element spacings and lengths for long Yagis were worked out experimentally by W2NLY and W6QKI some years ago, and similar work has been done by many others since. These projects have resulted in published figures that appear to be contradictory, but their seeming disparity merely shows that there are many ways to arrange the elements in a Yagi for roughly the same result in gain and bandwidth for a given length. This point explains the variations that will be found in comparing the Yagi systems shown in these pages.

One product of the W2NLY-W6QKI work was a 13-element 144-Mc. Yagi that has since become an almost standard long-Yagi design. Lengths and spacings for one version are shown in Fig. 9-35. This antenna is just under 24 feet long, but it can be made very light in weight and easy to handle. With the dimensions given

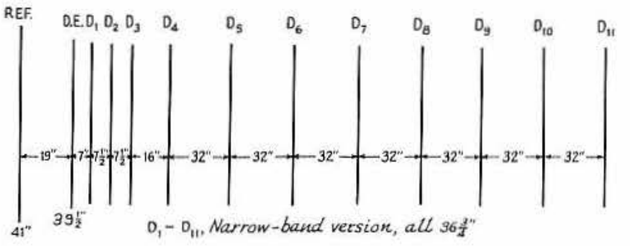


Fig. 9-35—Element lengths and spacings for a 24-foot high-performance 144-Mc. Yagi. Greater bandwidth can be achieved, at some sacrifice in forward gain, by tapering the element lengths as described in the text. Design information is from W2NLY and W6QKI.

in the drawing (all directors the same length) optimum performance is maintained essentially from 144 to 145 Mc. (The s.w.r. curve will show a rise at the low end, and a steeper rise approaching 145 Mc., but gain and minor-lobe content do not change markedly.

Polar plots of this antenna at various frequencies are given in Fig. 9-36. At the left are runs at 144 and 145 Mc. Note that the broken line for 145 Mc. shows a narrower main lobe, but some increase in size of the minor lobes. Plots at 145.5 and 146 Mc. show greatly reduced gain, and at the higher frequency almost no difference between the majo: and minor lobes. Above 145 Mc. appreciably, the array has no practical value.

This situation is improved somewhat by tapering the director lengths, as may be seen from Fig. 9-37, showing the relative gain level across the band with elements tapered ⅛ inch (solid line) and ¼ inch (broken line). The ⅛-inch taper extends the useful range to about 145.5 Mc., and the ¼-inch to above 146, but both involve appreciable sacrifices in peak gain. When tapered element lengths are employed, the first director, D_1 should be 37¾ inches, D_2 37⅝, D_3 37½, and then each additional director ⅛ or ¼ inch shorter, depending on the performance desired.

These curves and pattern plots represent the best that can be expected. In practice, the pattern of a long Yagi working above its upper useful frequency limit is little more than a mass of minor lobes and deep nulls. For work over all or most of the band, a collinear array, or a smaller Yagi, is to be preferred.

A convenient way to build a long Yagi is to obtain several telescoping aluminum mast sections, available from radio and TV distributors. These should be pinned together with self-tapping screws. Elements can run through the boom, if clamps such as shown in Fig. 9-32 are used to hold them in place. Hard drawn aluminum wire or rod ⅛ inch in diameter is a good element material. It is strong, and springy enough so that ice does not form on it readily. If an ice load does build up, the elements droop and the ice slides off, after which they spring back to their original position. A long Yagi of this construction, with suitable bracing for the boom, is practically indestructable.

With very long booms it is best to hang the antenna from its braces, though booms up to about 24 feet stand up well with bracing below the boom. The boom bracing of Fig. 9-16 is suitable for suspension. Where braces are below the boom, greater strength can be achieved by leaving the brace material round, and driving a tight-fitting wood plug into the end several inches, so that it will not compress when tightened in place. Lightweight aluminum angle stock also makes good boom braces.

Hauling a long Yagi up to its position is

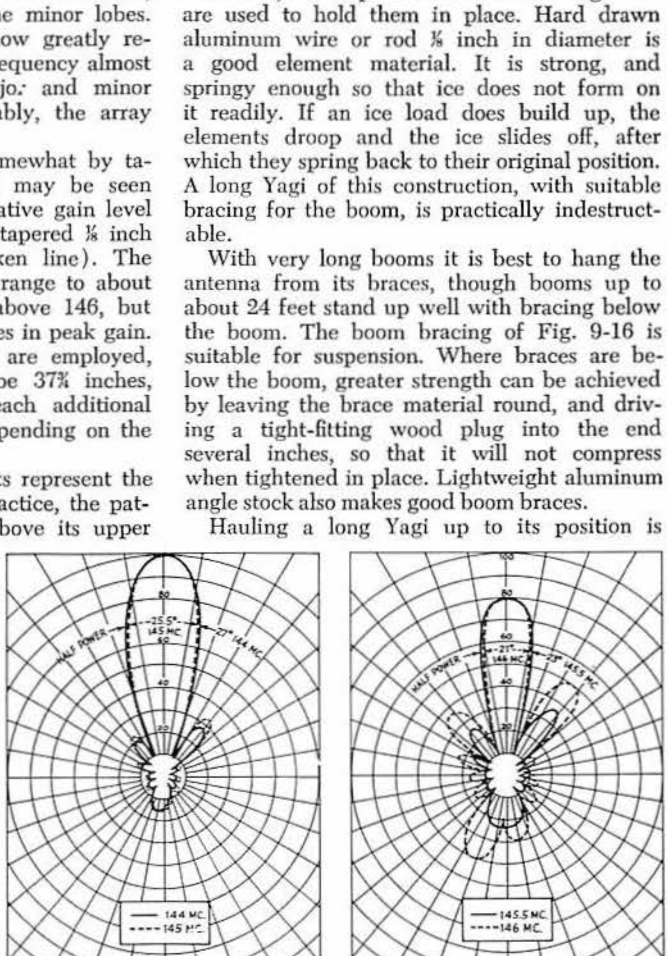


Fig. 9-36—Polar plots made with the long Yagi of Fig. 9-35. At the left are plots at 144 and 145 Mc. Note that at the higher frequency the main lobe is sharpening and minor lobes are larger. At the right we see that the gain is much lower and the minor lobes are much larger in the 145.5-Mc. curve, solid line. The array is of little value at 146 Mc., broken line, there being little difference between major and minor lobes.

often the most difficult part of an antenna installation. Many antenna workers solve this problem by designing the antenna so that it can be assembled atop the tower. The necessary components can be run up on a pulley rope with ease. This is practical if the boom is made in several sections that telescope together. The method has been used by the author of these lines on many occasions to put up large arrays single-handed.

Feeding the long Yagi presents no special problems, except that one should be sure that it is matched at the frequency to be most often used. Impedance of the array will depend on many factors, but if a basic impedance of about 25 ohms is assumed, and the matching system worked out accordingly, the result will never be very far off. The step-up folded dipole is usually the most convenient feed system. A half-inch solid portion, with the fed portion of ⅛-inch wire spaced 1¼ inches between centers, will give an 8-to-1 impedance step-up, and something close to 200 ohms feed impedance. Typical construction is shown in Fig. 9-32. This can be matched with 52-ohm coax and a balun, or a quarter-wave Q section can be installed at the feedpoint for higher impedances. If the spacing is made adjustable it will be possible to adjust the impedance of the Q section for perfect match to any desired feedline.

The Q section, if used, and the main feed line can be run along the boom toward the vertical support. This is usually preferable to letting the transmisson line dangle, as the feedpoint is a long way from the center of the system, and it puts considerable unnecessary strain on the support.

Stacking

Long Yagis have large "aperture;" consequently they require very wide spacing when stacked, if a real improvement is to be made. There is little to be gained from stacking two 24-foot Yagis closer than 12 feet apart, and 16 to 18 feet is better. In matching stacked systems, remember the rule that the feed impedance of a stacked pair is about half that of one bay, if the phasing line is any multiple of one wavelength long, electrically. If made any odd multiple of a half wavelength, the phasing system will act like a double Q section, when fed at its midpoint.

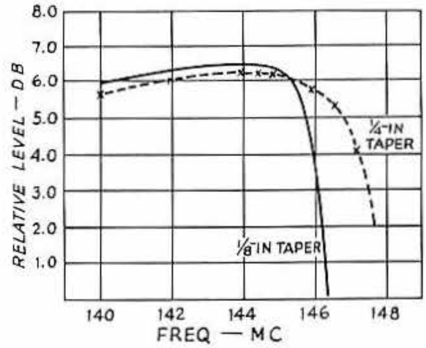


Fig. 9-37—Relative response of the long Yagi with director lengths tapered ⅛ inch, solid line, and ¼ inch, broken line.

Because of the mechanical and electrical problems in handling stacked combinations of long Yagis, more bays of shorter Yagis often may be a better solution. As an example, four 24-foot Yagis in a box configuration takes roughly 6000 cubic feet of space. It may give 20 db. of gain, if the job is done properly. A set of 8 6-element Yagis spaced one wavelength each way takes only about 1000 cubic feet of space, yet it should give almost as much gain, and probably over a wider frequency range.

4-BAY 20-ELEMENT 144-MC. ARRAY

The 2-meter array shown in Figs. 9-38 and 9-20, where it is stacked above a 6-element array for 50 Mc., has four Yagi bays of 5 elements each. The same general layout could be used with shorter bays, or longer ones up to about 7 elements. The spacing is one wavelength in the horizontal and vertical planes, which is optimum for bays of this approximate size.

The booms and frame are all ¾-inch aluminum tubing, hardware-store stock, available in 6- and 8-foot lengths. Four 6-foot pieces ($1.79 each) took care of the booms, and four 8-foot ones ($2.39 each) were used for the horizontal and vertical frame members.

The clamps are made as shown in model and drawing form in Figs. 9-22 and 9-39. Eight are needed. Elements can be anything from ⅛ to ¼ inch in diameter. Ours have ¼-inch center sections of tubing, with inserts of 5/32-inch aluminum wire or welding rod. Any stiff wire, tubing or rod stock could be used for the entire element. We used this combination for several reasons: it makes for exceptional strength, we had some of both but not enough of either for the entire array, and the inserts provided a convenient means of adjusting the element lengths. We will not bother with the various dimensions involved, other than the overall element lengths finally arrived at by experiment. See Fig. 9-38. The elements are run through the boom, and held in place by self-tapping screws, as seen in the mockup, Fig. 9-39. Aluminum screws for this purpose can be bought at the hardware store, and we recommend them over steel. They stay in place, and they won't rust.

The phasing system is shown at the right of

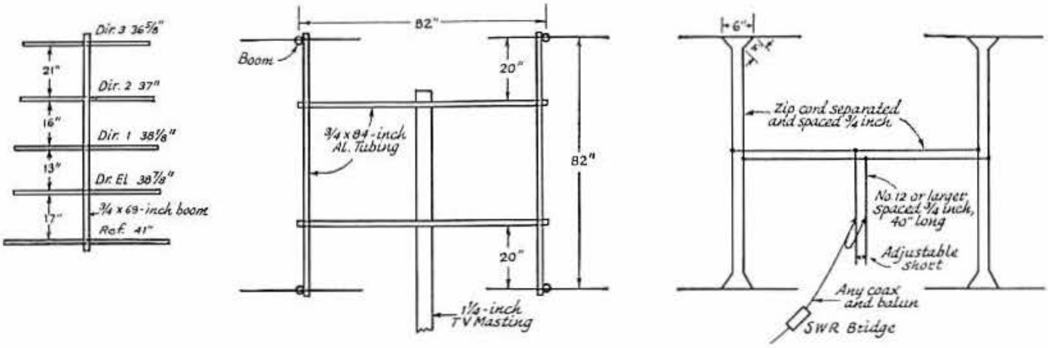


Fig. 9-38—Principal dimensions of the 144-Mc. array. Element lengths and spacings are given at the left. The supporting structure is sketched in the center. Details of the phasing harness and matching section are shown at the right. Impedances need not be known, since it is necessary only to adjust the position of the short and the point of connection of the balun for the frequency range most commonly worked. Dimensions of the fanned-out sections at the ends of the phasing harness are not critical, so long as all are the same size.

Fig. 9-38. A universal stub at the central feedpoint provides a simple means of matching without having to know the impedances involved. The phasing sections are fanned out near the point where they connect to each driven element. Here again, there is no precise dimension; just make the spacings and the traingular matching sections all the same. Be certain that a clean and permanently-tight connection is made to the driven element.

Phasing lines can be any balanced line, and most builders may prefer common TV open-wire line, either half-inch or one-inch spaced. We tried a different idea, and used ordinary zipcord from the electrical counter of the hardware store. We strung this on homemade spreaders cut from ⅜-inch wooden dowel, drilled to give about ¾-inch spacing. The insulation on the zipcode lasts well out of doors, and the wire is strong, yet flexible. Despite some early misgivings about this phasing setup, it has turned out to be durable and effective, as well as convenient and inexpensive.

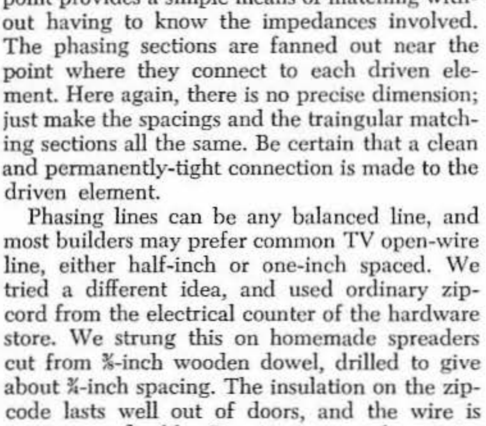


Fig. 9-39—Mockup showing methods for assembling the framework and elements of the 2-meter array.

Star lugs were soldered to the ends of the phasing lines to bolt to clips that wrap around the driven elements. The junction of the zipcord and the lug was wrapped with plastic tape and sprayed with Krylon. The wood-dowel spreaders were also sprayed. The line is supported at several points, using TV-type insulated standoffs which wrap around the appropriate dowel spreaders in the line.

The element lengths given are for maximum performance in the bottom megacycle of the band. If you want the beam to be most effective above 145 Mc. it would be desirable to shorten all elements by ¼ inch for each megacycle higher. Element lengths are not too critical, provided that the short and points of balun connection on the matching stub are adjusted for zero reflected power at the center frequency you select. A single 5-element Yagi was made as a preliminary to the 4-bay system, and adjusted carefully for optimum performance between 144.5 and 145 Mc. We found only a discernible difference in forward gain from 144.0 to nearly 147 Mc., after readjusting the matching stub for each frequency change. Both gain and front-to-back ratio dropped off markedly above 147.

It was interesting to note that the frequencies of optimum gain and front-to-back (they're not the same) moved down about 500kc. with the stacked system, compared with the single Yagi. Presumably this was the result of coupling between bays, and the introduction of more metal in the field of the array. The element lengths given are corrected for the 4-bay system. The single 5-element would have ⅛ inch more per element for peak performance over the same frequency range. Precise adjustment of element lengths is not too important, however, if the matching is adjusted in the frequency range most used.

THE SKELETON-SLOT ANTENNA

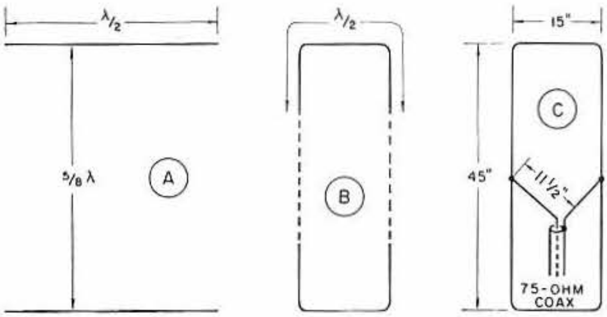


Fig. 9-40—Derivation of the skeleton-slot v.h.f. array. Only the driven element is shown, in the interest of clarity. Parasitic elements are lined up with top and bottom portions of the driven element, giving the effect of a stacked Yagi with ⅝-wavelength bay spacing.

A v.h.f. antenna that is very popular with British v.h.f. enthusiasts is shown in Fig. 9-40. Developed by B. Sykes, G2HCG, and sold by his company, J-Beams, Ltd., on both sides of the Atlantic, this so-called "skeleton-slot" array gets its name from the nature of its driven element, derivation of which is as follows:

Start with two half-wave dipoles spaced ⅝ wave-length, one above the other, as at A. Radiation is mainly from the center portions of these, so the ends are bent toward each other, as in B. Then they are joined with what is essentially a wide-spaced transmission line, C, and fed with a fanned-out Y section and coaxial or balanced line. Balanced-to-unbalanced conversion, for feeding with 75-ohm coax, is accomplished with a coaxial sleeve as detailed in Chapter 8. This is not shown in the sketch. The name of the array comes from the fact that this radiator behaves in much the same manner as a slot in a plane of metal, but in this case the plane is reduced to a closed loop.

Polarization is in the plane of the 15-inch portions, or horizontal in the example. These replace the usual driven elements in a stacked-Yagi system, and parasitic elements are lined up with them in the same way as in a conventional Yagi. Vertical spacing is ⅝ wavelength. Dimensions given are for 145 Mc., and broader frequency coverage is claimed than would be the case for a Yagi of similar dimensions. Up to 7 parasitic elements are commonly used in each half of the array. Element spacing is similar to that employed in Yagi design.

Two or more of these stacked slot-fed systems can be placed one above the other or side by side, and fed in phase in the manner of stacked Yagis. Starting with about 72 ohms for the first set, a stacked pair will have a feed impedance of about 36 ohms, and so on. Spacing of the sets varies between 1 and 3 wavelengths, depending on the length of each bay, following the rules for Yagis set forth in Chapter 8. Slot-fed arrays are common on both 144 and 432 Mc. throughout the United Kingdom, and elsewhere in Europe. The fore-going information is published with the kind permission of The Radio Society of Great Britain.

QUADS FOR 144 MC.

Though it has not been used to any great extent in v.h.f. work, the Quad antenna has interesting possibilities. It can be built of very inexpensive materials, yet its performance should be at least equal to other arrays of its size. Adjustment for resonance and impedance matching can be accomplished readily. Quads can be stacked horizontally and vertically, to provide high gain, without sharply limiting the frequency response.

The 2-Element Quad

The basic 2-element Quad array for 144 Mc. is shown in Fig. 9-41. The supporting frame is 1 by 1-inch wood, of any kind suitable for outdoor use. Elements are No. 8 aluminum wire. The driven element is one wavelength (83 Inches) long, and the reflector 5 percent longer, or 87 inches. Dimensions are not particularly critical, as the Quad is relatively broad in frequency response.

The driven element is open at the bottom, its ends fastened to a plastic block, which is mounted at the bottom of the forward vertical support. The top portion of the element runs through the support and is held firm by a screw running into the wood and the bearing on the aluminum wire. Feed is by means of 52-ohm coax, connected to the driven element loop. For a perfectly nonradiating line the coax should be fitted with a detuning sleeve (see Fig. 8-26) but omission of this precaution does not seriously affect the performance of the Quad.

The reflector is a closed loop, its top and bottom portions running through the rear vertical support. It is held in position with screws, top and bottom. The loop can be closed by fitting a length of tubing over the element ends, or by hammering them flat and bolting them together, as shown in the sketch.

The elements in this model are not adjustable, though this can easily be done by the use of stubs. It would then be desirable to make the loops slightly smaller, to compensate for the wire in the adjusting stubs. The driven element stub would be trimmed for length and the point of connection for the coax would be adjustable for best match. The reflector stub could be adjusted for maximum gain or front-to-back ratio, whichever quality the builder wishes to optimize.

In the model shown only the spacing is adjusted, and this is not particularly critical. If the wooden supports are made as shown, the spacing between the elements can be adjusted for best match, as indicated in an s.w.r. bridge connected in the coaxial line. The spacing has little effect on the gain, from 0.15 to 0.25 wavelength, so the variation in impedance with spacing can be utilized for matching. This also permits use of either 52- or 72-ohm coax for the transmission line.

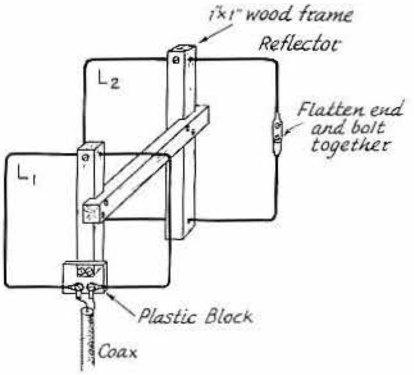


Fig. 9-41—Mechanical details of a 2-element Quad for 144 Mc. Driven element, L_1, is one wavelength long; reflector, L_2, 5 percent longer. Sets of elements of this type can be stacked horizontally and vertically for high gain with broad frequency response. Bay spacing recommended is ½ wavelength between adjacent element sides. Example shown may be fed directly with 52-ohm coax.

Stacking

Quads can be mounted side by side or one above the other, or both, in the same general way as described for other antennas. Sets of driven elements can also be mounted in front of a screen reflector. The recommended spacing between adjacent element sides is a half wavelength. Phasing and feed methods can be similar to those employed with other antennas described in this chapter.

Adding Directors

Parasitic elements ahead of the driven element work in a manner similar to those in a Yagi array. Closed loops can be used for directors, by making them 5 percent shorter than the driven element, or about 79 inches. Spacings can be similar to those for conventional Yagis. In an experimental model built by W8HHS the reflector was spaced 0.25 wavelength and the director 0.15. A square array using four 3-element bays worked out extremely well.

Workers using Quads on 144 Mc. have reported reduced fading, compared with horizontal Yagis. Possibly this is due to the presence of some vertical polarization with the Quad, making it less affected by polarization changes that tend to occur over long paths.

OMNIDIRECTIONAL 144-MC. ANTENNAS

The omnidirectional systems described for 50 Mc. are suitable for 144 as well, and in general they may be duplicated for the higher frequency by using dimensions from Table 9-I. Specific details of 144-Mc. versions will, therefore, be given only where information on tested designs is available.

Vertical Collinear

The smaller size, and greater need for gain at the higher frequency, tend to emphasize stacked systems at 144 Mc., whether directive or not. A vertical collinear array that can be used with any odd number of half-wave radiators is shown in Fig. 9-42. It can be fed with either 300-ohm line or coax and a balun at the midpoint of the center half-wave element.

As pictured, a three-half-wave system is made from two 97-inch pieces of aluminum wire, mounted with ceramic standoff insulators or TV hardware on a wooden pole. The coaxial line is taped to the vertical support, though a preferred method where balanced feed is used would be to run the line horizontally at least a quarter-wavelength from the feed point. A 5-element system would be made in the same way, using additional folded half-wavelengths of wire between each radiating element.

The Halo

Though not truly omnidirectional, the halo antenna is often used where a minimum of directivity is desirable, as in mobile work. In its 144-Mc. form, it is usually a halfwave element bent into circular shape. The diameter is not critical, so long as the element is resonant in the desired operating frequency range. The halo is widely used for 2-meter mobile, where small size is its principal virtue. Halos are used in stacked pairs occasionally, for increased gain and lowered radiation angle. Individual halos or stacked pairs can be fed by any of the methods shown for driven elements in Chapter 8. Halos of various types for 50 and 144 Mc. are described in the mobile section of this chapter.

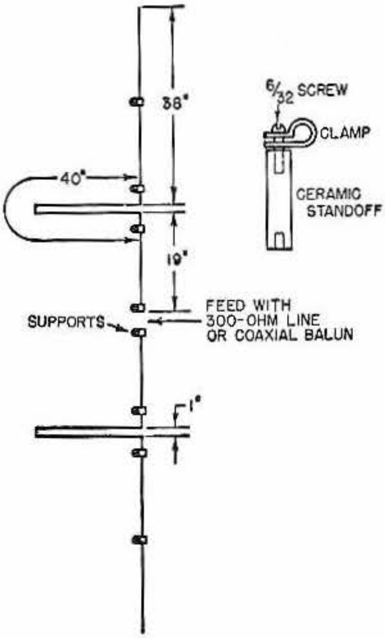


Fig. 9-42—Vertical collinear array for the 2-meter band may be made from two pieces of aluminum wire, bent as shown. Supports can be ceramic standoffs or TV hardware, fastened to wooden pole.

The Big Wheel

A weakness of the halo is its small size, and resultant low gain. It is considerably below a halfwave dipole, compared with the latter's pattern in its favored directions perpendicular to the dipole. The halo is also quite limited in frequency response, particularly when capacitively loaded. An omnidirectional antenna that is a considerable improvement over the halo in both respects is the "Big Wheel," developed by W1FVY and W1IJD, and shown in Fig. 9-43.

Fig. 9-43—The Big Wheel, an omnidirectional horizontal antenna for the 144-Mc. band designed by W1FVY and W1IJD. Radiating elements occupy an area approximately 40 inches in diameter.

Almost harder to picture and describe than to build, the Big Wheel consists of three one-wavelength elements connected in parallel and arranged in clover-leaf shape. The parallel connection results in a very low impedance, which is raised to 50 ohms with an inductive stub. Frequency response is very broad. With the stub adjusted to the proper length for perfect match at 146 Mc. the s.w.r. is negligible from 144 to 148 Mc. The radiation pattern is not perfectly circular, having slight dips in line with the notches in the antenna.

Elements (A in Fig. 9-45) are 80 inches long, of any convenient size tubing or rod stock. With tubing, the strength and stability of the antenna are improved if wooden plugs are driven into the element ends. One element

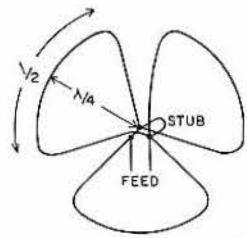


Fig. 9-44—Schematic representation of the Big Wheel. Three one-wavelength elements are connected in parallel. The resulting low feed impedance is raised to 52 ohms with an inductive stub.

end is fastened to a grounded angle plate, B in Fig. 9-45. The other connects to a floating triangular plate, C. The two plates are kept in alignment by a ceramic or bakelite insulator about 1½ inches high. The inner conductor of the coaxial line connects to the triangular plate, and the sheath to the angle bracket. The stub connects between these two plates.

Note from the schematic presentation, Fig. 9-44, that the elements are in parallel. Looking down at the antenna, if the left side of one element goes to the angle bracket, its right end goes to the triangular plate. Moving around to the right, the next element connects the same way, and so on, to the third. Only the stub length is critical, and since it is merely a strip of aluminum, several of various lengths can be made and tried. Another method is to slot the mounting hole in one end of the stub, so that its electrical length can be adjusted. Distorting the shape of the stub also will tune it to some extent. The only objective in this is to get an s.w.r. bridge in the line to show zero reflected power at 146 Mc. The s.w.r. will then be just detectable at opposite ends of the band. The size and shape of the elements contribute to the excellent broad-band characteristics.

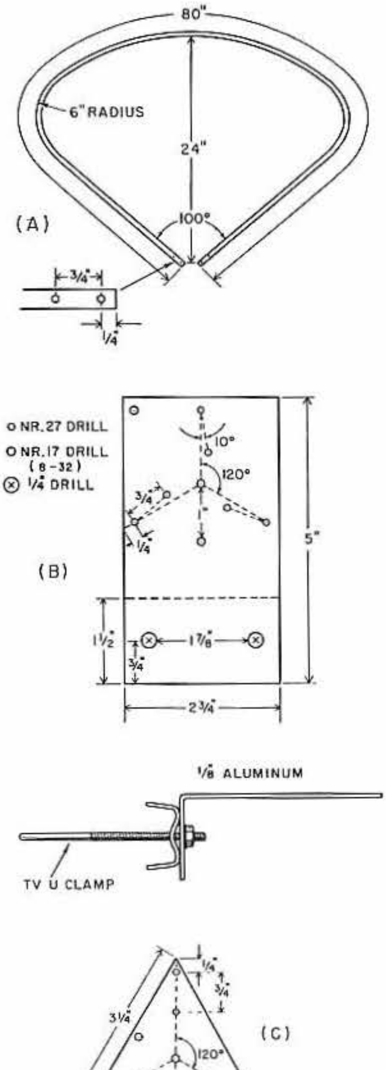


Fig. 9-45—Structural details of the Big Wheel. One element is shown at A. For strength the ends are plugged with wood. The grounded lower-support is shown at B. It is fastened to the pipe support with a TV U clamp. One end of each element is connected to this plate, and the other to the triangular plate, C. The tuning stub is shown at D.

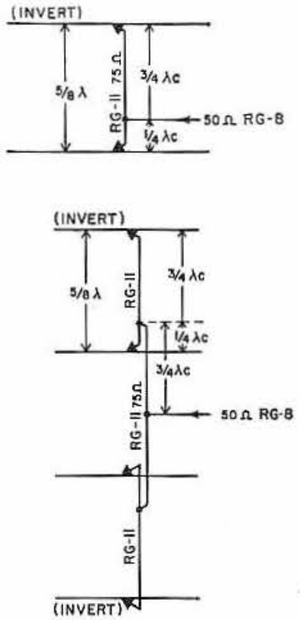


Fig. 9-46—Stacking arrangements for two and four Big Wheels. Off-center feed in the two-bay system requires that one bay be inverted with respect to the other. In the 4-bay stack the two center bays are the same side up and the two outer ones are inverted. Both systems are for 50-ohm feed. Dimensions given in λc should include the velocity factor of the coaxial line.

Fig. 9-47—Rear-deck-mounted turnstile for 144-Mc. mobile service.

A single Big Wheel is nearly as effective in all directions as a horizontal half-wave dipole in its favored directions, and it is some 2 to 3 db. better than a single halo. A very marked improvement comes with stacking a pair of Wheels, and this is easily handled electrically and mechanically. A stacked pair, and a pair of pairs, are shown in Fig. 9-46. Physical spacing is ⅝ wavelength. This is not critical; it may be set to whatever the phasing sections make convenient, in the vicinity of 50 inches.

The phasing lines are ¼- and ¾-wavelength sections of 75-ohm coax. These act as a double Q section, resulting in 50 ohms impedance at the T fitting junction. An electrical wavelength of solid-dielectric coax works out to be about ⅝ wavelength physically, the optimum stacking dimension. The length of the phasing sections should be checked out with a grid-dip meter, as described in Chapter 8.

The off-center connection for a pair of antennas requires that one be inverted with respect to the other. Polarities for sets of 2 and 4 are shown in Fig. 9-46. In the 4-stack, the feed impedance is kept at 50 ohms through use of two ¾-wavelength Q sections of 75-ohm coax, joined at the midpoint of the array with a T fitting, and fed there with 52-ohm coax for the main run.

The slight irregularities in the horizontal pattern of a single bay can be smoothed out in a stacked pair by positioning the bays so that the centers of the radiator elements of one line up with the notches of the other in the vertical plane. Coupling between bays of a stacked system requires some modification of stub length for perfect matching. The usual stub length for a single bay is 5 inches. For two bays stacked, the stubs will be 6 inches each. In a 4-bay system, the top and bottom stubs are 6 inches and the inner pair 7 inches. A longer stub than 5 inches may be needed for a single bay mounted near a car top or other large metal body, if perfect matching is to be achieved.

Gain from a stacked pair averages well above the 3 db. that theory would indicate, especially where the antennas are not at great height above ground. This probably results from the lowering of radiation angle that comes with stacking. Greatest improvement is observed with a pair, compared with one. With four, the overall gain is more in line with what one would expect, and the improvement over a stacked pair tends to be less than the theoretical 3 db. The end result, however, is omnidirectional gain roughly comparable to the gain of a 4-element Yagi in its favored direction.

Stacked omnidirectional systems are fine for control-station use in v.h.f. nets, or for situations where erection of a rotatable antenna is not possible, but they are not ideal substitutes for rotatable arrays. Interference problems may become acute when gain is built up in an omnidirectional array. Likewise, the broad frequency response of the Big Wheel is not an unalloyed blessing. The antenna may increase the trouble one has with spurious receiver responses, in a location where other v.h.f. services are operating close by.

Turnstile for Two

An unobtrusive but effective omnidirectional antenna for 2-meter mobile or fixed-station use is the "turnstile," Figs. 9-47–49. This adaptation by W1CUT is two half-wave dipoles crossed at

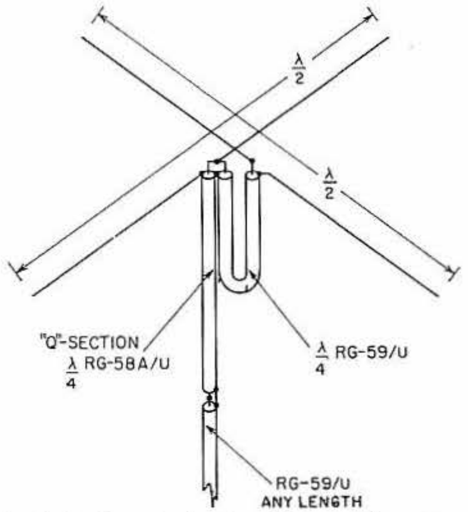


Fig. 9-48—Schematic drawing of the turnstile antenna. Crossed dipoles (dimensions from Table 9-I) are fed 90 degrees out of phase through a quarter-wavelength loop. The Q section is not ordinarily used in mobile installations where the line is short.

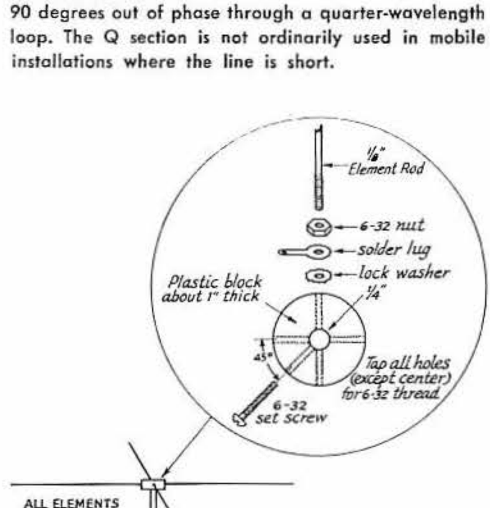


Fig. 9-49—Mechanical details of the lightweight turnstile. Design is by W1CUT.

right angles, fed with equal power but 90 degrees out of phase, as shown in Fig. 9-48. The quarter-wavelength stub provides this phasing of the second dipole. Note that this is not the usual coaxial balun; it is a *quarter* wavelength long, not a half wave-length. The pattern of the turnstile is essentially circular in the horizontal plane.

Mechanical details of the turnstile are shown in Fig. 9-49. The insulating support is a 1-inch piece of 1½-inch diameter polystyrene rod, drilled in the center to fit over a ¼-inch rod used for a support. A setscrew keeps this mount tight on the rod. Tapped holes 90 degrees apart take the four dipole elements, each 19 inches long, threaded to fit in the holes in the round block. Lock washers, soldering lugs and nuts hold the rods tightly in place and provide for connection of the line and stub, in the manner shown schematically in Fig. 9-48. Be sure that the dipole rods do not come in contact with the center support.

The Q-section arrangement shown is desirable if exact matching is important. In a mobile installation, where only a short feed line is needed, direct feed with any convenient small coax will work about equally well.

Though primarily for mobile use, this design may be adapted for neat home-station installation. The block can be fitted to the top of a pipe mast, and the coaxial line run down inside, if desired. Turnstiles may be stacked for additional gain, in a manner similar to that shown for the Big Wheel. Because of the light weight, up to four turnstiles may be stacked vertically with only a relatively small support. In mobile work the turnstile is somewhat superior to the halo, and is broader in frequency response as well. Length of the turnstile elements is not too critical on that account.

MOBILE ANTENNAS

The simpler antennas already described in these pages can be adapted to mobile service, usually with less in the way of permanent disfiguring of the car than may be required for mobile installations for lower frequencies. Often the v.h.f. mobile setup is casual and temporary —typically a Sunday afternoon drive with a rig that also serves the home station. For such ventures temporary antennas can be devised to fit almost any car, and they can be removed quickly, without leaving a trace. For more permanent installations (if any mobile station can be called "permanent") it is possible to devise ways of mounting equipment and antennas that are relatively unobtrusive and do not degrade the value of the vehicle when trade-in time comes.

No-Holes Mounts

Two temporary whip mounts are shown in Fig. 9-50. At the top is a sheet aluminum bracket with a coaxial fitting attached. This can be bent to fit the car door, or frame. The temporary coaxial lead can run over the top of the window, or in most cars it can run through the door opening, as it will not be subjected to long-term abrasion from opening and closing the door. The weather stripping around most car doors will pass RG-58 or 59 coaxial line easily. Similar arrangements can be made for temporary mountings around the rear deck openings of most cars.

The turret mount of Fig. 9-50 can be left in place as long as desired, and the antenna and line removed when not in use. Its construction from an ordinary can top, soldered to a sheet of copper or brass, should be obvious from the drawing. Plastic tape holds the mount to the car top, permitting removal without damage to the finish.

In many years of v.h.f. mobile operation in a variety of cars, the author has always found it possible to work out ways of mounting effective v.h.f. antennas without drilling visible holes. Removable trim and the holes for mounting it provided means for fastening the turnstile of Fig. 9-47 in place, and bringing in its coaxial line feed. Air vents offered ready-made holes and access to the car interior in the case of the all-purpose mount of Fig. 9-51. This takes antennas mounted in a PL-259 plug, or on a ¼-inch rod or tubing support. A compartment around the gas tank fill pipe was used for the whip bracket of Fig. 9-52. Even the screws and

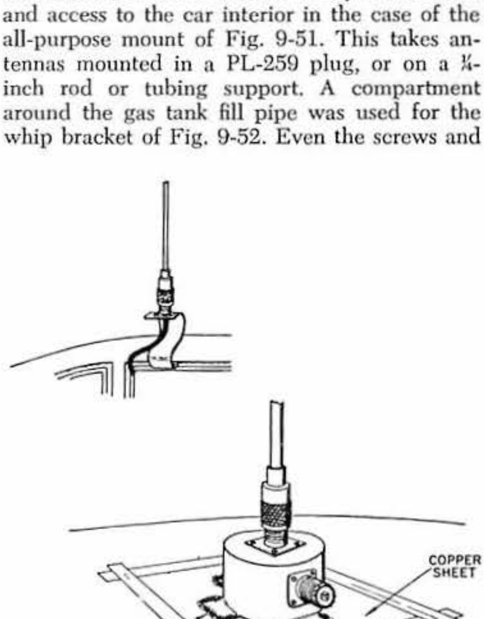


Fig. 9-50—Two no-holes mounts for 144-Mc. mobile whips. The clip, a temporary expedient, is merely a coaxial fitting mounted on a light strip of aluminum, which can be bent over the top of the window or around the edge of the door opening. The turret is cut from a can top, soldered to a thin copper plate. This is taped to the car top. Tape, or a sheet of thin plastic, under the plate will permit easy removal of the assembly, leaving no damage to the surface.

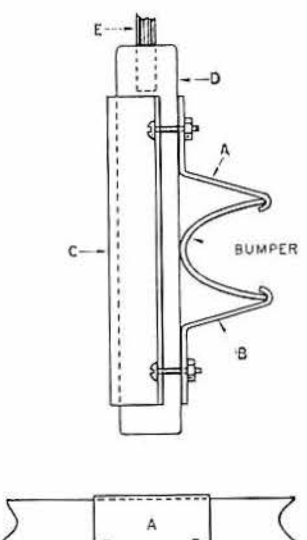


Fig. 9-51—Top and side views of a bumper mount easily made from sheet aluminum. Clamps A, B and C, $\frac{1}{8}$-inch stock, hold vertical member, D, tightly to the bumper. Vertical support can be tubing, for heavy antennas like a 50-Mc. halo, or wood, as shown. The rod E is $\frac{1}{2}$-inch aluminum, drilled at the top end to take the turnstile support of Fig. 9-47 and 49.

holes already in the car were used for mounting this one, and only a small hole in the side of the compartment, inside the rear deck, was needed for the coaxial line.

A detachable bumper mount for mobile antennas such as the turnstile or halo is detailed in Fig. 9-51. This is handy for supporting portable beams, as well. See Fig. 9-58. It requires only a few pieces of sheet aluminum, a section of round wooden closet pole or broom handle, and an aluminum rod or tubing for the vertical member. Dimensions will vary with each installation, so only the basic ideas are given here. For a 6-meter halo the support should be aluminum masting, and the brackets should be made of $\frac{1}{8}$-inch stock. For light 2-meter antennas, brackets of $\frac{1}{16}$ or $\frac{3}{32}$-inch aluminum should suffice.

Mobile Whips

Where vertical polarization is in general use the whip is quite a satisfactory mobile v.h.f. antenna. It is also used by casual mobiles in horizontal areas, regardless of polarization disadvantage, because of its convenience and unobtrusive appearance. Cross polarization does not pose too severe a problem in mobile operation; the polarization of the received signal is likely to be mixed, as a result of multiple reflections. The vertical whip may work well in reception of ionospheric DX on 50 Mc., again because of the random nature of the received-signal polarization. The chief weakness of the whip is that, like other verticals, it is much more susceptible to ignition noise than horizontal systems.

Whips for 144 Mc. or higher bands should preferably be mounted near the center of the car roof, as this gives a large ground plane, nearly omnidirectional coverage, and a low radiation angle. Roof mounting is good for 6-meter whips, too, but not everyone likes a 5-foot vertical on a car top. Family circumstances may dictate use of the broadcast whip, or reasonable facsimile. The whip of Fig. 9-52 serves three purposes. It is connected to a coaxial switch on the dash, and leads from this run to the broadcast receiver and to mobile rigs for 6 and 2. Other positions of the switch can be used for testing other gear, without disrupting the "permanent" installations.

Length of a 50-Mc. whip is not particularly critical. A field-strength indicator may not show too much variation with changes of an inch or two either way, but the best length should be found experimentally, as it may not be the theoretical quarter-wavelength. The position of the antenna on the car may affect both the length and the feed impedance. Do not adjust length for best match, as the nominal impedance of a whip is below 50 ohms. Trim the coax line length, if necessary, for optimum loading.

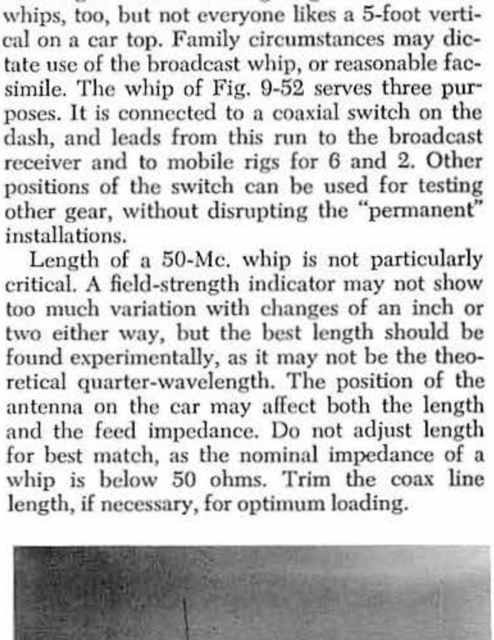


Fig. 9-52—Three-purpose whip serves for broadcast reception and for 50- and 144-Mc. mobile work. Set to optimum 50-Mc. Length, it works as a $\frac{3}{4}$-wave antenna on 144, though performance is improved slightly if length is adjusted when changing bands. Mount is an aluminum bracket fastened inside the well around the gas tank opening.

A 50-Mc. whip of optimum length, usually 54 to 56 inches, works reasonably well in the ⅝-wave mode on 144 Mc. If you extend the whip about 4 inches for 2-meter work, resonance can be obtained for both bands. The radiation angle is high for a ⅝-wave whip, but it is a convenient expedient for casual two-band work. If you work only the 2-meter band, a 19-inch car-top whip will almost certainly be better.

Novel 2-Meter Halo

A quick-disconnect 2-meter halo that can be dropped onto the broadcast antenna or added to a 6-meter whip is shown in Fig. 9-53. The brainchild of W3KDZ, it uses the ancient principle of single-wire feed. In this case the whip acts as the transmission line, and is connected to the halo off-center, at the approximate matching point. The halo should preferably be about 40 inches up on the whip, and if possible the whip should not extend above it. Minimum vertical radiation is obtained in this way. The distance off center on the halo should be adjustable, but 3 to 5 inches is a good starting point.

The length of the halo element will depend to some extent on the diameter of the circle. The smaller the circle the shorter the element, because of increasing capacitance end-to-end. The one shown is 40 inches long, and resonates at 145 Mc., when bent so that the opening is 9 inches. In the original by W3KDZ the element was 34 inches, and the ends were fitted into a polystyrene insulator about 2 inches long. A grid-dip meter resonance check is desirable, in any case.

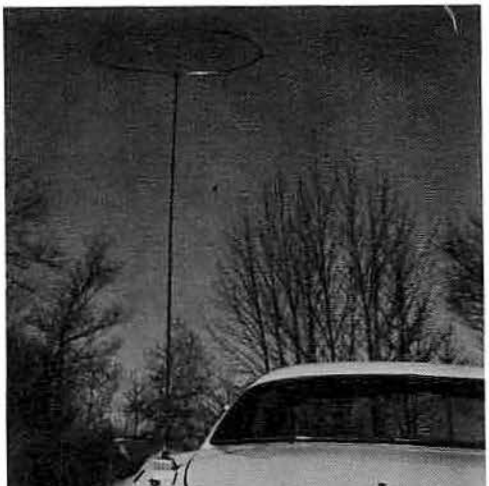


Fig. 9-53—This 144-Mc. halo appears to have no transmission line or matching device. Attached off-center to the whip of Fig. 9-52, it uses the whip as a single-wire transmission line. Optimum whip length is approximately 40 inches, so top section is telescoped when the halo is used. Very little vertical polarization is in evidence.

Walking in circles with a field-strength meter showed some interesting pattern and polarization variations with the antenna of Fig. 9-53. With the whip alone the 144-Mc. polarization was predominantly vertical, with a major lobe off the back of the car, at an angle of about 15 degrees to the right of the line of travel. With the halo at 40 inches up the whip, left in its extended position, the polarization was mixed, with vertical strongest on the main lobe, but with several horizontal and 45-degree lobes elsewhere. With the whip run down to the halo (whip now 40 inches long, thus mismatched in the vertical mode) horizontal was mainly evident, with some energy at 45 degrees, but almost no vertical. In other words, the single-wire feed principle *does* work. The halo is by no means omnidirectional, however; its main lobe is forward and to the left, perpendicular to that portion of the halo near its high-current midpoint. A lesser lobe appears off the back and to the right of the car heading.

2-Band Halo

A halo that can be set up to work on either 50 or 144 Mc. is shown in Fig. 9-54. This antenna is customary 50-Mc. size, 67 inches in circumference, with 2¼-inch square capacitor plates fitted to each end. The gamma matching arm is 14½ inches long, of the same material as the halo, and separated from it by ceramic insulators. A clip of sheet aluminum provides sliding contact between the arm and the halo. As always with the gamma match, be sure that this makes a clean tight contact.

The halo shown was put together mainly to try out the two-band idea, so its mechanical details are not spelled out here. Important points in a permanent installation are the arm-to-halo contact already mentioned, waterproofing of the series capacitor, and some adequate provision for keeping the halo rigid during driving. Any flopping of the halo causes intermittent detuning, and severe mobile flutter, in addition to that normally encountered.

Operation on two bands is achieved by changing the point of connection on the halo arm, the setting of the variable series capacitor, and the spacing of the square-plate capacitor at the element ends. The plate spacing is changed by using a ¼-inch ceramic standoff to hold the plates in position for 50 Mc. and changing to one ⅝ inch long to bring resonance down to 48 Mc., so that it will operate on its third harmonic.

Resonance must be achieved before the antenna can be matched on either frequency. This can be facilitated by use of a grid-dip meter, coupling to the halo near the vertical support. Resonance does not need to be exact for 144-Mc. work, as the antenna tunes more broadly on this band, but it is critical in the 50-Mc. band. When it is resonated near the desired frequency range, apply power and move the slid-

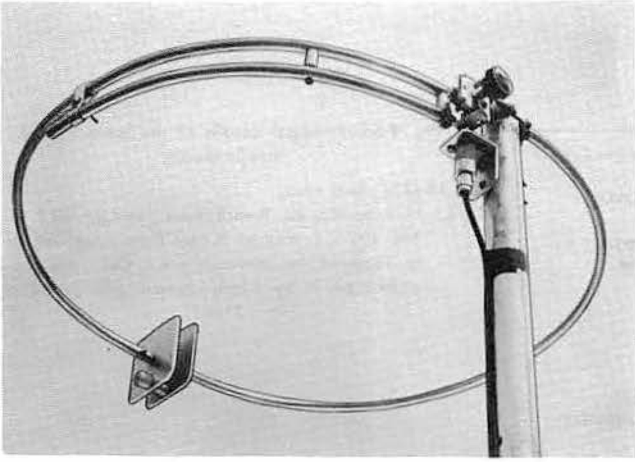


Fig. 9-54—Halo for 50 or 144 Mc. As shown it is set up for the 50-Mc. band. Change to 144 is made by moving connecting clip closer to the center post, and changing the spacing of the capacitor plates by swapping the ceramic insulators.

ing clip and adjust the variable series capacitor for lowest reflected power. These adjustments must be made with care for either band.

For 50-Mc. work the clip connection is near the outer end of the arm. For 144 it works at about 3 to 4 inches out from the capacitor. With a 25-pf. capacitor in series with the arm, tuning is near the middle of the range for 50 Mc. and close to minimum for 144. Properly adjusted the halo works well over only about 50 to 50.5 Mc. without readjustment. In the 2-meter band, satisfactory operation is possible over about half the band without retuning. The range with the antenna set up for 50-Mc. service seems to be normal for a halo on 6. On 144, the antenna seems to give somewhat better coverage than the conventional 2-meter halo, probably because of its larger size.

Two-Band Turnstile/Dipole

The turnstile of Fig. 9-47 can be modified to work on both 50 and 144 Mc, though it is no longer a turnstile when this is done in a manner which permits it to work on both bands without adjustment. It can be converted to a 50-Mc. turnstile with loading coils in each element, and the substitution of a 50-Mc. phasing loop. Or, loading coils can be put in one pair of elements, and the phasing loop left as it was. Now it will work as dipoles for 50 and 144 Mc. though admittedly not as well as either pair of elements would do alone.

The pair of loading coils can be left in place, and the system reconverted to 2-meter turnstile service, by shorting out the coils, or the coils can be removed and the elements replaced in their original positions. It can be seen that there are many options here, and any of the modifications are to some extent makeshift, but DX has been worked on 50 Mc. with the two-band setup, and the operation on 144 falls off only slightly from that of the original turnstile.

The coils are made of prepared stock, 11¾ turns each, ⅝-inch diameter, 16 turns per inch. The originals were of Miniductor No. 3007. They slip over half-inch ceramic pillars one inch long. The last turn of one coil can be broken loose, and bent to adjust the element to resonance at the desired part of the 50-Mc. band. Disconnect the coax, and jump a wire across the center insulator, so that the two elements are connected together. Resonance can then be found with a grid-dip meter. Remove the jumper, and reassemble.

A Neat 50-Mc. Dipole

Where a horizontal antenna is needed for 50-Mc. mobile, and a halo is too much of an eyesore, the dipole of Fig. 9-55 works reasonably well and is unobtrusive in appearance. If it can be mounted near the middle of the car the element ends will not extend far enough to be dangerous to passersby. Because radiation is largely from the center of a dipole, this design works better than one in which the loading coils are at the center of the element, as described just previously. The main lobe of radiation is

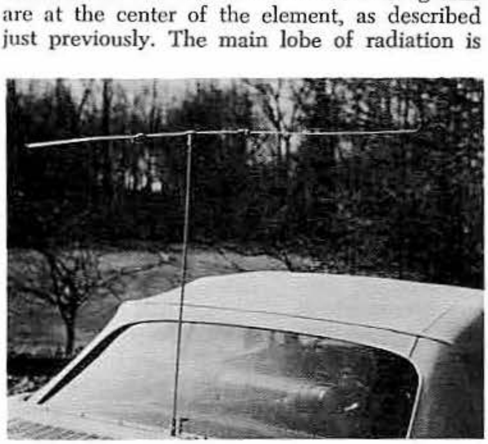


Fig. 9-55—Shortened dipole for 50-Mc. mobile service. Loading coils are inserted either side of a solid center section. Inserts are 19 inches long, threaded into forms that support the loading coils. See Fig. 9-56.

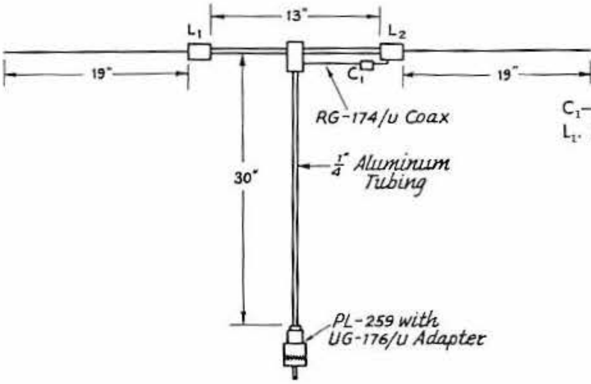


Fig. 9-56—Principal details of the loaded 50-Mc. mobile dipole.

C_1—15-pf. dipped mica.

L_1, L_2—11 turns No. 20, ⅝-inch diam., 16 t.p.i. (B & W No. 3007). L_2 tapped ¾ turn from inner end, or as required for minimum s.w.r. Coils are supported on ½ by 1-inch ceramic pillars (Millen 31001).

normally to the rear, with a lesser one forward. In stationary operation the element can be rotated for best signal.

As may be seen from Fig. 9-55, this horizontal dipole has loading coils at equal distances either side of center. The dimensions given are one of many possible arrangements. They were dictated by a desire to use 2-meter turnstile elements for the outer portions, and keep overall length to about 50 inches.

The 13-inch rod which is the center portion of the dipole is drilled and tapped at each end for 6-32 thread. The loading coils, L_1 and L_2, are made in a manner similar to those used with the two-band antenna. Prepared coil stock is slipped over ½ by 1-inch ceramic standoff insulators, and the wire ends are soldered to lugs at each end of the insulators. The element ends are ⅛-inch aluminum welding rod, threaded 6-32 for about one-half inch at their inner ends. A 6-32 nut is threaded onto the element, and this acts as a stop when the element is screwed into the insulator.

The 13-inch center section is supported in a ½-inch piece of solid aluminum rod about one inch long, with a setscrew running in from the top to hold the rod tightly in place. The lower portion of the block is drilled to take the vertical support, which is ¼-inch aluminum tubing. This can be any length that will stand the strain; ours is 30 inches long.

The diameter of the bottom end of the vertical member is filed down just enough so that it can be forced into the UG-176/U adapter, which, in turn, screws into the PL-259 coaxial plug. Small-diameter coax was first used for the feedline, bringing it out through a hole in the vertical support. This turned out to be fragile, so a piece of zip cord (one conductor and its covering discarded) was substituted and found to work just as well.

Adjustment

The top end of the line extending through the hole as shown forms the arm of a gamma match. The series capacitor, C_1, was first set up as a variable, permitting the right combination of capacitance and tap position on L_2 to be selected experimentally—but we're getting ahead of our story.

First the antenna by itself must be resonated at the center of the frequency range you want to work over. This will be a narrow frequency range, a limitation not too important, with most 6-meter operation being in the first 500 kilocycles of the band ordinarily. The resonant frequency can be checked with a grid-dip meter, putting the g.d.o. coil adjacent to the 13-inch center section of the antenna, close to the center block. The trick now is to trim the lengths of the outer elements, or the number of turns in the loading coils, until you hit the desired frequency. It will be a sharp indication; when you approach the desired frequency, do not trim elements by more than one-half inch at a time, or the loading coils by more than ¼ turn. Whichever you cut, be sure that the same change is made on both halves of the antenna. When you're through, the coils should be identical, and the outer ends of the element the same length.

The antenna used by the writer was trimmed for resonance at about 50.25 Mc. The next step was to find a value of series capacitor, C_1, and a point of connection on the antenna or loading coil, L_2, that would provide a 50-ohm termination for the coaxial line. This was done experimentally with the antenna support clamped in a vise on the workbench. A recheck of the s.w.r. and operating frequency, when the antenna was installed on the car, showed little change.

Frequency response is about the same as with a capacitively-loaded halo. Resonated and matched at 50.25, the dipole is usable from the low end to 50.5 Mc. before the s.w.r. rises above 2 to 1, a mismatch that is tolerable in a mobile setup.

2-Meter Dipole with Built-In Balun

A horizontal dipole makes a good 2-meter mobile antenna. Nearly all mobile antennas have some directional characteristics, depending on their position on the car, so the natural bi-directional pattern of a simple dipole may not be too much different from what usually results from the use of a supposedly omnidirectional mobile antenna.

The dipole of Fig. 9-57 is clean in appearance, and it has an invisible built-in balun. Installed on a car top it will have very nearly a true bi-directional dipole pattern, without the distorted lobe that usually results from direct feed of a split dipole with coax. It is very easy to make, and is handy for portable as well as mobile use.

The construction is similar to that for a turn-stile. A center insulator is mounted at the top of a support made of ¼-inch aluminum tubing, of any convenient length. The bottom of the support is forced into a UG-176/U adaptor, which screws into a PL-259 coaxial plug. The feedline is RG-58/U or RG-174/U coax, running inside the tubing. The coax should not make contact with the tubing at the top. The outer conductor is connected to one half of the dipole and the inner conductor to the other. The support is made to act like a 1-to-1 balun by running a self-tapping screw in at a point about 18 inches down from the top, to contact the outer conductor at that point.

The screw is not needed if the support is approximately a quarter wavelength high, as the ground will be made automatically by the connection of the coax to the PL-259 plug. In fact, several different lengths have been tried, and the sleeve support seems to work like a balun regardless of length. In the dipole pictured the grounding screw is about 6 inches above the fitting (support just over 2 feet high) and the radiation from the dipole is a good figure 8, whether the screw is contacting the outer conductor or not. It might be useful to be able to ground the outer conductor at this proper point in longer supports, however, and it is a simple thing to do. Be careful that the screw is not run in far enough to puncture the inner insulation, and short out the coax.


Fig. 9-57—Horizontal dipole for 2-meter mobile. The vertical support is ¼-inch tubing, with small coax inside, thus serving as a 1:1 balun.

PORTABLE ANTENNAS

From earliest times one of the great joys of v.h.f. hamming has been "working portable." Every enthusiast dreams of someday having a station on a mountaintop, with an unobstructed view for miles in every direction. Few of us ever see this dream become a permanent reality, but with the mobility we enjoy today nearly everyone can bring it off for a few hours now and then.

The catch is that even the finest v.h.f. locations have a way of turning out to be somewhat disappointing unless a good antenna is available. Halos, whips and the like are pretty poor stuff, compared with a beam antenna of even moderate size. Fortunately, fixing up a v.h.f. array so that it can be moved about readily is no great task.

Antennas can be built to encompass many degrees of portability. Probably the simplest are antennas and supports that come apart enough to permit tying to a ski rack or other car-top carrier. Most manufactured beams are shipped knocked-down, so dismantling for some portability is no problem. Yagis made for the TV trade nearly always have folding elements, to make life easier for the serviceman-installer. These can be modified for amateur bands quite readily. Occasionally the element spacings can be left as they were in the original, and the element lengths adjusted according to Table 9-I.

The next step is to make your own boom and supporting structure, using TV masting. This can be cut to the maximum length that your car storage space permits. Usually something around 4 feet maximum length is convenient. Element mounting methods described elsewhere in this chapter can be adapted to portable beams handily. The principal problem then becomes how to feed the array, since the two most popular matching systems, the folded dipole and the gamma match, do not lend themselves readily to quick dismantling and reassembly.

One thoroughly practical feed method is the delta or Y match, Fig. 8-17A and B, and 9-60. If the connection to the driven element is made with removable clips the delta provides a connecting and matching system that can be coiled up and carried in your pocket. The bottom of the delta can be terminated in a coaxial balun and coax of the desired length, or Twin-Lead can be used for all or any desired part of the main run to the equipment.

If the portable beam is to be used alongside the car the support can be tied or clamped to the door handle or bumper. An example of an effective bumper clamp is shown in Fig. 9-51, and in

Fig. 9-58—Portable beam set up for 144 Mc., using the bumper bracket of Fig. 9-51.

actual use in Fig. 9-58. A screw-driver or small stake can be driven into the ground to anchor the bottom of the mast.

Lightweights for 6 and 2

If you're satisfied to operate only in high spots that are accessible by car, extreme light weight is of no great interest, so long as the antenna can be taken apart and packed away conveniently in the car. But you really need the performance of a beam when you work a flea-powered transistor portable from some remote mountain top, miles from the nearest road. If you go for this kind of v.h.f. hamming, light weight and easy carrying are of primary importance. The arrays of Fig. 9-59 were designed to be carried on foot. They sacrifice nothing in performance for their lightness, so they also serve nicely for general-purpose portable operating.

Provision is made for a 3-element 50-Mc. beam and a 5-element 2-meter one. The booms are duplicates, so that one can be used for either band, or the two antennas can be set up together. All the material for the 2-band setup, with booms, 15-foot vertical support and feedlines, adds up to less than 5 pounds. It stows in a canvas golf bag, as seen in Fig. 9-61. In more-or-less this form they have served the author in v.h.f. portable ventures from Maine to California and the Canadian Rockies to Florida. They evolved gradually, being made ever lighter and easier to set up, until it is felt that they are now near the ultimate in portability and performance.

The 3-Element 50-Mc. Yagi

The booms are ¾-inch aluminum tubing, originally two 6-foot pieces, now cut into 4 3-foot sections. An insert at the center made of ⅝-inch copper pipe joins two 3-foot pieces to make a 6-foot boom. This size is a nice sliding fit inside ¾-inch hardware store aluminum tubing. A TV-type U clamp holds the boom to the vertical support. Drilled as shown in Fig. 9-59 the boom supports the center sections of the elements for either beam. The 50-Mc. elements are ¼-inch aluminum tubing, with their ends drilled one inch deep with a No. 5 drill, to take collapsible whips that make up the balance of each element. Various sizes of these whips can be found in distributor catalogs. Ours are Lafayette Type 99-C-3005, 0.210 inch in diameter at the base, and 47 inches long, extended.

Because these whips are very thin at the outer ends, the elements must be made several inches longer than would be the case for normal element diameters.

The driven element is 120 inches, the reflector 124 inches, and the director 114 inches. With 47-inch whips inserted to a depth of one inch, center sections should be as follows: reflector 32 inches, driven element 28 inches, and director 22 inches. The elements can be held in place with clips in the manner shown for several other antennas in this chapter, or self-tapping screws can be run through the top of the boom, to bear on the element midpoint. The ends of the center sections are slotted with a hacksaw, and a small wrap-around clamp is used to squeeze the ends tightly around the whip.

The array can be fed conveniently in two ways, as shown in Fig. 9-60. The simpler is the

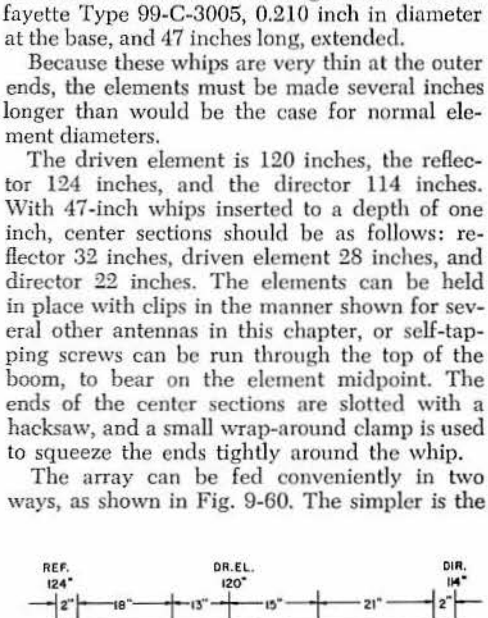


Fig. 9-59—Details of the portable beams for 6 and 2. One boom can serve for either array, or two sets of hardware can be made, permitting simultaneous operation on two bands. The hole just back of the U-clamp holes is used for both 6- and 2-meter elements.

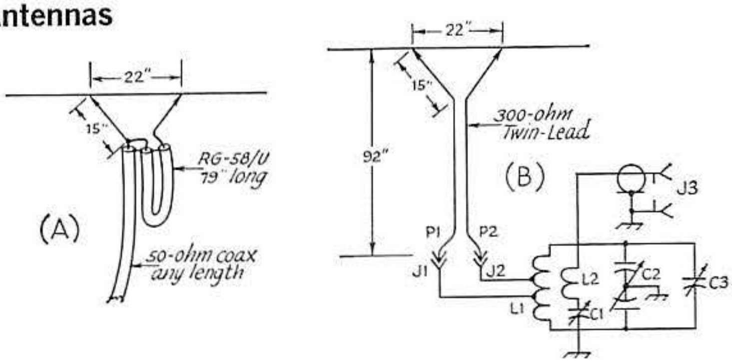


Fig. 9-60—Two methods of feeding the 50-Mc. portable array. A half-wave balum and a delta match of flexible wire are shown at A. The Twin-Lead delta and line, with adjustable antenna coupler, B, permits use of the array over a wider frequency range. With readjustment, it provides a constant load for the transmitter, from 50 to 52 Mc.

C_1—75-pf. miniature variable (Hammarlund MAPC-75B).
C_2—11-pf. miniature butterfly variable (Johnson 160-211).
C_3—30-pf. miniature mica trimmer (ARCO)
J_1, J_2—Insulated tip jack.
J_3—BNC fitting.
L_1—15 turns No. tinned, ½-inch dia., 16 t.p.i. Tap at 3½ and 11½ turns.
L_2—3 turns insulated hookup wire, around center of L_1. Coupler is assembled in a 1⅝ by 2 by 3¼-inch Minibox, with the tip jacks at one end and the coaxial connector at the other.

delta and balun, A. This works fine at the design frequency, but the s.w.r. rises quickly on either side. If an antenna coupler is used, at B, the useful frequency range of the array can be extended considerably. A small coupler unit can be connected at the bottom of the delta, and coax run from there to the equipment, or part of the transmission line can be Twin-Lead or other balanced line. This can be any length, though the electrical half wavelength shown is a good system.

With the short lines normally used in a portable setup, this array works extremely well, and its gain and directivity are very helpful in working out with low-powered gear. When the coupler is used, and retuned for each frequency change, the system can be matched perfectly over the first two megacycles of the band. Gain is 6 db. or better from 50 to 51.5 Mc., and in excess of 7 db. in the most-used lower part of the band.

The 5-element 2-Meter Beam

The boom for the 144-Mc. antenna can be the one used for 6-meter operation, or it can be another made similarly. The center sections of the 2-meter antenna are ¼-inch aluminum rods, with their ends drilled and tapped 6-32, to a depth of ½ inch. Element extensions are ⅛-inch aluminum wire or welding rod, threaded 6-32 at the inner end. To prevent confusion in assembling the antenna, the extensions are all the same, and the center sections are made progressively shorter, back to front, to give the optimum element lengths. The dimensions given in Fig. 9-59 are for best performance from the low end to about 145.5 Mc.

A delta and balun are used with this array also. The arms of the delta are 4 inches long, including the alligator clips used for connecting to the driven element. Pieces of zip cord make flexible long-lasting delta arms. The balun loop, of RG-58/U coax, is 27 inches long.

ANTENNAS FOR 220 AND 420 MC.

Since physical size must be maintained as we go to progressively higher frequencies, if communications effectiveness is to be maintained, the number of elements in large arrays for 220 or 420 Mc. tends to stagger the imagination of users of lower bands. On the other hand, a half wavelength becomes of such proportions (roughly 2 feet on 220, 1 foot on 420) that high-performance beams are not difficult to build and adjust.

Antenna work at these frequencies is an absorbing field in itself. There is no better way to work out pet antenna ideas than with models built for 420 Mc. If reasonable care is used in

Fig. 9-61—Beams for 6 and 2, with 15-foot vertical support and all hardware, ready for travel.

scaling significant dimensions, models can supply answers in a few hours of interesting work that would take days (and dollars) to work out on some lower frequency—if, indeed, they could be developed satisfactorily there at all, by amateur methods.

One frequently hears the statement, "Yagis don't work on 420!" This is a completely false notion that is mainly the result of the workers having neglected the question of scaling. Yagis and most other kinds of antennas work just as well on 220 and 420 Mc. as on any lower frequency, if they are made and fed properly. Anyone interested in working the full band, 420 to 450 Mc., would be foolhardy to use them, but where operation is confined to a narrow segment like 432 to 436 Mc., for example, frequency response should be no problem. This is likely to hold for 220-Mc. work as well, as activity is frequently channelled to one part of that band, by mutual local agreement.

Line losses *are* a problem at these frequencies. As pointed out in Chapter 8, the best line may be none too good. Matching is important, particularly if any real antenna evaluation work is to be undertaken. Line radiation is troublesome, and must be countered with every reasonable precaution.

Basic antenna systems already described work well on both 220 and 420. Collinear arrays may be built using the midband dimensions of Table 9-I and the mechanical arrangements suggested for 144-Mc. arrays, except that most parts can be of smaller cross-section and lighter in weight. In all-metal construction of collinears or Yagis, element and boom diameters should be kept in scale with the 144-Mc. designs, insofar as possible. A 12-inch element running through a 1¼-inch metal boom, for example, is out of place in a 420-Mc. array. Hard-drawn aluminum wire or welding rod, preferably not larger than ⅛ inch in diameter, is good for elements. Booms no larger than ½ inch are recommended, if made of metal.

Matching methods follow the basic principles detailed in Chapter 8, but the small physical proportions of the adjustable Q section and the universal stub make these probably the most desirable matching devices.

Plane and corner reflectors assume practical proportions. Where broad frequency response and high front-to-back ratio are desired, driven elements backed up by a wire grid or metal screening make a fine antenna. This is particularly inviting where two bands must be taken care of with a single structure, since elements can be mounted on opposite sides of such a reflecting plane. The gain of a screen reflector array is little affected by the spacing of the elements from the reflector, but there is a marked change of impedance when the spacing is changed. (See Fig. 8-9, 180-degree curve.) From this it is clear that matching to a transmission line can often be achieved by selecting a suitable spacing by experiment, if the impedance of the system is not known.

WOOD-BOOM YAGIS FOR 220 AND 432 MC.

Moderate-size Yagis for the 220 and 420-Mc. bands can be built at very low cost, and with only simple tools, if the suggestions of Figs. 9-62 and 63 are followed. Booms are 1 by 1-inch wood, available in any lumberyard. (Your dealer will call it "one by one" but the actual size will be more like ⅞ by ⅞ inch.) Be sure that it is straight, dry and free of knots. Take the man's advice as to which kind of wood will be best for out-door use, as available stocks vary around the country. Ours was red cedar. Prime and paint it well, if you want long life.

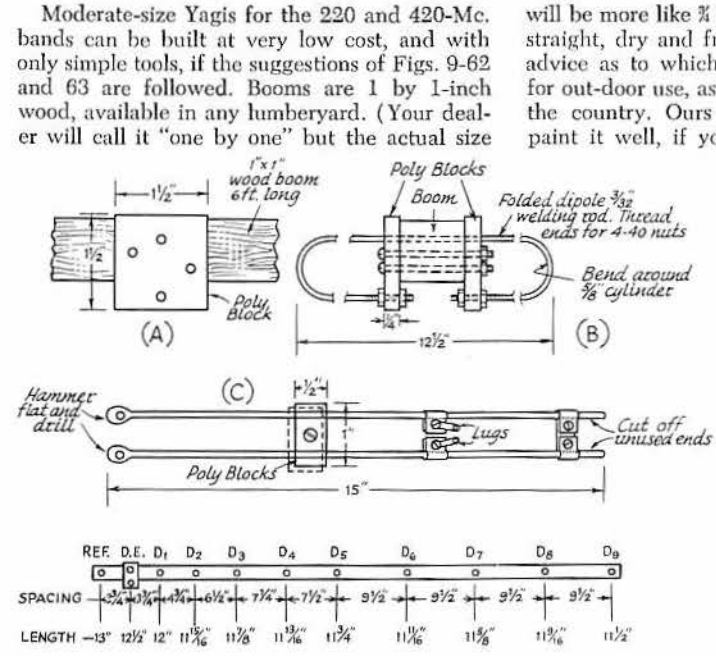


Fig. 9-62—Details of a 6-foot 11-element Yagi for 432 Mc. The square boom and one polystyrene mounting block for the driven element are shown at A. The blocks, element and boom are shown in detail in the end view, B. Matching stub, C, fastens to ends of the driven element, and is mounted under the boom between two poly blocks. Element lengths and spacings for the middle of the 420-Mc. band are shown in the side view of the complete array.

An 11-element array is shown for 432 Mc., and a 7-element one for 220 Mc., both using element spacings and lengths that are close to optimum for gain. The antenna should be supported near its mechanical balance point, roughly 2 feet from the reflector end. If a TV-type U clamp is used, it is well to bend up a U-shaped metal plate the width of the boom and about 3 inches long, and slip it over the boom at the point where the holes are to be drilled for the clamp. This protects the boom from crushing when the U-clamp nuts are tightened, and leaves it strong enough to stand up well without bracing.

Parasitic elements in the models tested were made of 3/32-inch aluminum welding rod, which can be purchased very reasonably at welding supply houses. Usually it comes in 3-foot lengths. Any stiff wire or rod up to ⅛ inch diameter will do. Drill the boom with a size that will just take the elements with a force fit, then run a ½-inch brass or aluminum screw into the boom to bear on the element and hold it in place. The screws can then be bonded together and connected to grounds for lightning protection, if desired.

The driven elements originally tried were step-up folded dipoles similar to those used in the 144-Mc. Yagis, but it was found that these did not work well at 220 and 432 Mc. This is probably the result of the spacing between the two parts of such a dipole being a considerable portion of a wavelength at these frequencies. The 432-Mc. Yagi was made with a driven element of the same material as the parasitic elements, mounted as shown in Fig. 9-62 A and B. Blocks of insulating material ¼ inch thick and 1½ inches square are fastened to the boom with two 1½-inch brass screws and nuts. The upper portion of the dipole runs through the boom, just above the center, and the lower is held in place with 4-40 nuts on either side of the insulating plates, as shown in the end view, B. The 3/32-inch rod is easily threaded for 4-40, if this is done before the element is bent. The total length of the wire is about 25 inches. An alternative to threading is to hammer the ends flat, and drill for 4-40 screws.

The antenna is matched by means of a universal stub, C, made of the same material as the elements. It should be cut about 15 inches long, and suspended under the boom. An adjustable short and two sliding clips for connecting the transmission line or balun are provided for adjusting the matching. The ends of the stub that connect to the dipole are pounded flat with a hammer, and then drilled to pass the threaded ends of the dipole. These are held in place by the 4-40 nuts shown in *B*. A ceramic cone standoff insulator (not shown in the drawing) is fastened to the underside of the boom. Two pieces of polstyrene similar to that used for the dipole mounting blocks, one above and one below the matching stub, are fastened to this cone, clamping the stub in place.

The short and the point of connection of the balun are adjusted for zero reflected power, as indicated in an s.w.r. bridge connected in the line. The bridge should be at a point in the line a multiple of a half-wavelength from the antenna, for the greatest ease of adjustment.

The 220-Mc. Yagi can be made in the manner just described, using a dipole made of a single piece of wire. A variation of the ratio-type folded dipole was made for the 220-Mc. antenna as shown at B in Fig. 9-63. Here a flat strip of aluminum comprised the fed portion of the dipole, and a ⅜-inch tube the unbroken portion. The aluminum strip is bolted to the underside of the tubing at the outer ends. The slope down to the feed point at the polystyrene blocks determines the impedance. With the dimensions shown the array can be fed with 52-ohm coax and a balun, connected to the lugs at the insulating plates. The s.w.r. is under 1.5 to 1 from 220 to 224 Mc., with optimum match at about 221.5 Mc.

6-Bay 220-Mc. Yagi Array

Six of the 7-element Yagis of Fig. 9-63 can be combined in a high-gain 220-Mc. system with

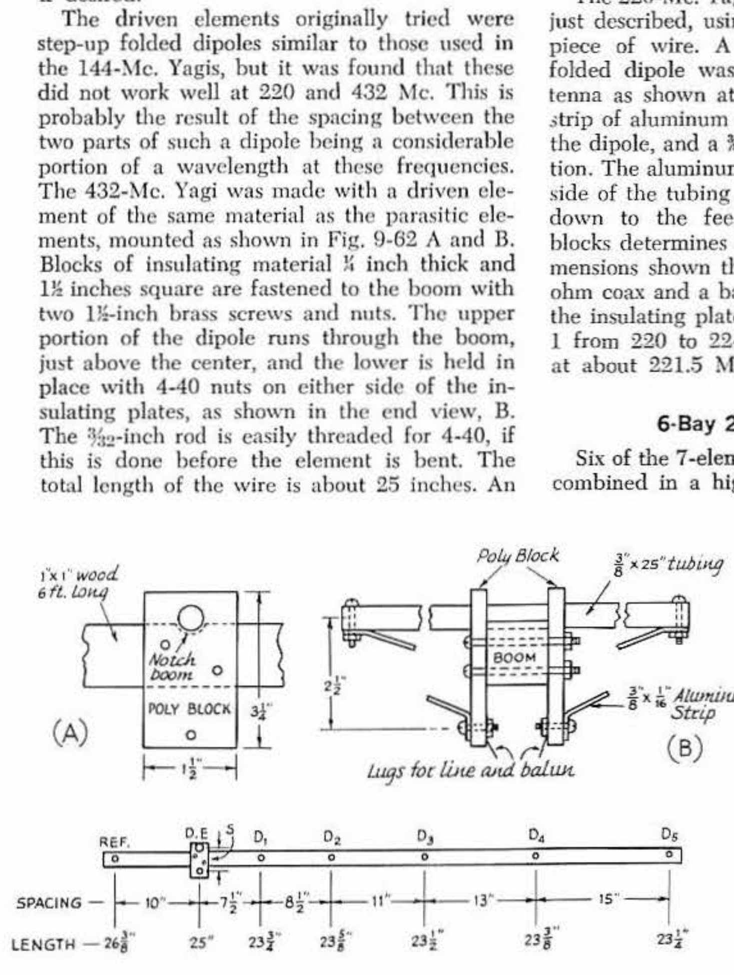


Fig. 9-63—7-Element 220-Mc. Yagi on a 6-foot wood boom. Poly blocks each side of the boom support the modified folded-dipole driven element. Latter has sloping lower portion, for matching 52-ohm coax and balun, connected to lugs at the bottom of sketch B. With element lengths and spacings given in the side view of the array, optimum performance is obtained over the first 3 to 4 megacycles of the band.

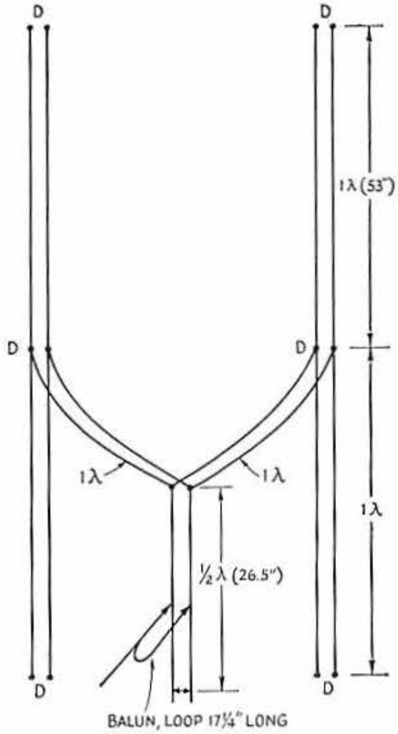


Fig. 9-64—Phasing and matching harness for the 220-Mc. array. The 6 driven elements are connected at points marked D. No transposition of the lines is made at any point. Main transmission line can be any impedance, if point of connection on the stub is adjusted for minimum s.w.r.

the phasing and matching arrangement shown in Fig. 9-64. This has also been used with commercially-made Yagis of short or intermediate lengths. If the booms are much over one wavelength long, however, more gain can be obtained by going to two wavelengths spacing between bays, as in the 432-Mc. array of Fig. 9-65.

The feed impedance of the individual bays need not be known for combining them in this way, so long as they are all the same, and are intended for balanced feed. In the H-shaped phasing system all parts are the same length (one wavelength, or 53 inches). Connection points for the 6 driven elements are indicated by *D*. The universal stub, lower center, tunes out any small reactance and provides matching for the main feedline, regardless of type or impedance. It is two pieces of ¼-inch copper tubing, spaced 1¼ inches, 28 inches long, mounted on ceramic standoffs attached to aluminum clamps on the main support.

No precise dimensions are given for bay spacing, as this is not critical and can be determined by the way the array goes together wher phasing lines of the right length are used. Bays should be tightened in place where the lines are without appreciable slack. Do not put tension on the phasing lines, as this will pull the booms out of alignment. The horizontal spacing leaves approximately ⅝ wavelength between the inner element ends.

2-BAY AND 4-BAY ARRAYS FOR 432 MC.

The 11-element Yagi of Fig. 9-62 can be used effectively in stacked pairs, or in a 4-bay system, as shown in Fig. 9-65. For convenience in stacking, delta-matched driven elements are used, with open-wire phasing lines and a universal stub. All director elements are as in Fig. 9-62. The driven element is 13 inches long, and the reflector is 13¼ inches. Essential mechanical details of the supporting frame are shown in Fig. 9-66.

The dimensions of the fanned-out portions of the phasing lines are not particularly critical, so long as all four are the same. The sides of the delta are about 2 inches long, and they fasten to the driven element about 1⅜ inches each side of the center point. The phasing lines can be ½-inch spaced TV line, though No. 12 or No. 14, with a minimum of spreaders, preferably Teflon rod, will be much better and more durable. The lines should be supported on standoffs at several places, to prevent flexing at the connection points.

Construction

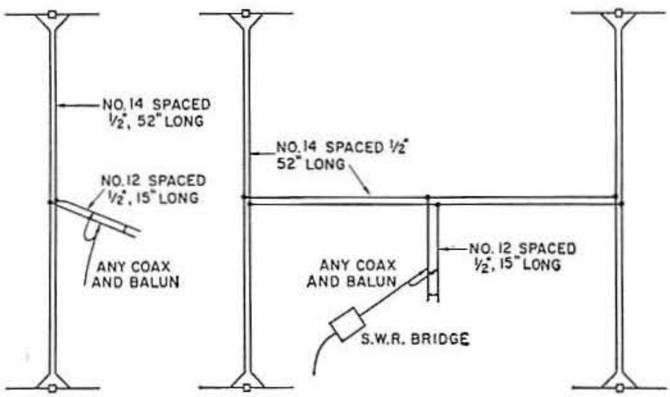
All-wood construction was used for low cost, ease of assembly, and freedom from worry over large amounts of metal in the field of the array. Lightweight wood design would be none too strong for large arrays on lower frequencies, but at 432 Mc. the wood frames are sturdy enough to stand up longer than most u.h.f. enthusiasts will want to stay with one array.

The wood is mostly 1 × 1 stock. Like all lumber dimensions for width and thickness, this is a misnomer. The actual size is likely to be more nearly ⅞ by ⅞ inches, but this is not important for our purposes. It merely makes it impossible to give precise dimensions for the supporting frame. Get good-quality dry wood, free of knots, and preferably a kind that is not subject to severe warping. Most lumber dealers will be glad to advise you on the best materials for outdoor use, and available woods vary around the country.

The holes for the elements are drilled the size of the elements or slightly smaller, and the elements are forced into place. Half-inch brass wood screws that run in from the top or bottom hold the elements in position firmly.

Bracing can be whatever the wind and weather conditions in your locality demand. The principal feed details of the array are given in Fig. 9-65. At the left is the assembly for two of the 11-element bays. The main vertical mem-

Fig. 9-65—Phasing arrangements for two and four 11-element Yagis. Bay spacing of approximately two wave lengths is set by the length of the phasing lines. The universal stub matching device may be used with any type of transmission line, as well as with the coaxial line and balun as shown.



ber, also 1 × 1, is held perpendicular to the booms by means of gusset plates of ¼-inch Masonite, as shown at the right. If only an 11-over-11 is to be built, this vertical member can be dispensed with, and the bays clamped to the main vertical support by means of U clamps.

When four bays are to be used additional bracing is needed, and the gusset plates and forward bracing become necessary. The front brace is ½ × 1-inch stock, bolted between the two booms to keep them in alignment. The two vertical supports with the gusset plates are tied together horizontally with two 1 × 1-inch cross braces and a 1× 2-inch main support, as shown. Not shown in the sketch are two ½ × 1-inch wood sway braces that run from the mid-points of the two forward vertical braces to the 1 × 2-inch main horizontal member. These are held in place by small brackets cut to fit from sheet aluminum. The main vertical support, not shown, is 1¼- or 1½-inch round closet-pole stock. This is inexpensive and strong, and there is no extraneous metal in the array proper.

To make the wood members reasonably durable and waterproof they were sprayed with Krylon before assembly. The Masonite gusset plates were also well soaked with lacquer spray. The whole assembly was painted with ordinary outside white house paint.

Adjustment

Matching the array should be done with the bottom bay at least four feet above ground, if in the position that it will be in use; that is, with the booms horizontal. The region in front of the array should be free of trees, buildings, wires or any other materials or objects that can reflect 432-Mc. energy. A high-gain array has a strong field out front. An appreciable reflection back has a marked effect on its impedance. If you don't have a good large open area, prop the array up with the vertical support in a horizontal position, and the four booms pointing straight up. Ground under the array will have little or no effect on its impedance in this position, as the power radiated off the back is negligible, for this purpose.

With an s.w.r. bridge in the coaxial line near the antenna (preferably some small multiple of a wavelength away), adjust the short on the universal stub and the point of connection of the balun for zero reflected power. Once the proper points are found, permanent connections can be made.

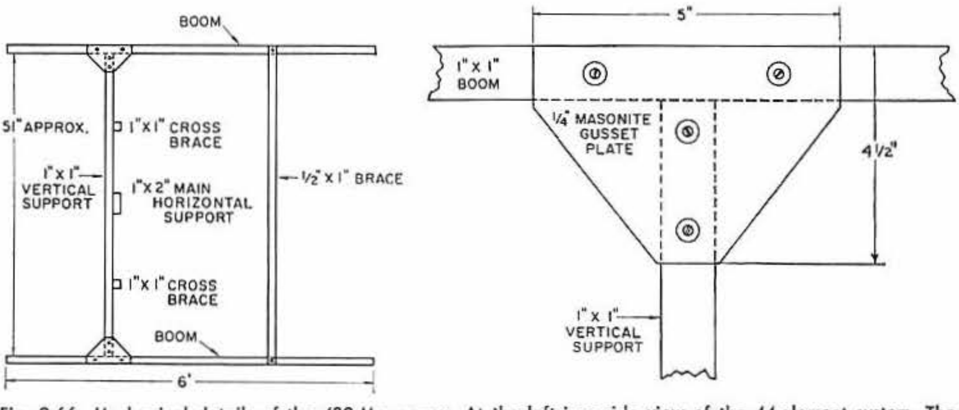


Fig. 9-66—Mechanical details of the 432-Mc. arrays. At the left is a side view of the 44-element system. The Masonite gusset plates used to hold the array in alignment are made as shown at the right. The array is supported on a round wooden closet pole, fastened to the three horizontal members in the sketch at the left, by means of U clamps.

A 48-ELEMENT COLLINEAR FOR 432 MC.

The high-gain 432-Mc. collinear of Figs. 9-67 and 68 is inexpensive, and easy to build and adjust. Though light in weight and mostly made of wood, it survived two rough winters in a hill-top location. Materials can be obtained anywhere, and only the simplest tools are needed in its construction.

Basically it is made up of four 12-element collinears, each having six half-wave driven elements, with reflectors. These are arranged in a square and connected by 2-wavelength phasing lines of open TV line, spaced ½ inch. To help to clarify this, in Fig. 9-67 the driven-element sets are numbered 1 and 2, at the left, and 3 and 4, at the right. The phasing harness is shown separately, with terminals correspondingly numbered. At the center is a universal stub (see Fig. 8-18), permitting the array to be fed with any balanced line, or coax and a balun.

We have ignored the 24 reflector elements so far, as they are not connected electrically, and showing them only complicates any drawing. They are 13¼ inches long, and placed approximately 5½ inches in back of the driven elements. The latter dimension has little significance as far as performance is concerned.

Most details of the supporting frame are in Fig. 9-68. The four vertical members, *A*, the two smaller horizontal braces, *B*, and the 24-element supports, *C*, are all "1 by 1" stock. This won't cost much, so get the best you can; free of knots and well dried. Take your lumberman's advice as to the kind of wood; the best available varies from one section of the country to another. As with all lumber, the actual size is less than the trade size, so the assembling can be done with 2-inch brass 6-32 screws, washers and nuts. These are indicated in the front view.

The element supports, *C*, are held in place with 2-inch brass wood screws. Drill holes in the *A* pieces just large enough to pass the screws, and in the ends of the *C* pieces about half the screw diameter. Now drill all the element holes. Paint all surfaces with a priming coat, and let it dry thoroughly. Coat the inner end of the *C* pieces with a good glue or cement, and screw them in place firmly—but don't overdo it. After the glue is dry, paint the frame again.

The elements can be about any conducting material: aluminum wire, welding rod or what-have-you. Put in the reflectors first. They are not centered exactly in the supports, but rather are placed so that their inner ends are about ¾ inch apart. Put a ½-inch brass wood screw into each support to bite lightly into the element and hold it in place. This is also done with each driven element.

Each set of six half-wave elements and interconnecting lines requires only four pieces of wire. At the upper left of Fig. 9-67, *E*, *F*, *G* and *H* are one piece, and *J*, *K*, *L* and *M* are another. Elements *N* and *O* are made from pieces of wire about 13½ inches long. Pound one end of each flat on an anvil, wrap it around a wire or rod of the same diameter, and drill through the overlapping flat portions for a 4-40 screw. This makes a loop that can be clamped tightly to the midpoint of the interconnecting lines.

For *E-F-G-H* and *J-K-L-M*, cut pieces about 52 inches long, feed them through the proper holes in the *C* supports, and bend so that the elements are 12½ inches long, with inner ends about ½ inch apart, and the phasing sections are arranged as shown in the sketch. Spacing at the cross-over points can be done with ⅜-inch nylon or teflon rods, about 1¼ inch long, and either drilled or notched to pass the phasing wires. Drilling is best, but it creates somthing of a "threading" problem in assembling the elements and phasing leads. We gave it up after an initial try, and resorted to notched insulators, cemented in place with epoxy glue. They have never come loose in nearly two years on the tower. Spacing of the lines should be maintained at ¾ inch or less.

Elements *N* and *O* are wrapped around the

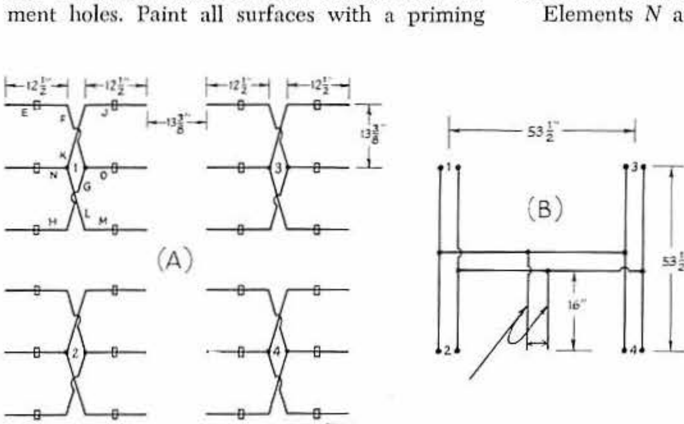


Fig. 9-67—Element details and phasing harness for the 48-element 432-Mc. collinear array. Reflectors, not shown, are 13¼ inches long, with their inner element ends ¾ inch apart. Phasing harness, B, should be spaced no more than ½ inch. Main transmission line can be coax or balanced line.

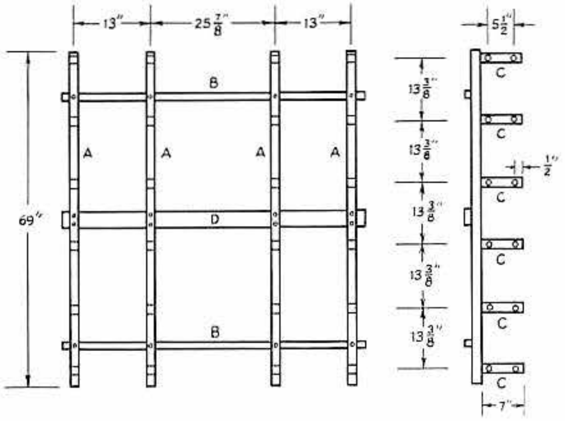


Fig. 9-68—Details of the wooden supporting frame for the 432-Mc. collinear array. Vertical members, A, are 1 by 1 by 69 inches; horizontal braces, B, 1 by 1 by 54 inches. Element supports, C, are 1 by 1 by 7 inches. Center brace, D, is 1 by 2 by 54 inches. Assembly dimensions are given on the drawing.

phasing leads and clamped in place. Trim each to 12½ inches over-all length, if necessary. When all elements are in place and they and their phasing sections are lined up properly, put in ½ inch brass wood screws bearing on each element to hold it permanently in position in its *C* support.

The phasing harness, 1-2, 3-4, and connecting line, is made of ½-inch open TV line, each piece 53½ inches long, this being 2 wavelengths as checked out for resonance with a dip oscillator at 432 Mc. Connection is made to the four sets of driven elements by soldering to lugs under the nuts that hold the center elements in contact with the aluminum phasing leads. The lines are longer than needed, physically, but the extra length permits looping them back to supports attached to the frame. TV hardware is useful here. Make all phasing-line bends on as large a radius as possible, and try to keep junctions perpendicular.

The universal stub can be any convenient material. The element wire was used in the original, 16 inches long, with the conductors spaced about 1 inch. When the proper points for the sliding short and the line connection are found, be sure that tight permanent connections are made. The excess of the stub below the short can be cut off, and the stub grounded for lightning protection, if desired. In the original the stub is mounted on brackets attached to the center vertical support, not shown in the sketch.

A U-shaped clamp of sheet aluminum holds the brace *D* to the supporting mast. Sheets of metal on each side of the brace prevent damage to the wood when it is clamped tightly in its final position. A similar but smaller clamp and plate arrangement was made for the lower *B* member, to steady the assembly.

Adjustment for matching was done with the beam in a horizontal position, aiming straight up. Two different feed systems have been used: a coaxial line and balun, and a balanced line. With the latter, tubular transmitting Twin-Lead runs from the stub to an anchor at the top of tower. Here it joins a 100-foot run of No. 12 wire, ½-inch spaced on teflon spreaders, one every 6 feet. This line is pulled up tightly and is straight throughout. At the house end there is a grounded adjustable short, and a balun connection, to complete the run to the antenna relay. The short and balun connection are adjusted for zero reflected power on this short section of coax. A mild s.w.r. exists on the balanced line, but this is of no consequence, the total loss having been measured at under 1.6 db.

The array has a very sharp main lobe, with two fairly large minor lobes on either side—an inescapable consequence of the wide spacing between bays. Response is sufficiently broad for uniform results over at least 4 Mc. in the vicinity of 432 Mc., when the stub is adjusted at the middle of the desired frequency range.

THE CORNER REFLECTOR

Corner and plane reflector principles were discussed in some detail in Chapter 8. The corner requires but one driven element, and is capable of giving a very clean pattern and moderate gain, with very broad frequency response. At 220 and 420 Mc. its size begins to assume practical proportions.

The corner angle can be 90, 60 or even 45 degrees, but the side length must be increased as the angle is narrowed. The driven element spacing from the corner can be anything from 0.25 to 0.7 wavelength for a 90-degree corner, 0.35 to 0.75 for a 60-degree one, and 0.5 to 1 for a 45-degree corner. Feed impedance for various corner angles and spacings was given in Fig. 8-9. Since the spacing is not critical as to gain, variation of it may be used to achieve impedance matching.

Gain with a 60- or 90-degree corner with 1-wavelength sides runs around 10 db. A 60-degree corner with 2-wavelength sides has about 12 db. gain. It will be seen that this is not outstanding for the size of such an array, but there are other advantages. A corner may be used for several bands, for example, or perhaps for u.h.f. television reception, as well as for amateur u.h.f. work.

A suggested arrangement for a corner reflector system is shown in Fig. 9-69. Sheet metal or wire mesh may be used with equal effective-

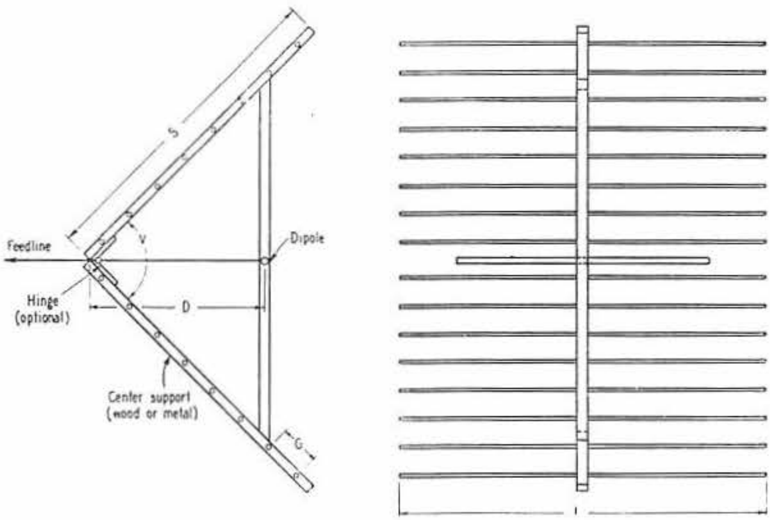


Fig. 9-69—Construction of a corner-reflector array. Frame can be wood or metal. Reflector elements are stiff wire or tubing. Dimensions for three bands are given in Table 9-II. Reflector element spacing, G, is the maximum that should be used for the frequency; closer spacings are optional. Hinge permits folding for portable use.

ness for the reflecting plane. A series of spines, as shown, is equally good, if the space between them is kept under 0.06 wavelength at the highest frequency for which the reflector is to be used. The frame may be made of wood, with a hinge at the corner to facilitate portable work or assembling atop a tower. Principal dimensions for corner reflector arrays for 144, 220 and 420 Mc. are given in Table 9-II. These dimensions are not at all critical, because of the broad frequency response of any plane-reflector system.

TABLE 9-II

Dimensions of Corner-Reflector Arrays for 144, 220 and 420 Mc.

Band (Mc.)	*Side Length "S" (Inches)*	*Dipole to Vertex "D" (Inches)*	*Reflector Length "L" (Inches)*	*Reflector Spacing "G" (Inches)*	*Corner Angle "V" (Degrees)*	*Feed Impedance (Ohms)*
144*	65	27.5	48	7¾	90	70
144	80	40	48	4	90	150
220*	42	18	30	5	90	70
220	52	25	30	3	90	150
220	100	25	30	screen	60	70
420	27	8⅜	16¼	2⅜	90	70
420	54	13½	16¼	screen	60	70

*Side length and number of reflector elements somewhat below optimum —slight reduction in gain.

BROAD-BAND ANTENNAS

In addition to utilizing harmonic resonances, as in most long-wire antennas, there are several ways to make an antenna work on more than one band. Most are variations of the broad-band dipole principle, in which the radiating element is modified in shape so that it has no precise *electrical* length. The "conical" and "bow-tie" antennas of v.h.f. and u.h.f. TV usage are familiar examples. In all such antennas the net effect is to make the transmission line gradually become the antenna, the point at which this happens varying with frequency.

A would-be v.h.f. enthusiast who must have an unobtrusive antenna that will work on 50 through 220 Mc. could do worse than to put up a TV conical. Its pattern will be far from perfect, and its gain low, but it will work after a fashion. Similarly, u.h.f. bow-ties and corner-reflector antennas have been pressed into service on 432 Mc. in a pinch. Performance of the

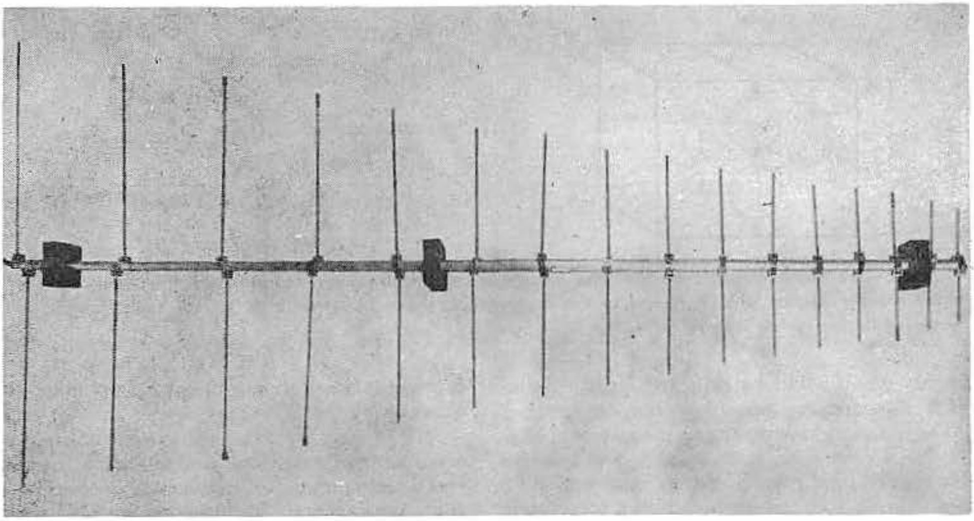


Fig. 9-70—The log-periodic array for 140 through 450 Mc. looks like a Yagi when viewed from top or bottom. Actually it has two electrically-separate booms, each with a set of elements arranged as shown in Fig. 9-71. Black objects are wood-block spacers for the booms. Design is by K7RTY.

TV conical on 50 Mc. may be improved slightly by extending the elements 3 or 4 inches.

A 3-Band Log-Periodic Array

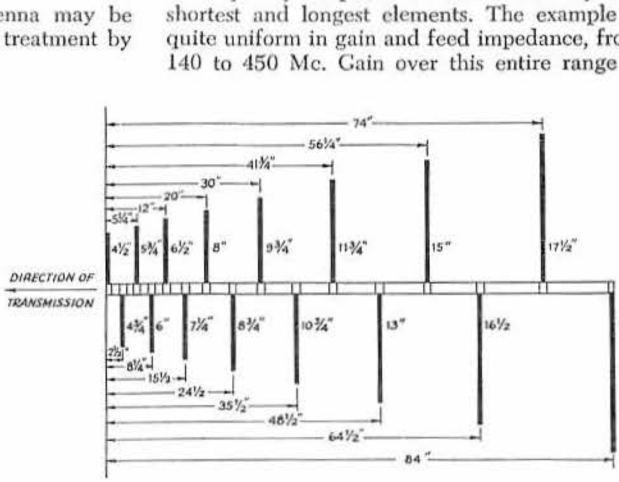
A more esoteric idea is the basis of the "log-periodic antenna," now widely used in military and commercial stations where many frequencies must be employed to maintain communication in the h.f. range. In theory, the frequency range of this type of antenna is almost unlimited, and in practice a spread of 4 or 5 to 1 is not uncommon. Arrays of this kind can take many forms. A simple version by K7RTY is shown in Figs. 9-70 through 73.

As author Heslin put it in his June, 1963, *QST* article: "This is an antenna whose resonance transfers smoothly from one element to the next, as the frequency is raised." More on the principle of the log-periodic antenna may be found in a November, 1959, *QST* treatment by Milner, W1FVY.

The version described here is not readily drawn or photographed in its complete form, to show full details. It has two booms one above the other, as shown in the sketch of the short (front) end, Fig. 9-72. Elements are progressively longer and wider spaced as we move toward the back of the array. The array is fed with coaxial line, which runs inside the lower boom its entire length. The outer shield connects to that boom, and the inner conductor to the upper. Each boom has a set of staggered elements, as shown in Fig. 9-71. These are assembled so that when the antenna is viewed from directly above or below it appears somewhat like a long Yagi, as in the photograph. The two booms are maintained 1¼ inches apart, by means of wooden blocks.

Frequency response is determined by the shortest and longest elements. The example is quite uniform in gain and feed impedance, from 140 to 450 Mc. Gain over this entire range is

Fig. 9-71—Dimensions for one section of the log-periodic v.h.f. array. The finished array consists of two of these sections, mounted one above the other, to give the appearance of Fig. 9-70.



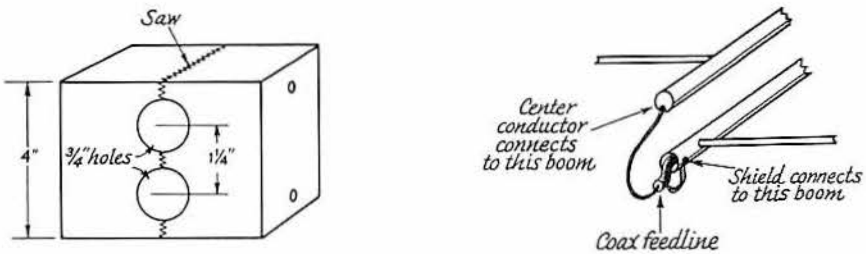


Fig. 9-72—The two booms of the wide-band array are kept in alignment and insulated from each other by three wooden blocks, left. Short end of the array, right, shows how the array is fed. Lower boom, with coax inside, acts as an infinite balun.

roughly what would be expected from a 3-element Yagi for any one frequency.

The element mounts were made from inexpensive TV antenna parts, modified to take an element with a threaded end, as shown. K7RTY used ¼-inch rod for elements, but other sizes are suitable if a mounting method is available. Note that two assemblies like that shown in Fig. 9-71 are required. These must be held in alignment, but insulated from each other.

An array of similar electrical properties was made by W1CUT, using ½-inch aluminum channel stock for the booms, and threaded-end ⅛-inch aluminum rod for elements. The coaxial line ran the length of one of the channels, apparently serving as an "infinite balun" in the same way as in the K7RTY version where the coax runs inside the boom. This array was tested and found to have an s.w.r. under 2:1 at 144, 220 and 432 Mc., and gain averaging 6 db. over this range.

The log-periodic, in common with all broadband arrays, does not give something for nothing. Gain is very low, compared with what a Yagi of the same physical size would deliver on one frequency, but the principal weakness is its broad-band nature. Being almost equally effective across more than a 3-to-1 frequency range, the antenna presents very much more of a problem with spurious receiver responses than does a Yagi, with the latter's inherent selectivity. By the same token, greater care must be exercised in keeping down spurious products in the transmitter. The log-periodic will accept power on any frequency in its wide range, and radiate it with some gain. This is quite a different matter from working with a Yagi that has gain only over a narrow frequency range, and considerable rejection ability on most other frequencies.

Still, if one antenna, with one feedline and one rotator, for several v.h.f. and u.h.f. bands (and perhaps the TV channels in between) is your requirement, the log-periodic will probably fill it as well as any one antenna can.

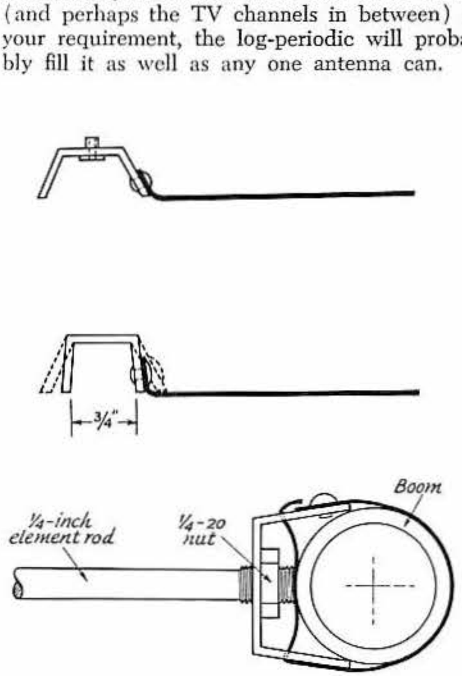


Fig. 9-73—Element mounts are made from TV line standoffs with stainless-steel straps, top. The insert is knocked or drilled out, and the clamp portion bent as shown at the center. Complete mount, with ¼-inch element in place, has lock and tension nut inside the clamp, to avoid need for drilling the latter.

Chapter 10

U.H.F. and Microwaves

Segments of the radio spectrum have labels. From 3 to 30 Mc. is "high frequency," 30 to 300 is "very high," 300 to 3000 "ultra high," and all higher "super high." But all through this book we've had trouble keeping within these semantic lines. This trouble with labels is, in a way, a capsule history of the radio art. Today's "ultra" is tomorrow's commonplace.

Webster defines ultra as "Going beyond others, or beyond due limit," yet within this writer's memory everything above 30 Mc. was called "ultra high." Our *QST* column started as "On The Ultra Highs" in 1939, but it wasn't long before we began to think of frequencies up to at least 220 Mc. as something rather below the "ultra" class. Today we even tend to take the 420-Mc. band out of this category.

Working on 420 is today approaching a routine business, but it is well to remember that the techniques we think of as "conventional" are that way because of continuing advances in tube design and in solid-state devices. We tend now to put the frontier somewhere around 1000 Mc. Who can say where it will be tomorrow?

Even a 1000-Mc. frontier is more philosophical than technical, challenging basic concepts of amateur radio communication. Throughout our history, amateurs have cherished the element of surprize. We call CQ to see who will come back. We listen, listen—everlastingly listen—for something new, or at least out of the ordinary. For all our emphasis on friendships made and maintained by radio, a strange voice, a new country, or even a new state stirs us. For reliable communication we tend to rely on the telephone. Worth for point-to-point has never done too well as a sales argument for new bands.

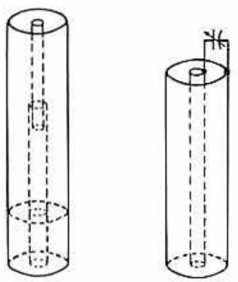
But communication above 1000 Mc. must, of necessity, be largely a person-to-person matter. Bands are incredibly wide. Beam patterns must be sharp, if communication over interesting distances is to be maintained. Cooperation in the matter of operating times, frequencies and beam headings is the only alternative to spending fruitless hours scanning the radio horizon for signs of life. Routine "activity" as we know it on lower bands seems thus unlikely ever to develop in our microwave bands, at least until we have space well filled with communications satellites.

We already have the know-how to transfer to the u.h.f. and microwave bands much of the talking we do on all-too-crowded lower frequencies. Will a new generation of amateurs seize upon the opportunities that the microwaves offer? It could be that the future of our avocation hinges on a positive answer.

Equipment and methods are already well within our capabilities. What we need most, in the world above 420 Mc., is people!

U.H.F. LINES AND CIRCUITS

The changing nature of tuned circuits as we move higher in frequency was discussed at length in Chapter 5. It may be well to review this material before going too far into u.h.f. circuitry. Earlier we were concerned with "lumped-constant" circuits; those where inductance and capacitance exist as separate entities. In u.h.f. work we have to think in terms of sections of transmission line.



(A) (B)

Fig. 10-1—A coaxial tank circuit can be tuned by means of a "false bottom," as at A, or by variable capacitance across the open end, as in B. In A the effective length is adjusted by means of the shorting disk, which is movable. Good electrical contact to both inner and outer conductors is important at this point.

Still higher in frequency we will abandon the line idea. Our circuits become resonant cavities, and an allied change occurs in methods used to transfer power from one circuit to another, and to an antenna. The parallel-conductor line disappears, giving way first to low-loss coax, and eventually to waveguide.

In Fig. 10-1 we have typical coaxial circuits, shorted at one end and open at the other. Such a line section an electrical quarter wavelength long can replace a coil-and-capacitor circuit as the tuning element of an oscillator, amplifier or other u.h.f. device. Its resonant frequency can be varied in several ways. Its length can be adjusted, if two pieces of tubing are made to telescope one inside the other, as in the upper portion of circuit *A*. A movable disk making firm contact to both inner and outer conductor can be moved up or down the line, as shown at the bottom of the sketch.

The line may be tuned with a variable capacitor, as in *B*. Adding capacitance in a tuning

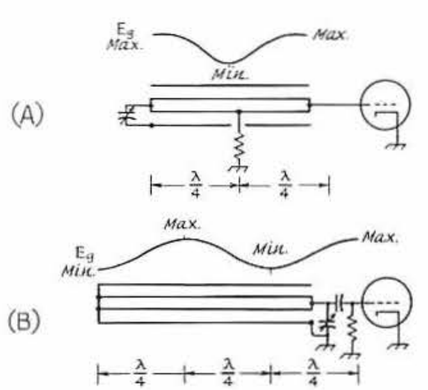


Fig. 10-2—The useful frequency limit of a coaxial-line tank circuit can be extended by making it a half-wave line, A, or three-quarter-wave line, B. R.f. voltage distribution along the lines is shown by curved line, E_g.

device, or in the form of the input or output capacitance of a vacuum tube, lowers the resonant frequency for a given length of line. The line must, therefor, be made physically shorter than a quarter wavelength. This loading limits the frequency over which a tube can be used with a quarter-wave line, just as in circuits discussed in Chapter 5.

The same steps can be taken to extend the frequency limit. A half-wave line of coaxial construction is shown schematically in Fig. 10-2A. A push-pull version could have a tube at each end. R.f. voltage distribution along the line is shown above it. Where a standing wave exists along a transmission line, the r.f. voltage and impedance are repeated every half wavelength. If tube capacitance and lead inductance tend to make us "run out of tank circuit" with a quarter-wave line, and a half-wave line is not convenient for our purposes, we can make the line any odd number of quarter wavelengths, as for example the ¾-wave line of *B*. This may have a quarter-wave resonance lower in frequency, but because of the different loading effect of tube and circuit capacitance at the two frequencies it will not be exactly one-third that of the ¾-wave mode.

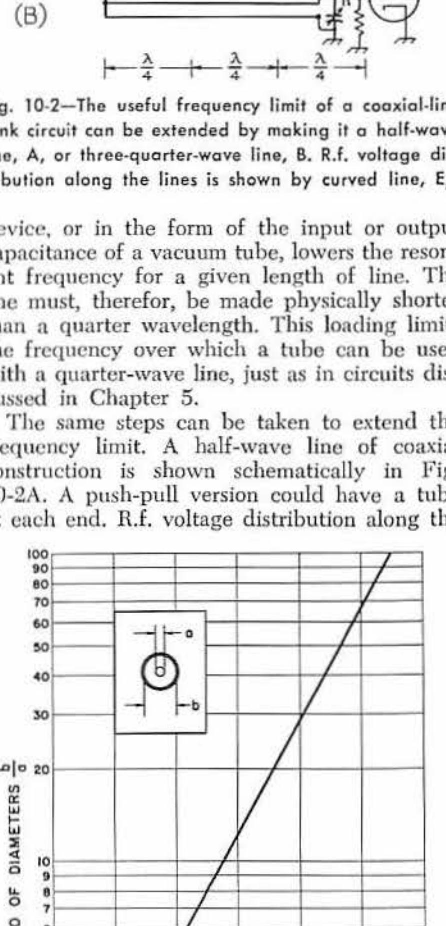


Fig. 10-3—Characteristic impedance of coaxial lines for various conductor diameter ratios. The outside diameter of the inner conductor and the inside diameter of the outer conductor are used.

Coaxial and Strip-Line Circuits

There is no need for the conductors to be round in cross-section, or truly coaxial in nature.

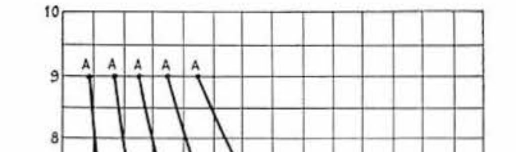


Fig. 10-4—Graph for determining the length of a capacity-loaded quarter-wave coaxial line of 71 ohms impedance, for frequencies from 150 to 1300 Mc. The value C includes tube output and tuning capacitances.

The outer can be rectangular, and the inner a flat strip of metal, with almost identical results. The strip line is often convenient for the amateur builder, and many examples appear in this book. Other shapes could be used, but coaxial or strip lines are most common.

The *Q* of these circuits is nearly always important, so the conductors should be of large size, and of metals having high conductivity. A coaxial line with a No. 20 wire inner conductor would be little better than a coil of the same wire size, for its ohmic resistance would be as high as the same wire wound into a coil.

Electrical conductivity is particularly important at points of high r.f. current (lowest r.f. voltage), notably at the shorted end of coaxial or parallel-line circuits. Insulation should be kept to a minimum, and preferably avoided entirely at or near the points of high r.f. voltage.

Insulation loss is introduced by a tuning capacitor, as it must be at a point of appreciable r.f. voltage. Movable-disk capacitors are favored, as they do not require insulating supports or metal frames that often introduce parasitic resonances.

Impedance of the line may be important in some applications. This can be obtained for coaxial lines from the formula:

$$Z_0 = 138 \log \frac{b}{a}$$

WHERE Z_0 is the impedance of the line, b is the inside diameter of the outer conductor, and a is the outside diameter of the inner conductor. Knowing the dimensions of available materials, the impedance can be obtained to an accuracy sufficient for most purposes from Fig. 10-3.[1] The impedance of a strip line can be obtained from the formula:

$$Z_o = 377 \frac{S}{W}$$

where S is the spacing between the strip and the outer conductor, and W is the width of the strip. Preferably W should be several times S. This information is from Brayley.[2]

The same author gave a formula for figuring the length of coaxial tank circuits when the total circuit capacitance is known. Solution of this formula for various capacitances, at frequencies from 150 to 1300 Mc., was worked out graphically by Garrett and Manly, as shown in Fig. 10-4.[3] The top curve in each set (C equals zero pf.) is for lines not capacitively loaded.

WAVEGUIDES AND CAVITY RESONATORS

Because of the loading effects of capacitance on parallel-line and coaxial circuits the useful frequency limit for any type of line circuit is somewhere around 1000 to 2500 Mc., depending on the tube or solid-state device used. At higher frequencies something different is needed, for transmission lines as well, for losses in even the best coaxial line become prohibitive. The answer lies in the use of cavity resonators and waveguides; devices in which the dimensions of conducting boxes or tubes determine the frequency at which they will operate.

Waveguide Principles

A waveguide is a metallic tube of circular or rectangular cross-section, through which electromagnetic waves can be transmitted. The walls of the waveguide are not considered as carrying current, in the sense of the conductors of a 2-wire line, but rather as a *boundary* confining the waves to the enclosed space. Energy injected at one end by capacitive or inductive coupling, or by radiation, flows through the guide to the load, by means of reflections from its inner walls.

There is an infinite number of ways in which the electric and magnetic fields can arrange themselves in a waveguide, depending on guide dimensions in wavelengths. These modes are separated into two groups: transverse magnetic (TM), and transverse electric (TE). The mode is identified by these letters, followed by subscript numerals, as $TE_{1,0}$, $TM_{1,1}$, etc. The number of possible modes increases with frequency, for a given size waveguide. The *dominant mode* (the only one for the lowest frequency the guide will pass) is generally used in practical work, as there is little point in using a larger guide than necessary for a given frequency.

Waveguide can be any cross-section, but only rectangular and circular are common. In the rectangular, Fig. 10-5, the width, x, is the critical dimension. It must be more than a half wavelength for the lowest frequency to be transmitted. Generally the height, y, is made about a half wavelength. It can be seen that waveguide has another advantage over other kinds of line, in addition to lower losses; by its very nature it can be an effective high-pass filter.

Five factors should be kept in mind in dealing with waveguide dimensions: *free-space wavelength*, λ; *guide wavelength*, λ_g, *the* actual length of the wave as it travels through the guide; *cut-off wavelength*, λ_c, longest *usable* wavelength, λ_u, that can be transmitted without excessive attenuation; and the *shortest* wavelength, λ_s, that can be transmitted before the next mode develops. These are obtained,

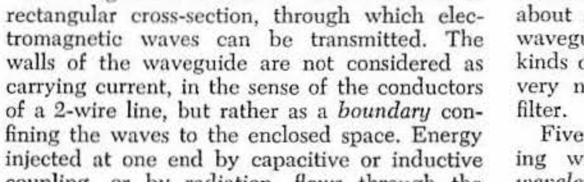


Fig. 10-5—Waveguide cross-section can be rectangular or circular. In the rectangular, the width, x, is the critical dimension. Cutoff wavelength is 2x or 3.41r.

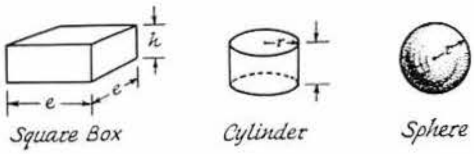


Fig. 10-6—Forms of cavity resonators. The square and cylindrical types are merely closed-off sections of waveguide.

for the dominant mode, as shown below:

$$\lambda,\ \text{inches} = \frac{11811}{\text{Freq., Mc.}}$$

$$\lambda_g,\ \text{inches} = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{\lambda_c}\right)^2}}$$

	Rectangular	Circular
$\lambda_c =$	2χ	$3.41r$
$\lambda_u =$	1.6χ	$3.2r$
$\lambda_s =$	1.1s	$2.8r$

Typical inside dimensions of rectangular waveguide for the various amateur bands are as follows: 2300 Mc. – 1.34 by 2.84 inches; 3300 Mc. – same; 5650 Mc. – 0.622 by 1,372 inches; 10,000 Mc. – 0.375 by 0.75 inch; 21,000 Mc. – 0.17 by 0.42 inch. These are standard waveguide sizes. If you were to make your own the dimensions would not have to be this precise.

The Cavity Resonator

Now suppose that we cut off a section of waveguide, and seal off the ends. We then have what is known as a cavity resonator. The term "cavity" is frequently applied to coaxial or strip-line circuits that are completely enclosed, but the name should be reserved for resonant boxes with no inner conductors, as shown in Fig. 10-6. The resonant frequency of these typical cavity shapes depends on the inside dimensions of the box and the mode of oscillation. The latter is comparable to transmission modes in waveguide. For the lowest-frequency mode, the resonant wavelengths are as follows:

Square box: 1.41 l
Cylinder: 2.61 r
Sphere: 2.28 r

The resonant wavelength of the cylinder or box is independent of the height, when this is less than half a wavelength. In other modes of oscillation the height must be a multiple of a half wavelength, as measured inside the cavity.

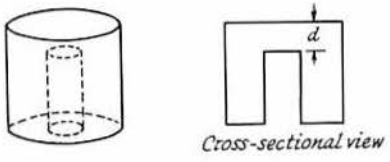


Fig. 10-7—Re-entrant cylindrical cavity resonator.

A cylindrical cavity can be tuned by means of an adjustable false bottom, when operating in such a mode. Other tuning methods include placing adjustable tuning paddles or slugs inside the cavity, so that the standing-wave pattern of the electric and magnetic fields can be altered.

Just as coaxial lines represented improvement in Q over coil-and-capacitor circuits near the upper limit of the useful frequency range of the latter, the resonant cavity is a means of obtaining very much better circuit efficiency at frequencies where the coaxial circuit begins to be inefficient or impractical. With care in the construction of a cavity as to silver plating and the best possible electrical conductivity where surfaces join, a circuit Q of several thousand is possible. A Q of 1000 or more is readily obtainable.

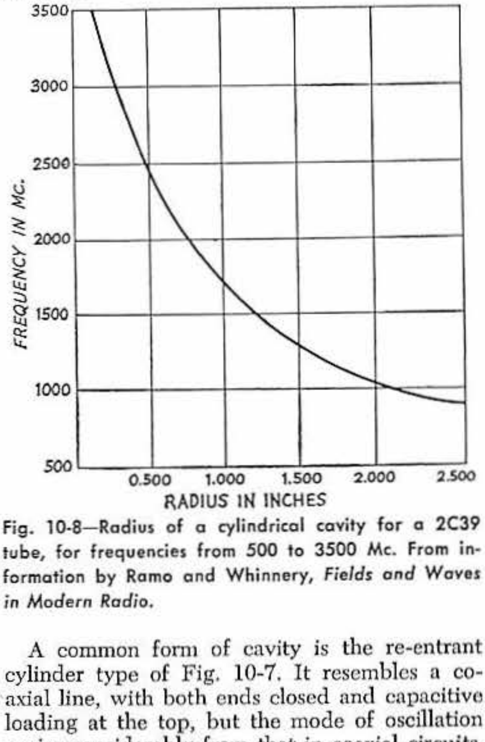


Fig. 10-8—Radius of a cylindrical cavity for a 2C39 tube, for frequencies from 500 to 3500 Mc. From information by Ramo and Whinnery, *Fields and Waves in Modern Radio.*

A common form of cavity is the re-entrant cylinder type of Fig. 10-7. It resembles a coaxial line, with both ends closed and capacitive loading at the top, but the mode of oscillation varies considerably from that in coaxial circuits. The resonant frequency of the re-entrant cylinder cavity depends on the diameters of the two cylinders, and the distance, d, between the cylinder ends.

A tube commonly used in amateur work in the u.h.f. region is the 2C39A or 3CX100A5. Fig. 10-8 makes possible a quick estimate of the size of a cylindrical cavity for this tube. The graph form was supplied by Garrett and Manly.[3]

Coupling to Waveguides and Cavity Resonators

Energy may be introduced into or extracted from a waveguide or resonator by means of

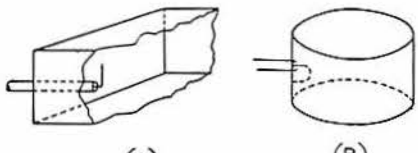


Fig. 10-9—Coupling to waveguides and resonators.

either the electric or magnetic field. The energy transfer frequently is through a coaxial line, two methods for coupling to which are shown in Fig. 10-9. The probe shown at A is simply a short extension of the inner conductor of the coaxial line, so oriented that it is parallel to the electric lines of force. The loop shown at B is arranged so that it encloses some of the magnetic lines of force. The point at which maximum coupling will be secured depends upon the particular mode of propagation in the guide or cavity; the coupling will be maximum when the coupling device is in the most intense field.

Coupling can be varied by turning the probe or loop through a 90-degree angle. When the probe is perpendicular to the electric lines the coupling will be minimum; similarly, when the plane of the loop is parallel to the magnetic lines the coupling will have its minimum value.

VACUUM TUBES AND TRANSISTORS FOR THE HIGHER FREQUENCIES

Even with the best possible circuits, the upper useful limit of frequency is determined finally by the characteristics of the devices connected to them. Capacitance between elements and inductance of leads brought out from them load down the circuit in the same way as wires and capacitors connected externally. In the tube, there is the seemingly infinitesimal time required for passage of electrons between cathode and plate. This may be only 0.001 *microsecond*, and of no importance at 1000 kc., but it is the equivalent of a full oscillation cycle at 1000 Mc. These facts set something around 3000 Mc. as the practical frequency ceiling for negative-grid tubes, and 1000 Mc. for inexpensive transistors.

Only specially-designed tube types will go anything like this high. At 420 Mc. we have left all but a very few types behind, and at 1215 Mc. no tube commonly used on lower bands is considered. Solid-state techniques for u.h.f. are promising, but suitable devices are costly, at present.

U.H.F. Receiving Techniques

In receiving, where power-handling capability is not a primary consideration, the useful frequency limit is somewhat higher than in transmitting, for more-or-less conventional designs. Probably the first mass-produced u.h.f. tube was the "acorn," a triode of very small element size, with leads brought out radially through the glass envelope. In suitable circuits it will oscillate up to around 500 Mc.

Today's Nuvistor carries reduction of electrode size and lead length close to the practical minimum. The inexpensive 6CW4 and 6DS4 work very well in our 420-Mc. band. The double-ended 8058, made for grounded-grid amplifier service, does even better.

Very high-transconductance planar triodes (416B, 7768 and others) are the best r.f. amplifier tubes available for 420-Mc. service, and they work fairly well even at 1215 Mc. in suitable circuits.[4] They are not mass-produced items, however, and consequently are rather expensive.

Actually there is little point in attempting to use vacuum tubes for receiving above 400 Mc., for even the best are incapable of satisfactory noise figure, compared with a well-designed transistor front end. Even at 432 Mc. a good crystal mixer followed by a low-noise i.f. amplifier will do just about as well as the best vacuum-tube designs. Transistor r.f. amplifiers are fast taking over the receiving job.

Conventional U.H.F. Transmitting Tubes

Several glass-envelope transmitting tubes work moderately well in the 420-Mc. band. The 6939 is a small dual tetrode much like the 6360, but with more compact design to extend the upper frequency limit for u.h.f. service. It will deliver about 6 watts output. The 6524 and 6252 are dual tetrodes with plate leads brought out through the top of the envelope. They are capable of around 15 watts output in the 420-Mc. band. The 5894 is a larger version of the 6252, widely used some years ago for 420-Mc. work. It will give up to 40 watts output at 432 Mc. Any of these tubes operated as a tripler from 144 Mc. will drive another as a straight-through amplifier, and they can be used with more-or-less conventional circuitry.

The best transmitting tubes for use at high amateur power levels in the 420-Mc. band are the external-anode types beginnng with the

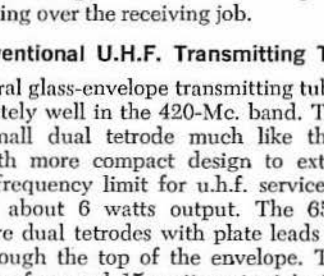


Fig. 10-10—Sectional view of the "lighthouse" tube. Electrode spacing and lead inductance are held to near the practical minimum, making the tube useful in the u.h.f. range.

4X150A, and continuing through later versions such as the 4CX250B and R, and the 4CX300A. These tubes are well adapted to coaxial-line circuitry, and they require forced-air cooling. A conduction-cooled version, the 8072, has considerable appeal for medium-power applications where use of a blower is not convenient. Many other types of this general tube family are available, but the ones mentioned above are most often seen in amateur u.h.f. circles. All require special sockets, which may include built-in bypassing, where desired.

Tubes generally available to amateurs for transmitting use above 1000 Mc. include the "lighthouse" triode, Fig. 10-10. This old but still useful type can be picked up inexpensively on the surplus market. It comes in several styles, of which the 2C40 is the most plentiful. The "pencil triode" series of tubes, not widely used by amateurs, work up to about 2000 Mc. The 2C39 planar triode, and later versions such as the 3CX100A5, can be used up to around 3000 Mc. in suitable circuits. The 2C39 tripler to the 1215-Mc. band was described in *QST* by W6DQJ.[5] WB6IOM described a 2-tube amplifier.[6] The 416B, more familiar to amateurs in receiving applications, is capable of operation up through the 3300-Mc. band in suitable circuits.

Even these u.h.f. types work poorly above our 2300-Mc. band, because of the fundamental limitations imposed by their physical dimensions. In our 3300-Mc. band and at all higher frequencies the negative-grid tube gives way to various specialized types such as the klystron, the travelling-wave tube and the magnetron.

VELOCITY MODULATION—THE KLYSTRON

At some point in the u.h.f. range, just where depending on power level and application, transit-time effects render conventional tubes unusable. Electron flow from cathode to anode, in tubes we're familiar with, is in the form of short bursts, regulated by the r.f. charge on the grid. When transit time is an appreciable part of the r.f. cycle these pulses become poorly defined, and performance falls off.

Of the many devices developed to get around this problem, the power klystron is probably the most interesting to the amateur. The klystron is complex and costly, and amateurs know it mainly as an occasional surplus market item, but if we are to do much with our assignments in the micro-wave region some understanding of its operation is mandatory. In one form or another, it has been used in nearly all amateur work above 3000 Mc.

The klystron uses the phenomenon of transit time to advantage, through a technique known as velocity modulation. It is capable of reasonable efficiency, high gain, good linearity and high stability. Its chief fault from the amateur point of view, other than high cost, is that the frequency-determining circuits are usually part of the tube itself. All too often, klystrons that look like surplus bargains turn out to be built for other than amateur frequencies. They are available for frequencies from a few hundred megacycles to many kilomegacycles, and for power levels from milliwatts to megawatts.

The Klystron Amplifier[7]

Referring to Fig. 10-11, the essential parts of an amplifier klystron are the electron gun, the *drift tube,* the *resonant cavities* and the *collector.* The cathode, anode, and focussing electrodes of the gun form an electron beam. Current flows as in a diode, when high voltage is applied to the anode. The gun electrodes focus the accelerated electrons through a hole in the

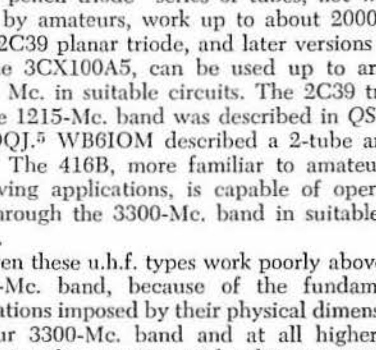


Fig. 10-11—Essential parts of the klystron tube, shown in simplified form to illustrate the theory of "bunching."

center of the anode, into a cylindrical beam which flows at constant velocity through the hollow drift tube and resonant cavities, to the collector.

R.f. drive applied to the input resonant cavity causes an r.f. voltage to exist between adjacent sections of the drift tube. (Spaces between these sections are called *interaction gaps.*) Electrons flowing past the input gap are affected by the r.f. voltage across it. When the voltage is positive (in the direction of electron flow) electrons in the gap are accelerated slightly. During the alternate half of the r.f. cycle they are retarded. Some of the electrons in the beam are thus moving faster and some are moving slower than the average rate—the beam is *velocity* modulated.

As the beam moves toward the output cavity, the fast electrons tend to catch up with the slower ones, by a process called *bunching.* When the bunches reach the output cavity gap they are well-formed, and they charge the output resonant circuit as sharp pulses of r.f. current. The cavity acts as a resonant coupling device, and the power is fed to the transmission

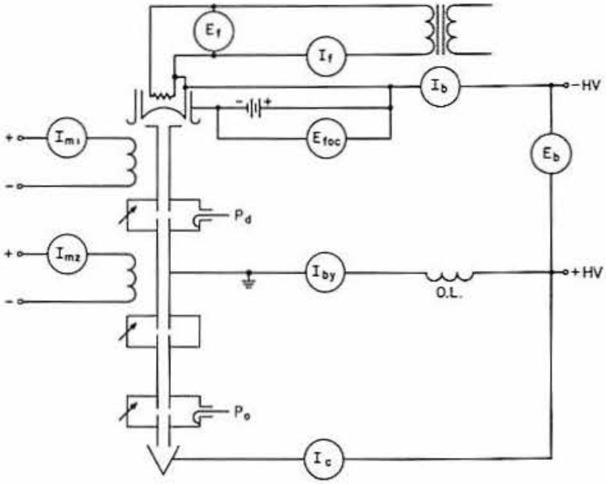


Fig. 10-12—Basic circuit of the klystron, identifying the various voltages and currents: E_f—filament voltage; I_f—filament current; E_b—beam voltage; I_b—beam current; I_{by}—body current; I_c—collector current; O.L.—overload relay coil; I_m—magnet current; P_d—driving power; P_o—output power.

line, and on to the antenna. The action is not unlike that in a conventional tetrode at lower frequencies, the essential difference being that in the klystron the electron bunches are not formed by a control grid, but rather by this velocity modulation of a continuous electron beam.

There is more to a klystron amplifier than this, and some of it is suggested in the simplified schematic of the klystron d.c. circuits, Fig. 10-12. Most power klystrons have 3 or more cavities, because multiple-cavity designs provide higher gain and efficiency than the simple two-cavity type described. Klystrons have been built with six or more cavities, capable of power gain in excess of 90 db. Amplifier klystrons have not been used extensively in amateur u.h.f. communication to date, their principal employment having been in 1296-Mc. moonbounce experiments;[8] where high power and extreme stability are mandatory.

The Klystron Oscillator

Most amateur work with klystrons has been with the simpler oscillator types, at power levels in the milliwatt range. Basic principles are similar to those described above, but the oscillator is of the two-cavity type. *Buncher* and *catcher* cavities are connected by a feedback loop to sustain oscillation. The catcher cavity is made resonant at the frequency of the velocity modulation of the electron beam, by changing the shape of the cavity physically and by adjustment of electrode voltages. The bunched beam current is rich in harmonics, but the high Q of the catcher cavity suppresses the unwanted harmonics and keeps the output wave form pure.

Practical Microwave Communication with Klystrons

Though klystrons and other microwave devices seem strange to the uninitiated, and admittedly they are costly when purchased new, microwave communication of a practical and highly satisfactory nature can be achieved with equipment that is surprisingly simple. Thanks to the surplus market and traditional ham ingenuity, it can be relatively inexpensive as well. The techniques employed vary in detail, but the principles are basically similar—and elementary.

The method is applicable to any microwave band for which klystrons are available. Two klystron oscillators are built and mounted in parabolic antennas. Each serves the dual role of transmitting oscillator and local oscillator for receiving. A crystal mixer is built into each unit, and some of the klystron oscillator output is diverted for injection to this mixer. The two stations are identical, except for the frequency of oscillation. The two oscillators are separated in frequency by an amount equal to the intermediate frequency to be used in reception.

Let us assume that the band is to be the one at 5650 Mc. One oscillator will be on, say, 5700 Mc. We will use f.m. broadcast receivers for our i.f. systems. This will require a mixer output frequency of about 100 Mc., so our other oscillator will be on 5800 Mc.

A klystron oscillator is very readily frequency modulated. Just a small audio voltage applied to the d.c. voltage on the klystron repellor element does the trick. This can be as simple a device as a microphone transformer connected in the repellor lead, with an ordinary carbon microphone the only other "speech equipment" needed. More advanced designs will include a speech amplifier, preferably with some provision for automatic level control, to keep the modulation at a constant level.

Keeping the transmitting and receiving functions separate may be handled in several ways. Separate antennas can be used for the transmitting oscillator and the receiving mixer. This was done in some of our earlier microwave work, but more recently it has been supplanted by the simpler expedient of "polarization duplexing." Here an open-ended cylindrical waveguide section has probes for transmitting and receiving, placed at angles of 90 degrees from each other. The "waveguide" may be nothing more than a beer can, as shown in later examples. The i.f. system can be an f.m. receiver, or even a simple superregenerative detector operating at the selected intermediate frequency.

This is inherently a duplex system, and one in which the actual operating frequencies are unimportant, so long as they are within an amateur band, and separated by the chosen intermediate frequency. There could hardly be a simpler communications system, yet "polaplexers" of this general type have been used in all our amateur bands above 3000 Mc., and for communication over some quite remarkable distances. Examples of this technique, too numerous for inclusion in this book, may be found in many issues of *QST*.[9]

OTHER MICROWAVE TUBES

Demand for microwave radar during World War II resulted in many devices being developed under very high priority. The klystron, the magnetron and the travelling-wave tube had existed in principle for some time but the wartime emergency brought them into mass production and use.

The magnetron and travelling-wave tube had little application to amateur communication, so they are dealt with only briefly in these pages. Essentially a magnetron is a thick-walled cylinder of copper, with a series of identical "keyholes" in the wall around the inner diameter.[10] Each keyhole represents a transmitter circuit, the hole itself providing the inductance and the slot, or base of the keyhole, the capacitance. As can be seen from the picture of the interior of a 10-centimeter magnetron, Fig. 10-13, these keyholes ring a central emitting cylinder or cathode.

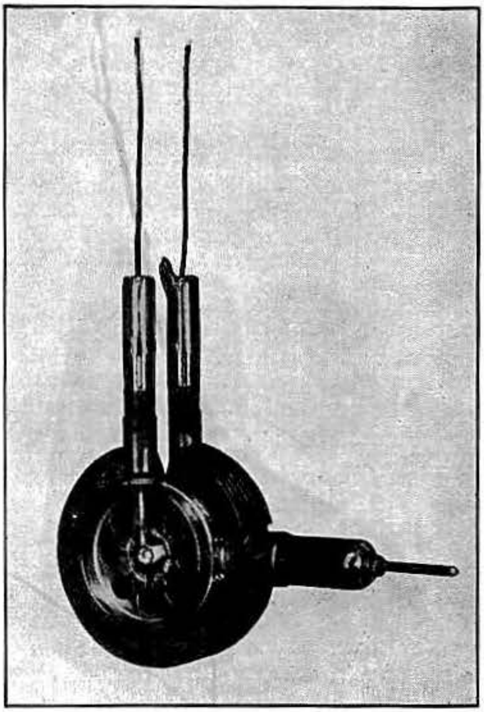


Fig. 10-13—Interior of a 10-centimeter magnetron, showing the 8 cavities surrounding the cathode. Complete tube has this structure operating in the field of a powerful magnet, not shown.

A magnetic field is applied axially, causing electrons to describe circular paths about the cathode when a high-voltage pulse is applied between anode and cathode. The critical velocity of the electron stream is reached when adjacent cavities represent positive and negative portions of the output wave. This is an oversimplified explanation, but it will suffice for our purposes, in view of the limited application of the magnetron to our kinds of communication. The main uses of magnetrons are for pulsed service, where very high peak voltages are applied for very short periods, at high repetition rates. Magnetron peak power of the order of a megawatt is common, but there has been limited use of the device for continuous-wave applications.

Gains as high as 50 db. over very wide bands in the microwave region are possible with the travelling-wave tube, shown schematically in Fig. 10-14. An electromagnetic wave travels down the helix, and an electron beam is shot through the helix from the electron gun at the left end, in the direction of the wave propagation. When the electron velocity is about the same as the wave velocity in the absence of electrons, turning on the electron beam causes a power gain for the wave propagation in the direction of the electron motion.

The input and output ports in Fig. 10-14 are coaxial lines to which the ends of the helix are coupled. The beam is focussed electrically at the gun end, and magnetically along the helix, by a series of opposing-polarity magnets stacked between ferrous pole pieces.

Outstanding features of the t.w.t. are great bandwidth and large power gain. Efficiency and power output are both rather low. The term "tube" is really a misnomer; the t.w.t. is actually a complete broadband r.f. amplifier in a vacuum envelope.[11]

Pulse Communication

In the table of amateur bands and modes in Chapter 2 pulse emission is shown as usable on amateur bands above 2300 Mc., with the exception of the 10,000-Mc. band. As the name implies, pulse transmission involves bursts of energy of very short duration, typically no more than a few microseconds. This is called the

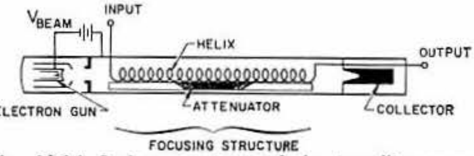


Fig. 10-14—Basic components of the travelling-wave tube. Device is actually a complete amplifier in a vacuum envelope.

pulse length. There may be up to several thousand pulses per second. This is the *repetition rate*, typically about 1000 per second.

Because power is on the equipment over such a small percentage of the time (low *duty cycle*), tubes and other components can stand many times their c.w. ratings, both voltage and current. An amateur station using pulse may thus have a *peak power* output of many kilowatts, so long as the average power does not exceed 1 kilowatt input.

A simple pulse system for the amateur is described later in this chapter. The possibility of employing surplus magnetrons for pulse communication with much higher power should not be overlooked. It is believed that reliable ranges at least equal to those covered on 144 Mc. could be achieved on any microwave band, if the full potential of pulse communication were exploited.

RECEPTION ABOVE 420 MC.

Receivers for the 420-Mc. band are usually crystal-controlled converters, much like those for lower frequencies. The circuits can be of the coil-and-capacitor variety, as in the converter described in Chapter 4, or they may be coaxial or strip-line arrangements, much like those of the 432-Mc. preamp and 1296-Mc. converter described later in this chapter. R.f. amplifiers, using both tubes and transistors, offer some improvement over the crystal mixer at 420 Mc., and practical examples are given in both chapters. Amplifiers other than the parametric variety have been little used in amateur work above the 420-Mc. band, but transistors show promise.

The Crystal Mixer

With the best crystal diodes and circuits, the noise figure of a crystal mixer is almost independent of frequency, up to around 10,000 Mc. The crystal mixer, having no gain (there is actually a slight conversion loss) does not override the noise of following stages, as does a good r.f. amplifier, so the design of stages following the mixer is important. Their noise figure is added to that of the crystal mixer.

The i.f. should be as low as practical, as low noise figure is more readily obtained below about 30 Mc. than at higher frequencies. We cannot go too low, however, or image response (including noise at the image frequency) becomes a factor in overall performance. A desirable i.f. for a 432-Mc. converter may be 14, 21 or 28 Mc., since communications receivers for these frequencies are universally available. Their noise figures are usually inadequate for this purpose, so a low-noise i.f. amplifier is often built into the converter assembly.

Occasionally 50 or even 144 Mc. will be used, where there is a converter or receiver for these frequencies available, and its noise figure is suitably low. The i.f. need not be tunable at this frequency. A crystal-controlled converter for 432 or 1296 Mc. can work into another at 50 or 144 Mc., the output of which may be 14 Mc., where the actual tuning will be done. An example of a 1296-Mc. converter with 144-Mc. output is shown later in this chapter.

Selecting Crystal Diodes

The 1N21-series crystal diodes have suffixes from A through F, each progressively better, and higher in price. Nothing below the 1N21C is desirable if noise figure is really important. It is capable of a noise figure of 8.3 db. The 1N21F (most costly) is capable of 6.0 db., which will be hard to match with tubes of any kind, above 500 Mc.

Injection Considerations

Stability is critical if narrow-band u.h.f. reception is undertaken. The converter crystal oscillator should run at the lowest practical input, and the crystal should be isolated from temperature variations due to component heating and cooling or air circulation. Putting the crystal inside the box is recommended. Injection should be as free of harmonic content as possible. Any injection makes mixer noise, and injection on other than the desired frequency is sure to degrade the noise figure of a crystal mixer. Finally, the injection system should be coupled loosely to the mixer, to prevent loss of signal energy through the injection circuits.

The latter two points are taken care of well in the 1296-Mc. converter described in this chapter, by the use of high-Q mixer and injection circuits, aperture-coupled. They are often neglected in amateur designs, with the result that mixers for 420 Mc. and up having noise figures in excess of 15 db. are not uncommon. When a good r.f. amplifier is used ahead of the mixer the amplifier establishes the noise figure of the system, but with a crystal mixer alone the above considerations must be handled with care if satisfactory reception is to be achieved.

THE VARACTOR DIODE

Perhaps no single development of recent years has had more impact on u.h.f. and microwave communication than the variable reactance device known as the *varactor diode*. This offshoot of the booming semiconductor industry made possible many improvements in receiver and transmitter design, some offering advantages never hoped for with vacuum tubes.

Notable among the varactor's contributions to u.h.f. progress were the parametric amplifier, proposed in theory earlier but made practical by the varactor, and the solid-state frequency multiplier for transmitting applications. Diode multipliers were also known for many years, of course, but the varactor works at power levels useful for transmitting. We will discuss these applications in some detail later, but first let us see how the varactor diode works. This section is mainly the work of Wayne Taft, W1WID.

Varactor Principles

Consider a small block of germanium that has been doped with impurities, so that one half is p-type material (contains free positive charges) and the other half n-type material (contains free negative charges). The result is a p-n junction diode. If a voltage of matching polarity is connected across it, free charges in the material will be repelled from the terminals, and move toward the junction boundary. This nets an exchange of charge, or forward conduction.

If the applied voltage is reversed the free charges are drawn away from the junction boundary, leaving a neutral region called the *depletion layer*. No exchange of charge is possible; hence the condition of high back resistance. We are interested in the latter condition. The depletion layer is a dielectric (no free charges) and the regions outside it are conductors (they have free charges). The two conductors thus act as the plates of a capacitor, whose plate spacing (capacitance) is dependent on the applied back-bias voltage.

In Fig. 10-15A we see the depletion layer for two conditions of back bias: low voltage, for close spacing and high capacitance; and high voltage, for wide spacing and low capacitance. A typical curve of capacitance *vs.* applied back bias is shown at B. This curve is good between the limits of zero bias and the reverse-breakdown voltage (V_B) of the diode. Capacitance is inversely proportional to the square root or cube root of the voltage, depending on how the semiconductor is doped. The most common type of varactor follows the square-root law, the result of an abrupt change in doping at the junction.

Practical Varactors

Up to this point we have considered the reverse-biased p-n junction as a lossless voltage-variable capacitor. Unfortunately the regions of the semiconductor containing free charges are not perfect conductors. Unavoidably they have a built-in fixed series resistance (R_s) usually between 0.1 and 10 ohms, which degrades the performance of varactor circuits from that obtainable if there were a completely lossless varactor. Further complications arise from mounting the device, as any practical package adds two important parasitic reactances. These are the internal series lead inductance and the shunt case capacitance.

The resultant varactor equivalent circuit is shown in Fig. 10-15C. The semiconductor chip itself is shown inside the broken line, with its variable capacitance, C_j, and series resistance, R_s. In series with it is the lead inductance, L_c, and in parallel with this combination is the case

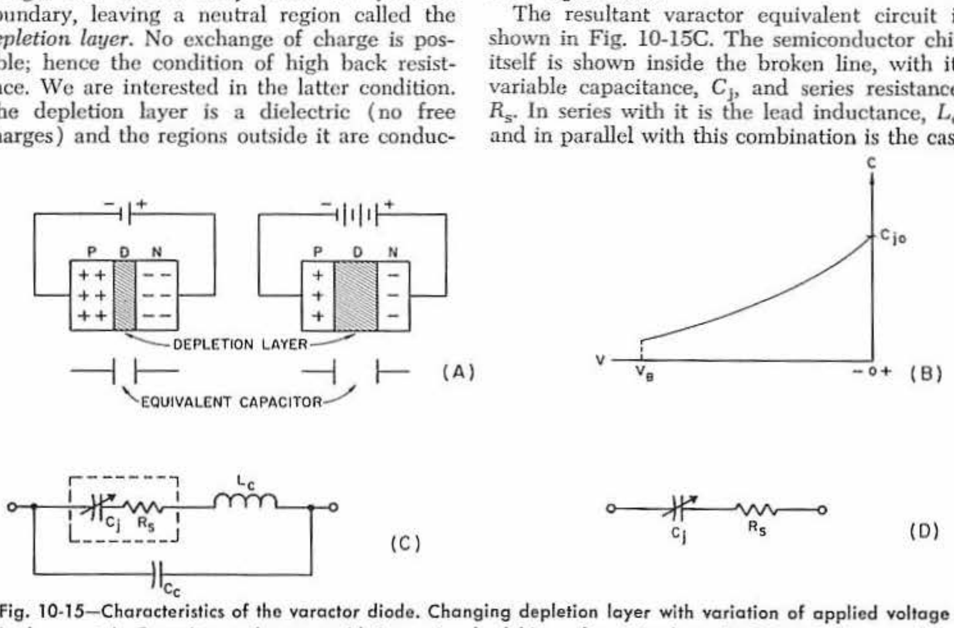


Fig. 10-15—Characteristics of the varactor diode. Changing depletion layer with variation of applied voltage is shown at A. Capacitance decrease with increasing back-bias voltages is shown in Curve B. The complete equivalent circuit of a varactor and its mount is given in C, and the practical effect of a varactor, within its design frequency range, at D.

capacitance, C_c. The parasitic reactances limit the maximum usable frequency of the varactor. Packages are available, however, for frequencies well into the higher microwave range. By choosing varactor packaging for the frequency range of interest, package reactances can be neglected. The simple equivalent circuit of 10-15D is then sufficient to describe the varactor.

Varactor units look very much like other diodes. The small glass-case version with pig-tail leads is useful only at frequencies below about 100 Mc., and at low power levels. A stud-mounted varactor of the type used in the frequency multipliers shown later in this chapter could be mistaken for a silicon rectifier diode, except for its price tag. It is useful up to 1500 Mc. or so, and at power levels up to 50 watts. Microwave packages commonly used for parametric applications include the 1N21 style and a related double-ended unit. Then there are tiny "pill" varactors for strip-line circuits, and various other mountings capable of working well up into the microwave region.

Varactor Terminology

In order to specify a varactor, certain measurable "parameters" are now in use:

C_{jVB} or $C_{j\ min}$—Junction capacitance at reverse breakdown.

$C_{j\ 0}$—Junction capacitance at zero bias.

C_{j-6}—Junction capacitance at some specified value of reverse bias, in this instance—6 volts.

R_s—Series resistance, sometimes called "spreading" resistance.

V_B—Reverse breakdown voltage.

θ—Thermal resistance in Degrees C per watt. Useful for power dissipation calculations.

Junction capacitance is usually measured with a bridge at some low frequency, on the order of one megacycle. The value of R_s is usually determined indirectly, by Q measurements at 500 Mc. or higher. In addition, two commonly-used terms involve combinations of the above:

Cutoff Frequency, f_c, at a specified value of bias, and hence C_j.

Normalization Power, P_{norm}, $= \dfrac{(V_B)^2}{R_s}$

These terms equate roughly with maximum usable frequency and plate dissipation as the latter would be used with vacuum tubes. In general, for equal cutoff frequencies, varactors with higher P_{norm} will handle higher power in multiplier service. In parametric amplifiers, varactors with higher P_{norm} require more pump power.

Applications and Availability

Varactors are useful for a variety of frequency-changing and amplifying applications, including electronic tuning, phase or frequency modulators in place of a reactance tube, parametric amplifiers and frequency multipliers. Some mention was made of the first two possibilities in our discussion of f.m. in earlier chapters of this book. We will look into the latter two below.

Once it was necessary to test various rectifier diodes in order to find one that would make a good varactor. Now many companies are producing diodes specifically for varactor purposes. These include such familiar names as Microwave Associates, Sylvania, Motorola, Amperex and others. Most varactors now are made from silicon, rather than germanium, for better high-temperature performance. Varactors made from gallium arsenide are also available. These have extremely high cut-off frequencies, but are somewhat more expensive than the silicon types.

Familiarity with the terminology outlined above will enable the would-be user to sort out units of greatest interest from the catalog listings. Presently-available varactors have characteristics as follows:

V_B—ranges from −6 volts for most parametric-amplifier diodes to −250 volts for the higher-power frequency-multiplier types.

C_{j-6}—from 0.1 pf. (microwave types) to over 100 pf. for v.h.f. frequency multipliers.

f_c—10 to 300 Gc. for silicon, and up to 800 Gc. for gallium arsenide.

R_s—0.1 to 10 ohms; usually higher-capacitance units have lower R_s.

Price is roughly proportional to cutoff frequency and P_{norm}.

THE PARAMETRIC AMPLIFIER

Lowest-noise devices for u.h.f. reception include the maser, the travelling-wave tube, and the parametric amplifier. There is little point in dwelling on the first two here. The maser must operate in a strong magnetic field. It requires certain gases, or exotic substances like rubies or garnets. Worst of all, it must be cooled to very low temperatures, requiring cryogenic techniques quite beyond the reach of the amateur worker. We have already discussed the t.w.t. briefly in this chapter. It, too, is an expensive device; its strong point, very wide-band amplification, is not required in amateur service.

This leaves the parametric amplifier, a development of some potential worth to the amateur, but one that has been little used and even less understood by most of us. Probably the best treatment of this admittedly involved subject, for the amateur reader, was a series of *QST* articles by W4AO and W4LTU[12] writen at the time that early development work on paramps was just getting underway. These articles are still required reading for anyone who would understand noise problems and the potential and partial solutions. Readers without engineering background or extensive electronics experi-

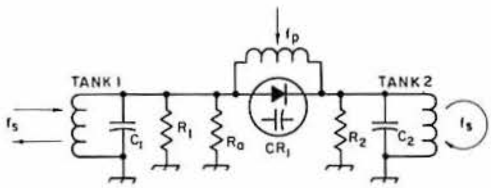


Fig. 10-16—Basic circuit of a parametric amplifier shows its similarity to a crystal mixer.

ence may not find them easy to assimilate, but they are about as simple as they could be without leaving the authors open to the charge of over-simplification. What follows is, to a large extent, a condensation of their excellent work.

Bateman and Bain made a valiant effort to stamp out the almost meaningless term, "parametric," in favor of "reactance amplifier," a name more indicative of the way the amplifier works, but "paramp" seems to have won out in the years since. There is a rather large family of parametric devices, and many mechanical and electrical analogies have been used to explain their operation. We will not go into these here, but they run all the way up from the children's swing, which is probably as apt (and as confusing) as any. Many systems are "pumped" in one way or another; we'll leave the analogies at this point.

The varactor is in effect a capacitor, the value of which changes with applied voltage. It can thus be used to modulate power from an external source, in relation to a signal voltage, and therefor amplify a signal applied to it. The parametric amplifier of most interest to amateurs is physically quite simple, being mainly a diode, fed at the signal frequency, and pumped at a higher frequency simultaneously. The basic circuit, Fig. 10-16, is quite similar to a conventional crystal mixer, and it may be used for either frequency conversion or straight-through amplification.

The signal frequency, f_s, is applied to *Tank 1*. In a frequency converter, *Tank 2* is tuned to the output frequency, f_o, which may be either higher (up-converter) or lower (down-converter) than the signal frequency. The pump tank, top, has only the job of providing an efficient means for exciting the diode capacitor (varactor). The terms "pump" and "pump frequency" are, in effect, merely new names for the more familiar local oscillator and its output frequency, in this case.

In an up-converter (output frequency higher than the pump frequency) a stable power gain equal to $\frac{f_o}{f_s}$ could be realized with ideal diodes and lossless circuits. If the output circuit is tuned to the pump frequency *minus* the signal frequency (though still an up-converter) the gain relationship is $-\frac{f_o}{f_s}$. The minus sign implies that regeneration is involved and, depending on conditions, very high gains can be achieved.

As a down-converter the output frequency is always lower than the signal frequency. Where the signal is higher than the pump the relationship $\frac{f_o}{f_s}$ remains, but since f_o is smaller than f_s, the device is an attenuator. When the signal frequency is below the pump frequency, and $\frac{f_o}{f_s}$ is still less than unity, the actual gain may be very high, because of regeneration, as in the up-converter.

In the regenerative arrangements the pump frequency is always the highest in the system, and is equal to the sum of the signal and output frequencies. In the regenerative conditions the signal in the input circuit is amplified by regenerative action, and the device may be used as an r.f. amplifier merely by taking the output from this point, instead of from the output circuit, *Tank 2*. The difference frequency must still appear in the output circuit, however. The terms "idler" and "idler frequency" have become standard names for the output tank and the energy therein. They have no purpose in our life, but they must exist.

For practical purposes, the approximate noise figure of the amplifier of Fig. 10-16 can be obtained from the formula:

$$F = 1 + \frac{R_a}{R_1} + \frac{f_s}{f_i}$$

where F is the noise figure, R_a is the shunt resistance across the input circuit represented by the antenna, R_1 is the shunt resistance represented by the losses directly associated with the tank circuit and diode, f_s is the signal frequency, and f_i the idler frequency.

The last two terms of the equation added together are a measure of the noise generated by the amplifier. Each should be kept small, so that their sum is a minimum. The second term can be kept small by coupling tightly to the antenna, so that R_a is much less than R_1. The third term may be kept small by using an idler frequency much higher than the signal frequency. This means, of course, a still higher pump frequency.

The way that the noise figure varies with pump frequency and various values of $\frac{R_a}{R_1}$ is shown in Fig. 10-17. The bottom curve, for $\frac{R_a}{R_1} = 0$, represents an idealized case in which R_1 is considered infinitely large. This curve illustrates the value of a high pump frequency. For example, if a pump frequency 5 times the signal frequency is used, the contribution from idler noise will be 0.2. The noise figure under idealized conditions would then be 1.2, or about 1 db. In any practical circuit, however, the contribution from $\frac{R_a}{R_1}$ will add to this. Thus,

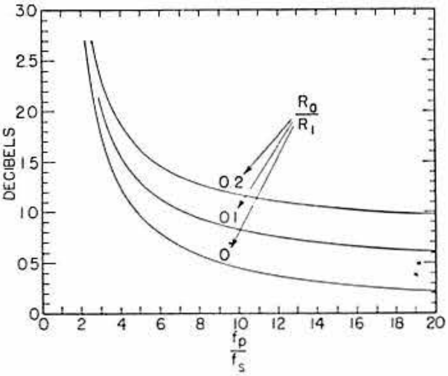


Fig. 10-17—Noise figure of the parametric amplifier of Fig. 10-16, as a function of frequency and antenna loading.

when you are straining for lowest possible noise figure it would be more practical to use a pump frequency in the range of 7 to 10 times the signal frequency. The contribution from idler noise will then be in the range of 0.11 to 0.17, leaving some room to maneuver in with respect to the contribution from $\frac{R_a}{R_1}$

From here on, analysis of the parametric amplifier is very involved, and will not be dealt with in detail in this text. Though the noise figure equation, even in the simplified form given above, gives us indications on how to keep noise figure to a minimum, it is by no means the whole story. Nothing has been said as to how much capacitance variation is required from the varactor and its pump, but it may be said that the following conditions are desirable in a practical device:

1. High idler and pump frequencies relative to signal frequency.
2. High tank-circuit Q.
3. High-Q semiconductor capacitor, or varactor, CR_1.
4. High available capacitance variation, ΔC, in the varactor.
5. Small values for C_1 and C_2.

Practical Considerations

For a given gain, the regenerative amplifier configuration (basic circuit, Fig. 10-16) is the least stable of the arrangements outlined above. Its noise performance, however, is quite good. Furthermore, it may be used directly ahead of an existing receiver or converter. Another big advantage is that instability in the pump does not affect the frequency stability of the output. Typically, 20 db. of fairly stable gain is available over a bandwidth of 100 to 200 kc. at 432 Mc. This is enough bandwidth for weak-signal communication as presently done on this band.

A practical paramp consists of a varactor properly coupled to the necessary tuned circuits and pump. In the v.h.f. range lumped circuits are useful, but at u.h.f. and higher, coaxial, strip-line or cavity resonators are necessary. In the microwave region these can be constructed from waveguide. Resonators should always be high-Q, to prevent losses and poor noise figure. Mechanical stability is extremely important in a regenerative setup, as small mechanical variations can completely upset the tuning.

Varactors designed for paramp service have low breakdown voltage, usually between –5 and –10 volts. Typical zero-bias capacitances are on the order of 0.2 to 5 pf. Cutoff frequencies range from 20 Gc. up, but in general the higher the cutoff frequency the better the noise figure. Typical diodes are the Microwave Associates MA-450 and 460 series, and the Sylvania D4075, D4140 and D4141 series.

Paramp Limitations

Someone may be wondering about the noise associated with the amplifier output load, and whether it is amplified along with the signal by the regenerative action of the circuit. It is, and the coupling problems involved can be solved only by the use of an esoteric device known as a *circulator*. The circulator has the unique property of permitting power to flow in one direction only, between certain pairs of terminals. By properly connecting a circulator in a receiving system using a paramp, the noise generated in the load can be made harmless by dissipating it in a resistive system.

Without a circulator, checking noise figure by means of a noise generator can lead the worker astray, in that *apparent* noise figures much lower than the actual are indicated. Another "fudge factor" in noise generator measurement with a paramp is the low *indicated* noise figure obtained when a pump frequency only twice that of the signal frequency is used. Such an arrangement is fundamentally limited to a minimum noise figure of 3 db., but noise generator measurements may indicate a noise figure of *zero* db. These factors were undoubtedly at the bottom of some early amateur enthusiasm for paramps.

All this is not to say that the parametric amplifier has no place in the amateur u.h.f. picture. It certainly does have, for those amateurs sufficiently skilled in receiver work to assess what is being accomplished with the many adjustments required. The paramp as generally used in the amateur field is a very tricky item. The pump frequency and power level should both preferably be adjustable, in the interest of precise adjustment, yet both should be reasonably stable, so that they will stay put when other adjustments are made. These two attributes are not readily combined in r.f. power sources at several thousand megacycles!

Most paramps built by amateurs have used klystrons for pump sources. A 432-Mc. paramp requires something around 4000 Mc. or higher, for best results. Since the power and frequency stability of a klystron oscillator are both relatively poor, adjustment of a paramp using one becomes something approaching black magic.

The pump frequency and the diode bias must be adjusted, and then the pump power increased, while fiddling with the other two items. All three react on each other. If the operator finally does get things peaked up for optimum results, a slight change in load impedance (such as may occur when the antenna is rotated and objects of differing reflecting properties appear in its pattern) will throw the adjustments off, and the work starts all over.

Measurement of the various "parameters," an over-worked word we'll use this once, since we're talking about parametric amplifiers, is all but impossible. Adjustment for optimum results is cut-and-try, to a degree probably not encountered in any other amateur electronic endeavor.

Results can be worth the trouble. Even without the circulator (and not many amateurs have access to one) it should be possible to develop noise figures around 3 db. at 432 Mc., at least 2 db. better than is likely with vacuum tubes or crystal mixers. It is not easy, nor very permanent, but you will have fun along the way! It is worth noting that great progress has been made in the development and production of low-noise transistors for u.h.f. applications. They pretty well take over the burden of low-noise reception at 432 Mc. and should do the same for higher frequencies eventually.

Until such times as they do, there is considerable to be gained from use of the paramp for 1000 Mc. and up. There is little practical value in a paramp for lower amateur bands, except for practice and experience. The principles are applicable at any frequency, and suitable pump sources for lower bands are readily obtained. The amateur who wants to learn more about paramp construction and adjustment can work with them at 50 or 144 Mc., where measurement of results is considerably easier than at u.h.f.

The Bateman-Bain series[12] describes practical paramp construction for 144 Mc. An effective paramp for 1296 Mc. was described in January, 1961, *QST*.[13] A modification of this for 432 Mc. appears in October of the same year.[14] A 432-Mc. paramp with crystal-controlled pump, the work of K2CBA and W1WID, was described in Edition 1 of this Manual.

FREQUENCY MULTIPLICATION WITH POWER VARACTORS

We are indebted to Henry H. Cross, W1OOP, for the first practical information on use of varactors for frequency multipliers in transmitters for 432 Mc. The following is mainly from his *QST* treatment of this subject.[15]

Power varactors now available to amateurs will give up to 15 watts output on 432 Mc. when driven with 30 watts on 216 Mc. They will do almost as well tripling from 144 Mc. The tripler described below will give a substantial signal on 432 when driven by nothing more than any of the popular a.m. transmitters such as the Communicator. No auxilliary power or audio is required.

The d.c. voltage-capacitance characteristics and the output voltage as a function of time, for sine-wave current input, are shown in Fig. 10-18. Once the diode draws conduction current, the theory gets more complicated, but harmonic output does not cease, so the complications can be ignored for small currents. If the

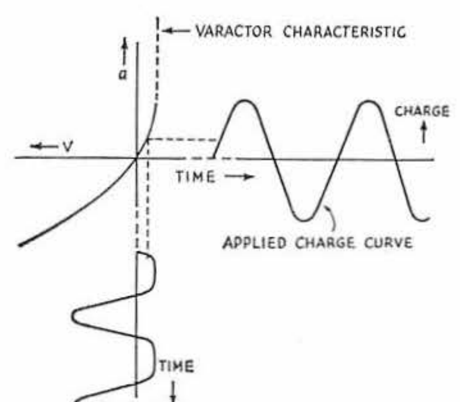


Fig. 10-18—D.c. voltage-capacitance characteristics, and output voltage as a function of time, of a varactor multiplier for sine-wave input current.

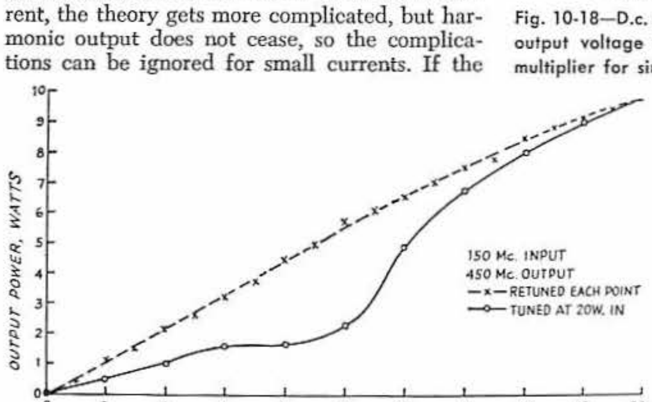


Fig. 10-19—Power output from a 450-Mc. tripler using a Type MA-4060A power varactor. The solid line shows the power available at various drive levels, when the tripler is tuned for maximum output with 20 watts drive. Uniform efficiency, upper curve, is possible if the system is retuned for each power level change.

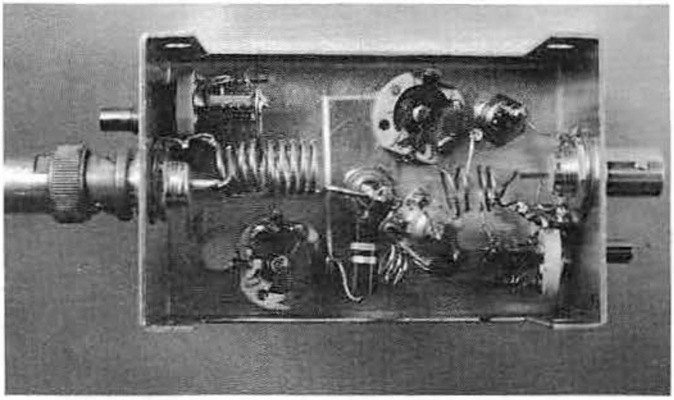


Fig. 10-20—Interior of a doubler stage using a power varactor. Driven with 20 watts on 216 Mc., it delivers 10 watts on 432, yet it requires no power supply or modulator.

multiplier is retuned each time the drive level is changed, an input-output curve similar to the upper curve of Fig. 10-19 is observed. For one tuning condition, the lower curve applies, and this is the case where a.m. is applied to the input. The function is not perfectly linear, but in on-the-air tests the 432-Mc. signal from a 2-meter phone rig and a varactor multiplier sounds quite satisfactory; better in fact than some 432-Mc. plate-modulated setups. Doubling from 216 Mc. to 432, with a unit like that in Fig. 10-20, the varactor does even better.

The circuit of the varactor doubler is given in Fig. 10-21. This works well with any 220-Mc. transmitter of moderate power. To operate in the segment of the 420-Mc. band usually reserved for narrow-band work, 432 to 436 Mc., the 220-Mc. rig is tuned down to 216 Mc., by using a crystal at about 8 Mc. even, and retuning the various stages to the lower frequency.

The trap, L_5C_5, is mostly to simplify tuning; without it changing the output capacitor, C_3, would also change the tuning of the 216-Mc. circuit. The double-tuned input and output circuits help to establish that the measured output is on the desired frequency. The circuits are rather low-Q, however, and use of a coaxial filter in the line to the antenna is recommended, to prevent radiation on the driving frequency.

Tripling to 432

A 432-Mc. tripler built by W1EHF is shown in Fig. 10-22, and the circuit in Fig. 10-23. There is an intermediate resonant loop on 288 Mc., and two traps, one on the input frequency and one to isolate the 288 from the output tuning. The "idler" at 288 gives improved 432-Mc. output. Theory stipulates that such an idler is needed, and tripling is not very satisfactory without it. Performance is shown in Fig. 10-19.

The traps can be tuned up with a dip meter before they are wired in place, and they should not require readjustment in the multipliers. The best way to tune the rest of the system is with an output indicator of some sort, on a dummy load. A directional coupler at the input is convenient for setting up the input network, which should be adjusted for zero reflected power. Maximum drive to the multiplier should then be obtained by adjustment of the driver output circuit, not L_1C_1.

When the driver is to be modulated the final peaking of the multiplier should be done while modulation is applied. Whistle loudly, while tuning for best linearity. This setting will *not* be the same as that for most carrier output. With a tripler, 20 watts of drive on 144 will give 8 watts c.w. on 432, and about 2 watts carrier for a.m., when tuned for best linearity.

The varactor multiplier is becoming increasingly popular as a means of developing power on 432 Mc., with f.m. and c.w., where its full capabilities are realized, and for low-powered a.m. with a modulated driver. A doubler similar to the one shown in Fig. 10-20 is used with the

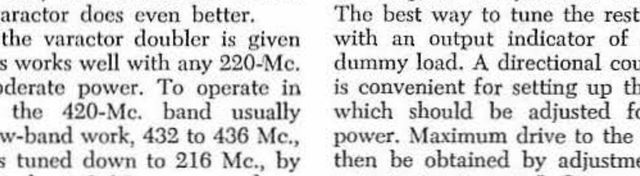


Fig. 10-21—Schematic diagram and parts information for the varactor doubler.

C_1, C_2—8.7-pf. miniature variable (Hammarlund MAC-10).

C_3, C_4—5-pf. miniature variable (Hammarlund MAC-5).

C_5—9-pf. subminiature variable (Johnson 189-503-4).

CR_1—Power varactor (Microwave Associates MA-4060A).

J_1, J_2—BNC coaxial fitting.

L_1—4 turns No. 20, ⅜-inch diam., ⅜ inch long. Tap at 1 turn.

L_2—5 turns No. 20, ⅜-inch diam., ⅜ inch long.

L_3—2½ turns No. 20, ⅜-inch diam., ¼ inch long.

L_4—2 turns No. 20, ⅜-inch diam., ¼ inch long. Tap at 1 turn.

L_5—3 turns No. 22, ¼-inch diam., ¼ inch long.

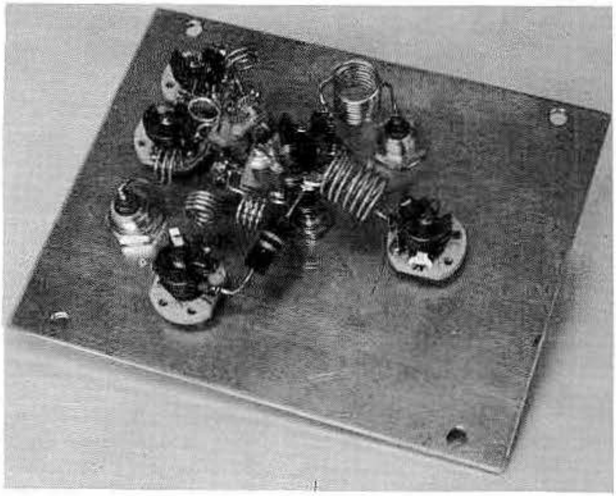


Fig. 10-22—Interior of the varactor tripler, for 432-Mc. output with 144-Mc. drive.

220-Mc. transmitter of Fig. 6-20, to drive a 4CX300A amplifier at W1HDQ. With less than 20 watts output on 216 Mc. enough 432-Mc. drive is developed to give up to 150 watts output from the amplifier, on f.m. or c.w. The multiplier is also used occasionally for low-power work, feeding the antenna directly. The strip-line filter of Fig. 12-12 is then used in the line to the 432-Mc. array, to prevent radiation on 216 Mc.

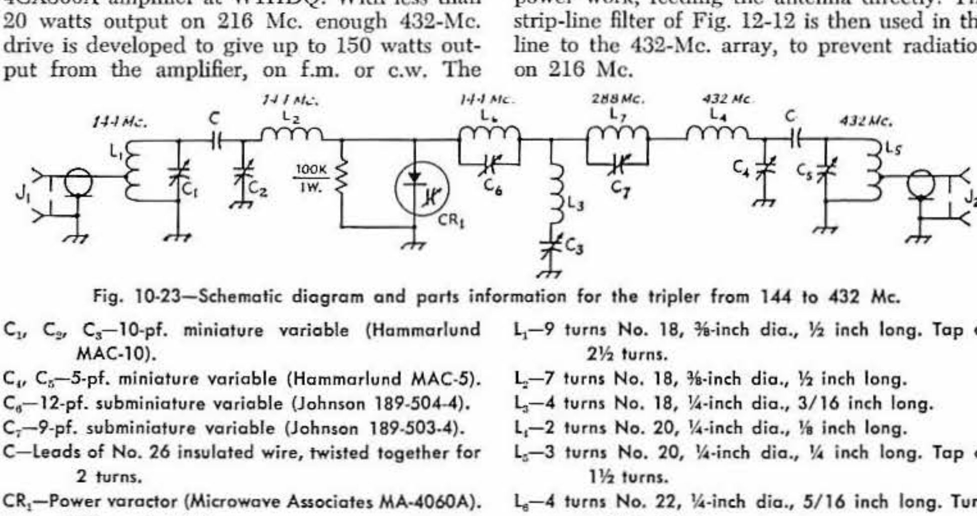


Fig. 10-23—Schematic diagram and parts information for the tripler from 144 to 432 Mc.

C_1, C_2, C_3—10-pf. miniature variable (Hammarlund MAC-10).

C_4, C_5—5-pf. miniature variable (Hammarlund MAC-5).

C_6—12-pf. subminiature variable (Johnson 189-504-4).

C_7—9-pf. subminiature variable (Johnson 189-503-4).

C—Leads of No. 26 insulated wire, twisted together for 2 turns.

CR_1—Power varactor (Microwave Associates MA-4060A).

J_1, J_2—BNC coaxial fitting.

L_1—9 turns No. 18, ⅜-inch dia., ½ inch long. Tap at 2½ turns.

L_2—7 turns No. 18, ⅜-inch dia., ½ inch long.

L_3—4 turns No. 18, ¼-inch dia., 3/16 inch long.

L_4—2 turns No. 20, ¼-inch dia., ⅛ inch long.

L_5—3 turns No. 20, ¼-inch dia., ¼ inch long. Tap at 1½ turns.

L_6—4 turns No. 22, ¼-inch dia., 5/16 inch long. Tune cold to 144 Mc.

L_7—1½ turns No. 22, ¼-inch dia. Tune cold to 288 Mc.

VARACTOR TRIPLER FOR 432 TO 1296 MC.

Happily varactor multipliers work almost as well on higher frequencies as in the 432-Mc. applications just described. A varactor tripler for 1296-Mc. output with 432-Mc. drive is shown in Fig. 10-24. It is the work of Wayne Taft, W1WID, who also contributed the basic information on varactors earlier in this chapter.

Except for the 432-Mc. circuits, coils and capacitors are out of the question for this applica-

Fig. 10-24—Varactor multiplier for 1296-Mc. output with 432-Mc. drive, designed and built by W1WID. Case is a 3⅛ by 1-inch brass box. Large screws at the left are the movable elements of capacitors C_4, C_5 and C_7.

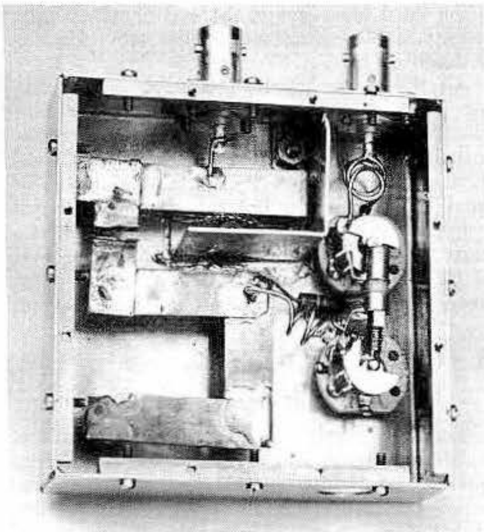


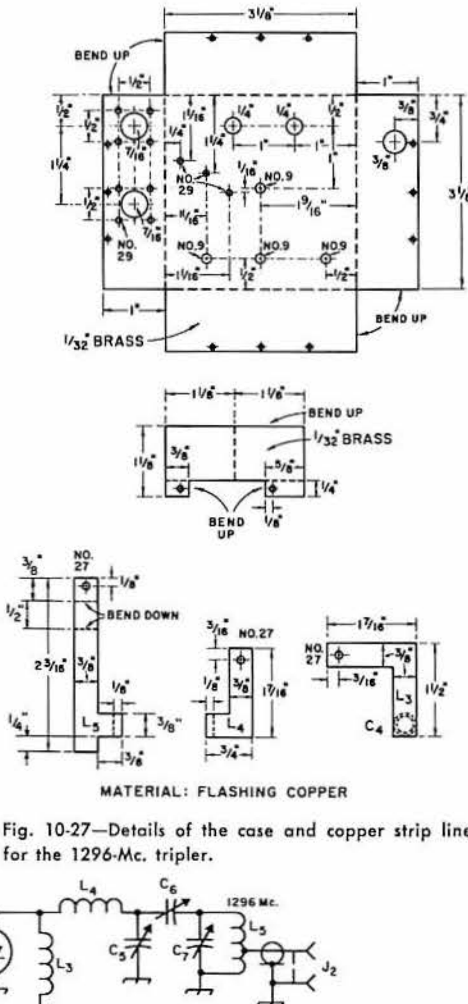
Fig. 10-25—Interior of the 1296-Mc. varactor tripler. Coils and variable capacitors are the 432-Mc circuits. Inductances for 1296 Mc. are copper strips. L-shaped shield of brass isolates input and output circuits.

tion. Strip lines are used in an ingenious and relatively simple manner. The circuit, Fig. 10-26, is almost identical to that of the W1OOP 432-Mc. tripler, but the circuits will require explaining, to the reader accustomed to the way such things look on lower frequencies.

The varactor is mounted in the center of a brass box 3⅛ inches square and 1 inch high. Adjacent to the BNC input fitting near one corner of the box is the 432-Mc. input circuit, L_1C_1. A small piston-type trimmer, C_2, couples energy to C_3 and L_2. The latter may be seen connected to the varactor at the center, though the varactor itself is out of sight under the strip-line circuits for 1296 Mc.

The line circuits are cut from flashing copper ⅜ inch wide. In the model shown they are made of separate strips soldered together, but they could be cut as shown in Fig. 10-27. L_3 is an "L" in shape as well as in function. The ⅛-inch hole in one end fits over the varactor post. At the other end is C_4, which is merely a piece of ⅜-inch brass or copper tubing, soldered to the strip, and facing toward the top of the chassis. Running through the chassis is a No. 10 brass screw, which is the "rotor" of C_4. It runs into the cup formed at the end of L_3 by the brass tubing.

Construction of L_4 is somewhat similar, except that coupling capacitor C_6 is built into it. The capacitor C_5 is merely another No. 10 screw that runs down so that its end makes a small variable capacitance to ground at the right-an-



MATERIAL: FLASHING COPPER

Fig. 10-27—Details of the case and copper strip lines for the 1296-Mc. tripler.

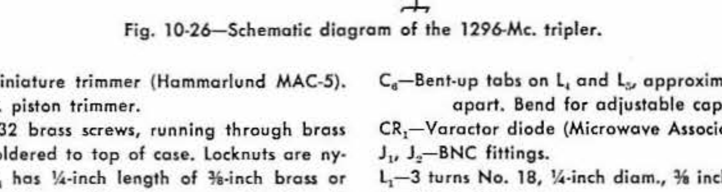


Fig. 10-26—Schematic diagram of the 1296-Mc. tripler.

C_1, C_3—5-pf, miniature trimmer (Hammarlund MAC-5).
C_2—0.5 to 5-pf. piston trimmer.
C_4, C_5, C_7—10-32 brass screws, running through brass nuts soldered to top of case. Locknuts are nylon. C_4 has ¼-inch length of ⅜-inch brass or copper tubing soldered to underside of L_3, to increase maximum capacitance.
C_6—Bent-up tabs on L_4 and L_5, approximately 3/32 inch apart. Bend for adjustable capacitance.
CR_1—Varactor diode (Microwave Associates MA4062D).
J_1, J_2—BNC fittings.
L_1—3 turns No. 18, ¼-inch diam., ⅜ inch long, c.t.
L_2—Like L_1, but 2 turns.
L_3, L_4, L_5—Copper strip lines. See text and Fig. 10-27.

gle turn in L_4. There is no brass cup at this point, as only a very small capacitance is required. Coupling between L_4 and L_5 (C_6 in the schematic) is made by bending up the ends of the short arms of L_4 and L_5. These ⅜-inch wide surfaces then face each other about 3/32 inch apart.

The output inductance, L_5, is the most complex piece. It is bent into U shape at one end to support itself at the same height from the chassis as the other inductances. The output tap for the BNC connector is made at a point 13/16 inch from this end. Capacitor C_7 is the third brass screw, the end of which provides variable capacitance in the same manner as described for C_5.

All this is an involved way of saying that tuned circuits really reach an elementary simplicity at frequencies this high. They are confusing only when we think of "coils" and "capacitors" in their 3-to-30-Mc. connotation. The small shield visible in the photograph is the full height of the box. It isolates the 432-Mc. circuits from the output, thereby keeping the level of the unwanted 432-Mc. energy in the output lower than it would be with an open layout.

TRANSMITTING CONVERTER FOR 50 TO 432 MC.

A varactor diode can be used as a combination mixer and multiplier, to produce a 432-Mc. signal from a 50-Mc. source. The transmitting converter of Figs. 10-28 to 10-30, originally described by W1ICJ in *QST* for March, 1966, uses 50-Mc. energy from a Heathkit HX-30 (s.s.b. exciter of about 6 watts output) and 10 to 15 watts of energy on 190.75 Mc. to produce about 3 watts of s.s.b. output on 432 Mc. With the HX-30's tuning range of 50 to 51 Mc., the resulting output frequency range is 431.5 to 432.5 Mc.

The 190.75-Mc. pump energy is doubled in the varactor to 381.5 Mc. The unwanted product, $381.5 - 50.5 = 331$-Mc., is also generated, and must be supported by the idler circuit, L_5, C_7 in Fig. 10-28. This frequency does not appear in the output. There is some 381.5-Mc energy in the output, but this is removed easily with a coaxial or strip-line filter.

Circuit Description

The input parallel-tuned circuits, L_3C_3 for the pump frequency, and L_1C_1 for the signal frequency, are lightly coupled to the varactor via pi networks. Although this appears unsatisfactory at first glance, it will be noted that the reactance of the 50-Mc. pi-network inductor, L_2, at the pump frequency is so high as to constitute an open circuit. Conversely, the pump inductor, L_4, presents essentially no reactance at the signal frequency but the capacitor to ground, C_4, does. Therefore essentially no loading of the signal frequency occurs.

The output has series-tuned idler circuits for 381.5 Mc. (pump frequency $\times$ 2) and 331 Mc. (pump $\times$ 2 minus signal), and two resonant circuits for 432 Mc. (pump $\times$ 2 plus signal). Output is taken from a tap on the output tuned circuit.

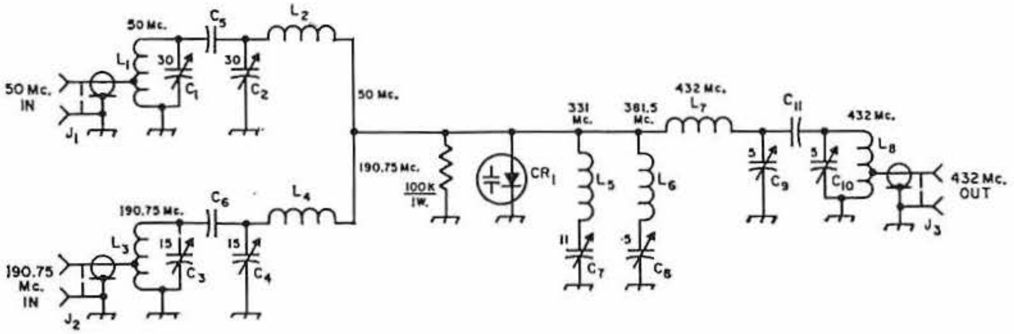


Fig. 10-28—Circuit of the parametric converter. Resistances are in ohms (K = 1000); capacitances are in pf. (μμf.)

C_1, C_2—30-pf. variable (Johnson 160-130).
C_3, C_4—15-pf. variable (Johnson 160-107).
C_7—11-pf. variable (Johnson 189).
C_8, C_9, C_{10}—5-pf. variable (Johnson 189).
C_5, C_6—Gimmick, 3 turns No. 18 solid plastic-covered hookup wire twisted together; ½-inch length for 190 Mc.; ¾-inch length for 50 Mc. (Johnson trimmers may be used as shown in photos).
C_{11}—Gimmick, 2 pieces ⅛×7/16 copper ribbon over-lapped ⅜ inch, spaced 0.020 inch.
CR_1—Varactor diode (Amperex H4A/1N4885).
J_1, J_2—BNC female.
J_3—Type N female.
L_1—10 turns No. 20, ½-inch diam.; tap at 3 turns (B & W 3003).
L_2—10 turns No. 20, ⅝-inch diam. (B & W 3007).
L_3—3½ t. No. 18, ½-in. diam.; tap at 1 t. (B & W 3003).
L_4—3 turns, same as L_1, without tap.
L_5—4 turns No. 18, ¼-inch diam., spaced wire diam.
L_6—3 turns No. 18, ¼-inch diam., spaced 2 times wire diam.
L_7—2 turns No 18, ¼-inch diam., spaced 2 times wire diam.
L_8—2 turns No. 18, ¼-inch diam., spaced, tap at ½ turn.

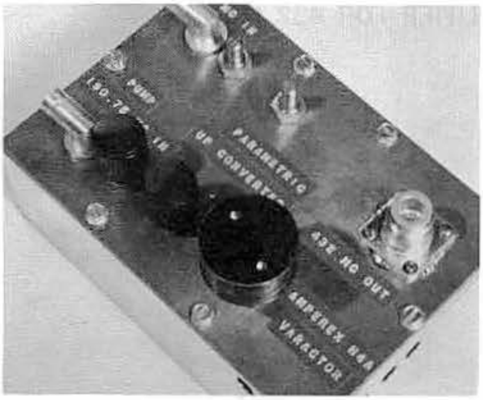


Fig. 10-29—Parametric up-converter for transferring a 50-Mc. s.s.b. signal to 432 Mc. The large black object, a heat-dissipating cap for the varactor, taken from a 2C39A tube, is not needed if an all-metal chassis is used.

As may be seen in the photographs, the unit is built in an inverted $4 \times 6 \times 2$ inch chassis. The mounting plate is 1/16-inch-thick double-sided printed-circuit board (0.040 copper may be used if desired.) All components are mounted on this plate. Wiring is done with at least No. 18 wire going point-to-point. Most leads are inherent in the components. The varactor heat sink shown is not necessary if a solid metal mounting plate is used.

Adjustment and Operation

Adjustment of the converter is a little tedious since there are eight interacting controls. First apply about 10 to 15 watts of pump power through an s.w.r. indicator, and adjust C_3 and C_4 for minimum s.w.r. A field-strength meter tuned to 381 Mc. placed nearby will serve to detect the doubling operation. Tune C_8 for maximum 381-Mc. signal. Go back and forth a few times between C_3, C_4 and C_8, adjusting for maximum 381-Mc. signal and minimum s.w.r. (they should coincide).

Now connect a load, preferably a wattmeter, to the output and apply about 1 watt (30-per cent scale on the HX-30) to the 50-Mc. input. Tune the wave-meter to 331 Mc. and adjust C_1, C_2 and C_7 for maximum as above. Now tune the wavemeter to 432 Mc. and adjust C_9 for maximum signal. Adjust C_{10} for maximum output to a wattmeter or other output indicator. Note that a high wavemeter indication at 432 Mc. indicates only circulating current in L_7C_9—*not output*. At this point it is well to go back and start again. Since all the adjustments interact to some extent you should go through at least three times. Do not be upset if the output indicator on the HX-30 goes up when connecting to the converter; this is normal. A 6-db. pad between the HX-30 and the converter gives better carrier suppression since you can use more audio (sideband power) while the carrier level output of the HX-30 remains essentially constant.

The unit exhibits good linearity when used to drive a 2C39 g.g. amplifier to about 12-15 watts output. There is some leakage of the 381.5-Mc. signal into the output; this is removed by a couple of tuned amplifiers or by a simple high-Q filter (see Chapter 12). With no filter the 381-Mc. signal is at least 20 db. down from the 432 output.

This same scheme can be used to convert from a 28-Mc. s.s.b. exciter, with appropriate changes in pump frequency and idler resonances. It is possible to triple from the pump source instead of doubling, with little change in efficiency. The overall performance is nearly the same except for slightly higher pump power requirements to make up for additional loss.

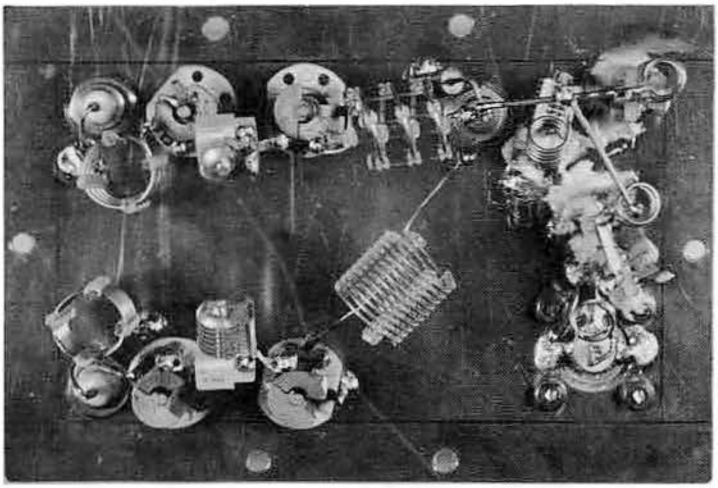


Fig. 10-30—Under side of the 50-to-432 converter. The 50-Mc. circuits are at the lower left, 190.75-Mc. circuits at the upper left, and 432-Mc. output circuits with idler tanks along the right side. The varactor diode is at the top just to right of center.

TRANSISTOR PREAMPLIFIER FOR 432 MC.

The transistor preamplifier of Fig. 10-31 has an appreciable edge in performance over any r.f. amplifier that could be made with vaccuum tubes, so no tube designs are given for this and higher frequencies in this Manual. The design follows one used in a 432-Mc. converter built by John Clark, K2AOP, and described by him in *QST* for December, 1965. The basic converter information is also in The 45th Edition of the *Radio Amateur's Handbook*.

Bipolar transistors are used. The model shown was tested with RCA 2N3478s and 2N2857s, but there are many other u.h.f. types available now that should give equal or better performance. Reversing voltage polarity permits use of p-n-p transistors. Side-by-side comparisons with the best vaccuum-tube r.f. amplifiers demonstrated that signals very close to the noise are definitely more readable when the transistor preamplifier is used. Its gain is adequate to override the noise of almost any crystal-diode or tube mixer, however poorly it may be working.

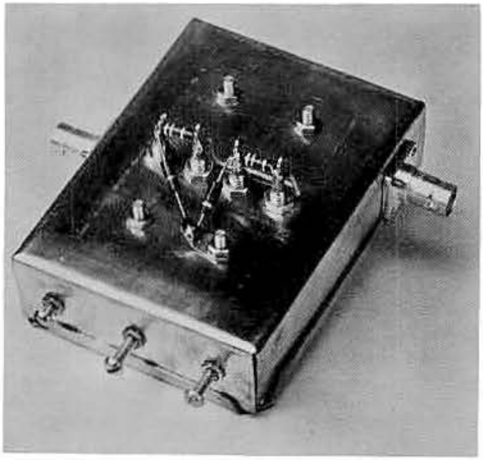


Fig. 10-31—A two-stage preamplifier for 432 Mc. The box is silver-plated brass, but flashing copper could be used with equally good results. Connections to the bases and collectors are brought out on feedthrough bypass capacitors, to permit changing the operating conditions.

Construction

Coil-and-capacitor circuits are usable with transistors at 432 Mc., but the trough-line arrangement shown is easy to make, and it is probably better in rejection of unwanted signals than if lower-Q coil circuits were used. Lines L_1, L_2 and L_3 are ¼-inch copper tubing, fitted tightly into holes in one end of the box, and soldered directly to the fixed elements of the ceramic trimmers at the other. No need to use expensive glass trimmers—the Centralab 829 series, 30 cents each, do the job nicely. The end of the tubing is countersunk slightly with a ¼-inch drill, to fit over the silvered end of the trimmer. This is better mechanically and electrically than using the flexible wire lead on the trimmer for making this connection.

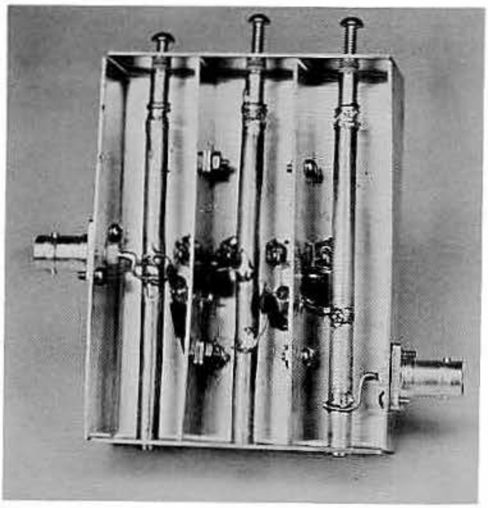


Fig. 10-32—Interior of the 432-Mc. amplifier, with the input circuit at the left. Partitions are held in place with spade lugs, and no heavy soldering is required.

Dimensions of the box are shown in Fig. 10-34. We used 1/32-inch brass, but flashing copper is good enough, and easier for the kitchen-table worker to handle. We silver-plated the box and lines, which made the completed amplifier pretty (for a while) but probably accomplished little else. Without silver plating, copper is better than brass electrically, though brass works beautifully with hand tools and is easily silver plated. See Chapter 13. The partitions are also brass, held in place with two spade lugs each.

The transistors are in the left and center compartments, about 1¾ inches up from the bottom, as seen in Fig. 10-32. They hang by their leads, a method that might not be desirable for a receiver to be shot into space, but entirely satisfactory for amateur use. The base leads go directly to feed-through capacitors, C_4 and C_6. The bias networks, R_1-R_2 and R_3-R_4, are connected externally.

The emitter leads are connected to the junctions of the blocking capacitors and 1000-ohm resistors, without support other than that afforded by these parts. The collector leads run through ¼-inch holes in the two partitions. As indicated in Fig. 10-33, the collector circuits are in the center and right-hand compartments. Col-

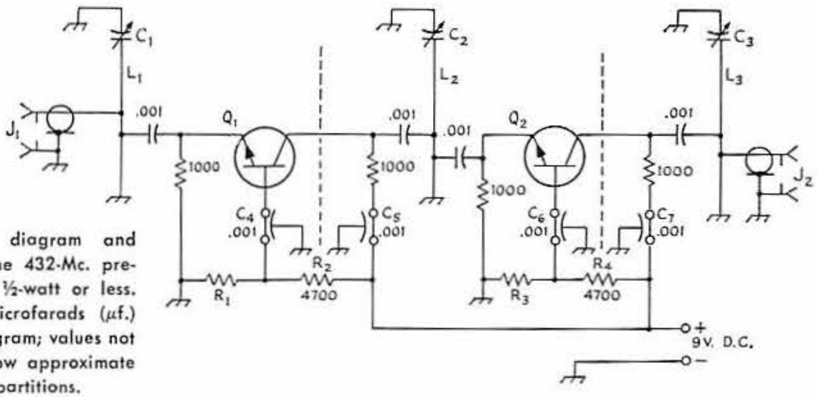


Fig. 10-33—Schematic diagram and parts information for the 432-Mc. preamplifier. Resistors are ½-watt or less. Capacitances are in microfarads (μf.) where shown on the diagram; values not critical. Broken lines show approximate positions of shield partitions.

C_1, C_2, C_3—1 to 7.5-pf. cylindrical trimmer (Centralab 829-7).

C_4, C_5, C_6, C_7—0.001-μf. feedthrough bypass; 500-pf. also usable (Centralab FT-500 or 1000).

J_1, J_2—Coaxial connector, BNC type.

L_1—¼-inch copper tubing 3½ inches long. Drill out end slightly to fit over capacitor body. Tap L_1 at 2 inches and 1¾ inches, L_2 at 1 inch and 2 inches, L_3 at 2 inches and ½ inch, all up from grounded end. See Fig. 10-32.

Q_1, Q_2—2N2857 or 2N3478. See text.

R_1—Adjust for maximum gain and best signal-to-noise ratio. Value in original was 2800 ohms with a 2N2857 for Q_1.

R_3—Adjust value for maximum gain, if necessary. 1000 ohms used with 2N2857 for Q_2, purposely lower than maximum-gain value. See text.

R_2, R_4—Labeled for text reference.

lector voltage is fed in through C_5 and C_7, from the top of the box.

Adjustment

Tuning of the preamplifier is very simple. The circuits are first peaked for maximum gain, and the input circuit is adjusted for best signal-to-noise ratio. No attempt was made to adjust the tap positions, as the amplifier seemed to work up to the specifications for the transistors, just as assembled. The value of R_1 in the bias network of the first stage is the principal critical factor, and it will vary with different types of transistors. Use a 5000-ohm control temporarily at this point, with a 10-ma. meter connected in the negative lead to monitor the total current drain. With the 2857s the optimum value for R_1 was about 2800 ohms, and the current to the first stage was about 2 ma. Higher current drain causes noise to rise faster than signal level, and much lower current costs some gain. About 200 ohms either way is enough to make a noticeable difference in noise figure or gain. With the 2N3478 a lower value may be better.

The value of R_3 can be juggled to suit require-

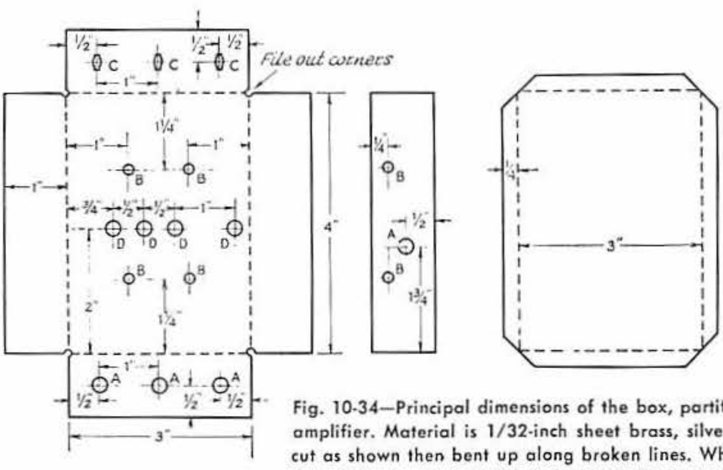


Fig. 10-34—Principal dimensions of the box, partitions and cover for the 432-Mc. amplifier. Material is 1/32-inch sheet brass, silver plated. Flat plates should be cut as shown then bent up along broken lines. Where precise bending cannot be done it is recommended that the cover be bent up to fit after the box is made. Hole sizes should be checked with available parts. Those shown are as follows: A—¼ inch, B—No. 28 drill, C—No. 28 drill, with 3/32 by 1/32 notches, D—3/16 inch. The three "A" holes in the bottom lip of the case should be a press fit for the tubing used for L_1, L_2 and L_3.

ments. It is not often necessary to run this stage at maximum gain, since noise figure is controlled mainly by the first stage. With about 1000 ohms at R_3 we had plenty of gain, with complete stability. More gain is available, with higher resistor values (more current drain) but instability may develop with some transistors. If there is a justification for the higher-priced units, greater stability under high-gain conditions is probably it. There should be no problem in getting adequate gain with the 2N3478s, and holding gain down by means of R_3 need not "cost you" in noise figure.

Total drain at 9 volts is about 4 ma. Higher or lower voltages may be used if R_1 and R_3 are adjusted in the manner outlined above, using the lowest current drain that gives optimum noise figure (R_1) and gain (R_3).

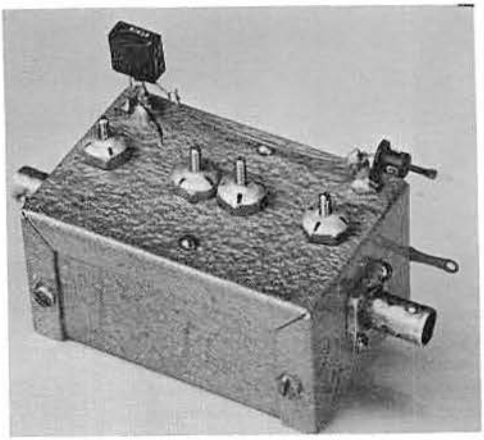


Fig. 10-35—The 1296-Mc. transistor preamplifier by K2UYH. Rectangular object at the upper left is a miniature control, R_1, for regulating bias. The diode at the upper right is in the plus 9-volt lead, for transistor insurance in case of inadvertent battery polarity reversal.

TRANSISTOR PREAMPLIFIER FOR 1296 MC.

Really effective use of amateur v.h.f. and u.h.f. bands has always hinged on the development of good receivers. This has been true from the earliest days, and today the threshold is the region just above 1000 Mc. Until the advent of u.h.f. transistors for r.f. amplifier service, the best receivers for the 1215-Mc. band used crystal mixers and low-noise amplifiers at the intermediate frequency, as in the 1296-Mc. converter described later in this chapter.

Transistors for r.f. amplifier service above about 500 Mc. are in limited production at this writing, and thus are fairly expensive. Channels are available whereby experimentally-inclined amateurs can obtain them at somewhat below current market prices.* The 1296-Mc. preamplifier of Fig. 10-35 was worked out experimentally by K2UYH and others, and described by him in *QST* for November, 1967. In the same issue was another preamplifier of more complex mechanical design, using the TIXM101 transistor, described by K4QIF. Either amplifier will make a worthwhile improvement in the performance of the best crystal-mixer converter.

Construction

The amplifier was built into a 3¼ by 2⅛ by 1⅝-inch Minibox. Disk ceramic capacitors are used to couple into and out of the preamp. Their leads are cut as short as possible to keep inductance down. The pi-network tuned circuits are composed of two 10-pf. glass trimmers (listed in some surplus flyers for about 30 cents) connected to each end of 1-inch copper straps. The width and shape of the copper straps, L_1 and L_2, are adjusted so that maximum gain comes within the tuning range of all capacitors. A ⅜-inch width gave good results with a majority of the K-2500s tried, but variations were noted from one transistor to the next. Notice that the strap is soldered across the tops of the capacitors, to keep inductance down. The position of the end of the strap on the capacitor terminal can be changed to give a slight range of inductance adjustment.

One need not be overly concerned with overheating of the transistor in soldering, if reasonable care is used. Ordinarily the leads will not conduct enough heat to do any damage. The small wire breaks off easily, however. The transistor is positioned between the capacitors C_2 and C_3, directly above a thin copper shield mounted across the center of the box. This extends the full height of the interior but has a notch cut at the point where the transistor will be. The emitter wire is soldered to this shield with the shortest possible lead.

Adjustment

For the initial tuneup, the bias should be set, by R_1, so that the transistor draws less than 1 ma. A signal source can then be connected, preferably through an attenuator, and the four trimmers adjusted for maximum gain. Be careful of oscillation and false tuning combinations. The proper capacitor settings will result in uniform

* The KMC 2500-series transistors used by K2UYH, and the improved 5200 series not available when the model shown was built, may be obtained at moderate prices from Samuel G. Nelson, W2MHK, Reaville Associates, RFD 1, Box 200, Flemington, N.J. 08822.

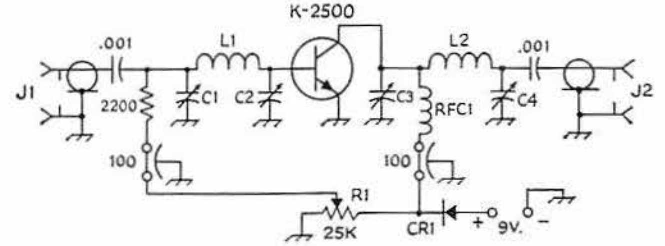


Fig. 10-36—Schematic diagram and parts information for the 1296-Mc. preamplifier. Decimal values of capacitance are in μf., others in pf.

C_1, C_2, C_3, C_4—1 to 10 pf. glass trimmer.
CR_1—Any power-supply type diode rectifier.
J_1, J_2—BNC fitting.
L_1, L_2—Thin copper strip, 1 by 3/8 inches. See text.
R_1—25,000-ohm trimpot.
RFC_1—5 turns No. 24 enamel spaced on 10,000-ohm or higher ½-watt resistor.

gain over a wide frequency range. The bias can then be adjusted for maximum gain, which will occur at a collector current somewhere between 2 and 4 ma. Do not permit the drain to exceed 10 ma., or the transistor will be damaged.

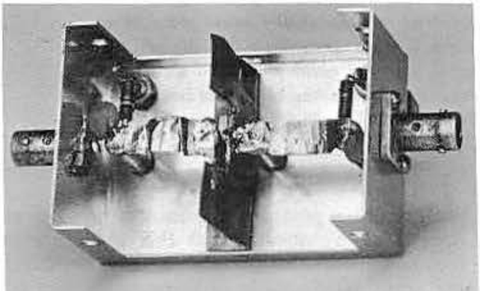


Fig. 10-37—The important item in this picture is practically invisible: the transistor is a tiny black dot at the center of the assembly. Tank circuits are u.h.f. versions of the pi network, for matching the low input and output impedances of the transistor.

Best amplifier noise figure occurs at bias settings that give less current than that for maximum gain. The optimum noise figure will occur at around 1 ma. The setting which will deliver the best overall noise figure may depend on the noise figure of the mixer, and on the nature of the stages following, if the mixer is the crystal-diode variety. In the light of the high noise figure of most 1296-Mc. converters in use today, it may be that best overall performance will be obtained with the amplifier adjusted for maximum gain, as the full 9 db. that this amplifier can deliver may be required to override the mixer and i.f. amplifier noise. The setting that gives best signal-to-noise ratio on weak signals is the one to use.

The diode, CR_1, prevents damage to the transsistor if the wrong polarity is applied. If you're sure that you'll always have the polarity right, CR_1 can be omitted, but it is cheap transistor insurance.

A CRYSTAL-CONTROLLED 1296-MC. CONVERTER

Crystal-controlled reception is a must if narrow-band work is attempted above 1000 Mc. The converter of Fig. 10-38 was built by W6-CGV, with help from K6UQH, K6ONM and W6VSV. It is not too much more of a project than a converter for any of the v.h.f. bands, yet its performance on 1296 Mc. is about all that can be achieved without going to parametric amplifiers.

The injection chain has only two 6J6s and a multiplier diode, using a 57.6-Mc. crystal to give injection on 1152 Mc. The output frequency is 144 Mc., chosen to avoid the need for building a low-noise i.f. amplifier stage as part of the converter. Most v.h.f. men already have good converters on 144 Mc., so the needed low-noise amplification at the intermediate frequency is taken care of easily in this way.

The front end, a simple crystal mixer designed as an integral part of a trough-line assembly, is seen from the bottom in Fig. 10-41 with the mixer input line at the top of the picture. The diode multiplier is in the bottom trough. Diode multipliers generate harmonics at all multiples of the driving frequency, so another trough is used to reject frequencies other than the desired 1152 Mc. This middle trough acts like a filter, and as a coupling circuit to the

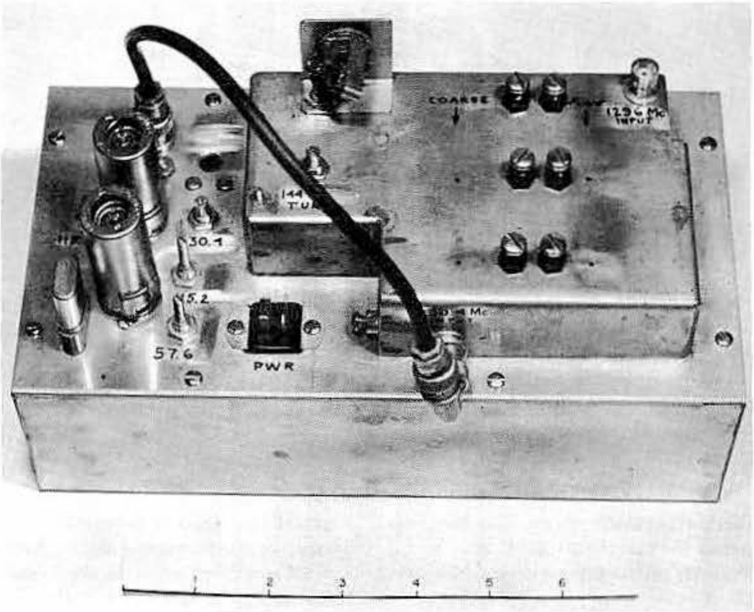


Fig. 10-38—The 1296-Mc. crystal-controlled converter is built on the cover plate of a chassis. The oscillator and multiplier stages at the left are coax-coupled to the crystal-diode multiplier, which is built into the penthouse atop the cover plate. The six screws with nylon nuts are for tuning the three half-wave tank circuits. The i.f. output frequency, 144 Mc., is taken off through a BNC fitting not visible in this picture.

mixer. Aperture coupling is used into this filter, and between it and the mixer. The mixer crystal is visible in the photograph, centered in the aperture between the mixer and filter troughs. The aperture coupling system does not load the Q of the mixer trough as much as a tapped mixer type, and improved rejection of both unwanted crystal harmonics and out-of-band signals results.

The i.f. tuned circuit, L_9 and C_7 in Fig. 10-45, is built into a separate compartment of the mixer assembly, at the right side of the photograph, to provide maximum shielding of the 144-Mc. circuits. Unless good shielding is used at this frequency, a few strong locals on 2 meters can cause a lot of trouble. Details of the mixer assembly metalwork are given in Fig. 10-39.

Oscillator and Multiplier Circuits

As may be seen from its circuit diagram, Fig. 10-40, the vacuum-tube portion of the multiplier chain is very simple. The first stage is an overtone oscillator on 57.6 Mc. The second half of the first 6J6 doubles to 115.2 Mc. This is link-coupled to the grids of a second 6J6, which is a push-push doubler to 230.4 Mc. The 230-Mc. energy is coax-coupled to the multiplier trough, where the diode multiplier output is picked off at the fifth harmonic, 1152 Mc. A fair amount of drive is required to make the diode quintuple effectively, and the 6J6 push-push doubler provided the most output of any tube tried. Substitutions at this point are not recommended, though almost any dual tube will serve satisfactorily in place of the first 6J6.

The diode multiplier is the heart of the converter. The secret lies in the impedance-matching LC network, and in the choice of the diode. Credit for the network and aperture mixing techniques, both essential for successful operation of the converter, rightfully belongs to K6-UQH. Several diodes, including the 1N72 and 1N82, were tried, the best producing a maximum of 120 microamperes of mixer crystal current. Diodes were then salvaged from plug-in u.h.f. converter strips for the widely-used Standard Coil TV tuner. Of these, the CBS 1N133 and the Raytheon CK710 worked equally well, yielding 300 to 500 μamp., which is more than enough. This permitted detuning the LC network to decrease the crystal current to the value that gave optimum noise figure for the diode used.

These plug-in converter strips are available for the asking, or at the worst at very low prices, at most TV service shops in areas where there is or has been u.h.f. television. Several of the diodes have since been used in other work with good results. Other diodes are undoubtedly suitable, one widely-used type being the Radio Receptor DR-303, also available at moderate cost.

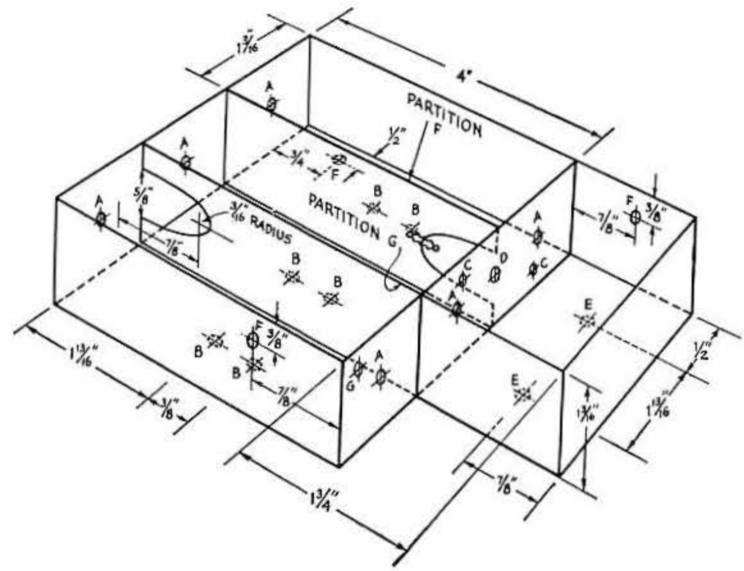


Fig. 10-39—Details of the mixer-multiplier trough assembly, as viewed from the bottom. The builder recommends 0.025- to 0.050-inch sheet brass, but with minor modifications in design thin materials such as flashing copper could be used. Holes are as follows: A—3/8-inch drill, on center line of each trough. B—No. 29 drill, tapped for 8-32 screw. C—No. 35 drill, tapped for 6-32 screw; to line up with No. 27 holes in capacitor parts. D—5/16-inch drill, on center line of partition E of Fig. 2. E—1/4-inch drill. F—3/8-inch drill, BNC fitting clearance. G—Trimmer hole, to suit type of trimmer used; location not critical. The notches at the ends of partitions F and G are coupling apertures.

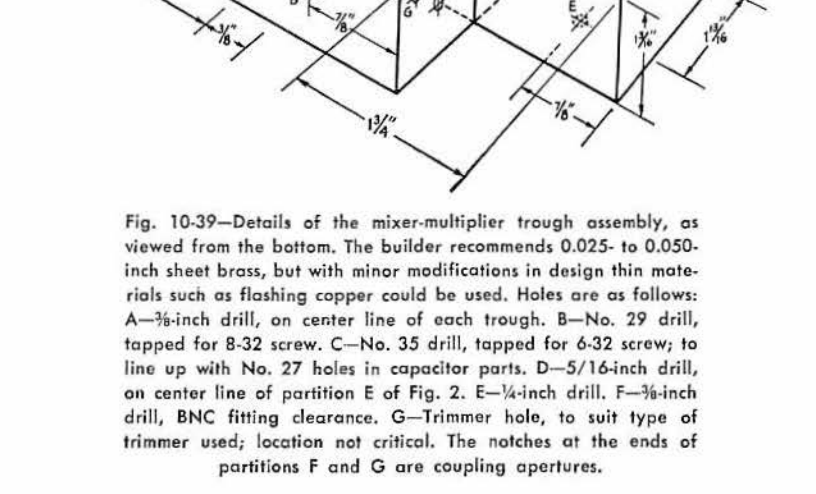


Fig. 10-40—Schematic diagram of the oscillator and multiplier section of the 1296-Mc. converter.

C_1—11-pf. butterfly variable (Johnson 11MB11 or 160-211).

C_2—9-pf. miniature variable (Johnson 9M11 or 160-104).

C_3—7—45-pf. ceramic trimmer.

L_1—10 turns No. 24 enamel on 3/8-inch iron-slug form.

L_2—6 turns No. 20 enamel like L_1.

L_3—2 turns No. 24 enamel around cold end of L_2.

L_4—Like L_3, but at centr of L_5. L_3, L_4 and link of one piece of wire.

L_5—8 turns No. 18, 3/8-inch diam., 5/8-inch long, c.t.

L_6—1 turn No. 18, 3/8-inch diam.

L_7—1 turn insulated hookup wire coupled to L_6.

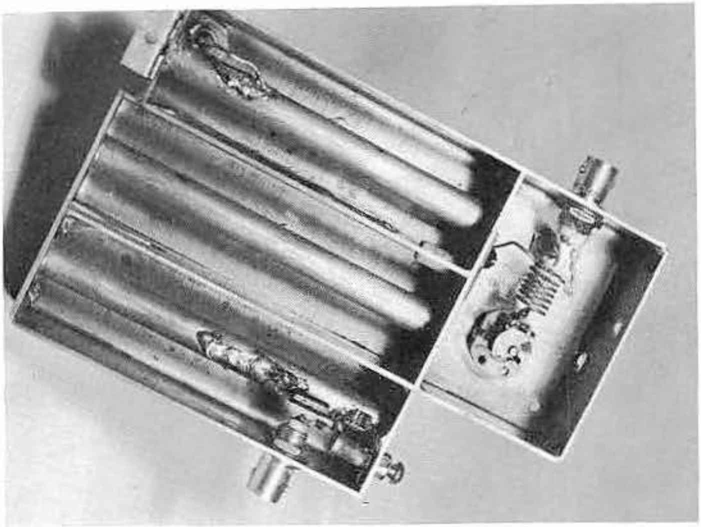


Fig. 10-42—Bottom view of the r.f. end of the 1296-Mc. converter. The multiplier circuit is the bottom trough. Here a diode delivers 1152-Mc. energy when driven at 230.4 Mc. by the oscillator-multiplier stages. The top trough is the 1296-Mc. mixer. Separating the two is an 1152-Mc. filter and coupling circuit. The mixer crystal may be seen in the aperture between the filter and mixer sections. The small compartment at the right houses the 144-Mc. output circuit.

Front-End Metal Work

The front-end assembly is constructed of sheet brass or copper, 0.025 to 0.050 inch in thickness. Brass is easy to work and makes a solid assembly. Fig. 10-38 shows the original model, which was made with the mixer signal-input cavity slightly shorter than the others. Later work proved this shortening to be unnecessary, so the drawing shows all troughs of equal length.

In making the trough, the sheet metal should be first cut to the dimensions and shape shown in Figs. 10-39 and 10-42. Drill all holes and tap where required. Before bending, cut along the line indicated in Fig. 10-42, then bend as shown. This is easy if you have access to a sheet-metal brake. If not, and you want a particularly neat job, you can have it done by a sheet-metal shop for a nominal fee. In doing the bending yourself, start with the lower lip of the right-hand portion of the assembly first. When the bending is completed, soldering of the joints at A, B, C, and D (Fig. 10-42) with intermediate or hard solder is recommended. Anything from 30/70 to Easy-Flo will do. Partition E is then soldered in place with the same type of solder. Partitions F and G may be soldered with 60/40 soft solder. The harder variety may be used for all work, but it is not recommended unless you are patient, and skilled with the torch.

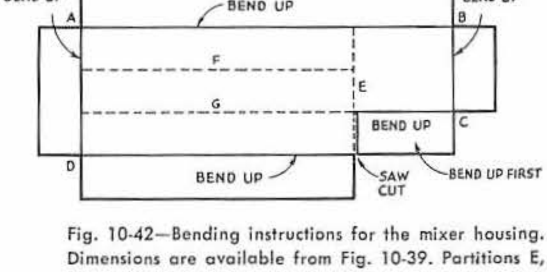


Fig. 10-42—Bending instructions for the mixer housing. Dimensions are available from Fig. 10-39. Partitions E, F and G, indicated by dashed lines, are soldered in place after the bending operation is completed. Note that the lower lip of the i.f. output portion at the right should be bent up first.

When the partitions have been soldered in place, insert the coarse-tuning screws, after first having run an 8-32 nylon nut up to the head of each screw. Now solder a large 8-32 brass nut to the end of each screw. Do this quickly and with a minimum of heat, and do not disturb the nylon nuts until the screws have cooled completely. Now insert the fine-tuning screws, each with nylon nuts, as before, but do not solder the brass nuts to these screw ends.

Now insert the ⅜-inch hollow brass lines in place (in 6 holes marked A, Fig. 10-39) and soft-solder. File the inside surface of the i.f. compartment, partition E, completely smooth, so that no sharp projection will puncture the insulation that is part of the u.h.f. bypass capacitor. Next, a contact pin removed from an octal socket is soldered to partition F, at the deepest point of the aperture, to make contact with the tip of the mixer diode. Solder a 2-inch length of No. 18 wire to the brass plate (see Fig. 10-43) for making connection to the i.f. output coil later. The combination crystal-retaining plate and u.h.f. bypass capacitor is shown in Fig. 10-43. This may be assembled with nylon screws as shown, but if these are not available, insulating shoulder washers and brass screws will do equally well.

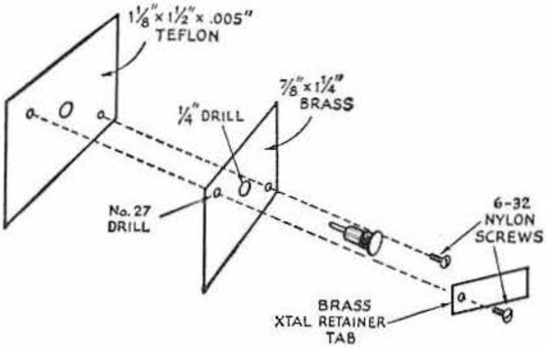


Fig. 10-43—Details of the mixer crystal mounting and u.h.f. bypass capacitor. These mount on the left edge of the i.f. output section, as seen in the bottom view. Locations of the mounting holes are not critical, so long as these and the mating holes in the mixer assembly line up. The center of hole D should line up with the center line of partition F.

Next, referring to Fig. 10-45, the feed-through capacitor, C_6, L bracket and closed-circuit jack for monitoring crystal mixer current are mounted as shown in the top-view photograph. The three BNC connectors are then mounted, along with the 7-turn i.f. coil and tuning capacitor, L_9 and C_7. The appropriate-sized hole is then carefully drilled in partition E at the end of the multiplier compartment to accommodate the small trimmer capacitor, C_4. In the unit pictured, the trimmer capacitor was padded with a small fixed capacitor to bring the tuning range of the trimmer to the proper point. The trimmer pictured is a 0.5-3-pf. unit salvaged from an old TV tuner. Use of the next larger size would eliminate need for padding. The small 4-turn coil, L_8, is soldered from the BNC connector to the trimmer, and the multiplier diode is soldered to the line approximately 1¼ inches from the inside wall of partition E. The optimum point will have to be determined later on, but this is a good place to start.

Connect the mixer output to the i.f. coil, using the 2-inch No. 18 lead previously soldered on the capacitor plate, 1¼ turns from the cold end of the i.f. coil. This connection will be adjusted later for maximum output. The i.f. output coupling loop, L_{10}, is installed with loose coupling to the cold end of the i.f. coil.

The 1296-Mc. antenna coupling loop is made of No. 18 bare wire and soldered to the BNC connector. Then it is run parallel to the ⅜-inch line and grounded to the trough wall. Several methods of input coupling were tried: the loop as described above, a direct tap on the line, and probe coupling. All worked equally well and all are relatively easy to adjust. The probe method is worthy of further mention since, of the three, it appeared to be the least critical to

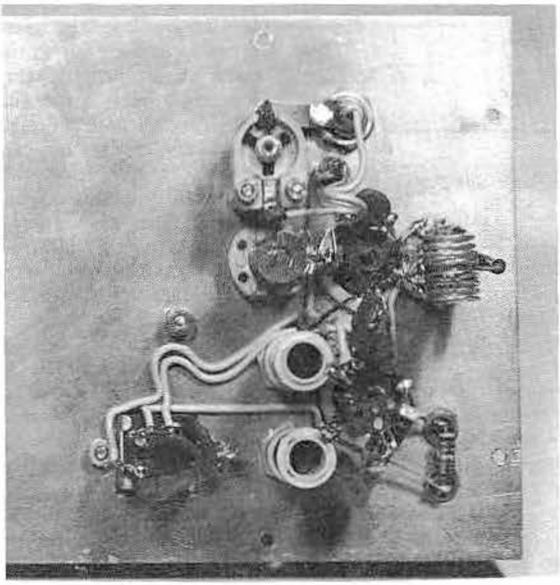


Fig. 10-44—Interior view of the oscillator and multiplier circuits of the converter. The two slug-tuned coils at the lower right are the oscillator and first-doubler plate circuits, L_1 and L_2. Above is the push-push doubler, with its 115.2-Mc. grid circuit at the right edge and the 230.4-Mc. plate and output-coupling circuits at the left and above the tube socket.

adjust. A $\frac{3}{16} \times$ 1-inch piece of brass was soldered into the center pin of the BNC connector and adjusted by moving it either closer to or farther from the line.

Multiplier Chain

The converter was constructed on the bottom plate of a $5 \times 9\frac{1}{2} \times 2\frac{1}{2}$-inch chassis. No specific mounting directions are given here since the techniques are quite straightforward. Fig. 10-40 shows the principal layout details. Subsequent models were constructed using a larger chassis. The 1296-Mc. trough assembly was mounted underneath the chassis, instead of on top as shown, to provide a little more shielding. In an effort to achieve greater stability, a longer multiplier chain was tried, to eliminate the third-overtone crystal. However, the unit constructed as shown is readily amenable to the application of more sophisticated techniques if they appear desirable later. Mounting the crystal underneath the chassis will help to insulate it from external temperature variations.

Adjustment and Operation

The power supply should deliver 250 volts d.c., 6.3 volts a.c. at 2.5 amp. and 150 volts regulated. An additional power plug may be added to run power to the 144-Mc. converter if desired. Design of the power-supply unit is left to the needs of the constructor.

When the trough assembly and multiplier chain have been constructed, apply power to the multiplier and tune up. With the voltage specified, the output at 230.4 Mc. should be capable of lighting a No. 47 pilot lamp to approximately half brilliance. If the output is much less than this, the preceding stages should be checked carefully, and adjusted until the output equals or exceeds the amount required.

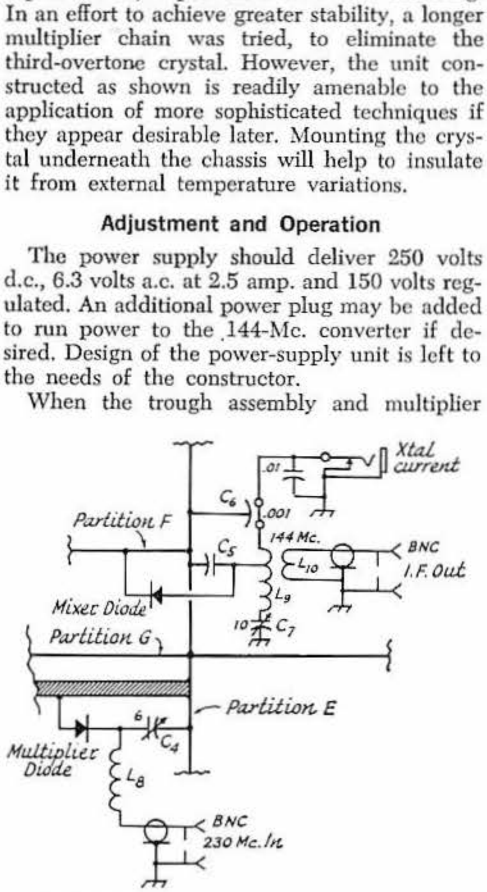


Fig. 10-45—Schematic diagram of the diode multiplier and i.f. output circuits of the 1296-Mc. converter. Decimal values of capacitance are in μf., others in pf.

C_4—6-pf. plunger-type trimmer.

C_5—U.h.f. bypass; see text and Fig. 10-43.

C_6—Feed-through capacitor, 0.0005 μf. or larger.

C_7—10-pf. miniature variable.

L_8—4 turns No. 26 enamel, closewound, 1/16-inch diameter.

L_9—7 turns No. 18, ¼-inch diameter, 7/16 inch long. Tap at 1¼ turns.

L_{10}—2 turns No. 24 insulated hookup wire inserted between turns of L_9. Twist leads to coax fitting.

The multiplier trough may be preset by turning the coarse-tuning screw until it bottoms on the trough line, then backing off approximately one turn. Set the fine-tuning capacitor to a depth of approximately ¼ inch in the trough. Set the coarse- and fine-tuning adjustments in the filter-mixer trough in the same manner.

The trimmer in the diode multiplier circuit should be set to approximately three-quarter capacity. Insert the mixer crystal (a 1N25 is preferable, but almost any of the 1N21, 1N23 series will do nicely), and plug a 0-100 microammeter into the mixer current jack. Couple the multiplier chain to the crystal multiplier with coax and BNC fittings. With power applied to the multiplier chain, a slight deflection should be noted on the meter. If no deflection is noted, check to make sure that the 1296-Mc. bypass capacitor, C_5, is not grounded. Caution: Remove the mixer crystal before measuring with an ohmmeter. If there is still no deflection, use a grid-dip oscillator tuned to 230 Mc. and lightly couple into the crystal-multiplier trough. Adjust C_2 and C_3 for maximum dip. A slight indication should now be seen on the microammeter. Adjust the coarse tuning on both the multiplier and filter troughs for maximum meter indication. Change the meter to a 0-1-ma. type and adjust the fine-tuning and trimmer capacitors for peak crystal mixer current. Adjust the diode multiplier tap on the trough line for maximum mixer current, being careful not to apply too much heat to the leads of the diode when soldering. A pair of long-nosed pliers will conduct most of the heat away if used to hold the diode pigtail during the soldering operation. When all adjustments have been completed, a reading somewhere between 200 and 500 μa. should be readily attainable, depending on the type of multiplier and mixer crystal used.

The injection frequency is 1152 Mc., the fifth harmonic of the multiplier chain. The trough will not tune to the fourth harmonic of the driver, but it will tune to the sixth, 1382.4 Mc. For this reason it is best to begin tuning adjustments from the maximum-capacity side.

If you have access to a stable 1296-Mc. signal generator, the rest is easy. A local 1296-Mc. amateur signal will serve nicely, or you may have to build a 1296-Mc. beacon. This is not too difficult. Use a 54-Mc. third-overtone crystal in a transistor oscillator circuit and feed the output to a diode multiplier trough similar to the one described here. The entire unit can be built in a small box about 2 by 3 by 4 inches, including the battery power supply.

Pretune the i.f. coil to 144 Mc. with a grid-dip oscillator. Connect the i.f. output to a good 144-Mc. converter and the input signal to the converter. Tune the signal trough and i.f. tuning capacitor for maximum signal. Adjust the tap on the i.f. coil for best match. This point will be ½ to 2 turns from the cold end of the coil, depending on the type of mixer crystal used. Carefully position the output pickup link to the point of maximum signal while returning the i.f. coil each time an adjustment is made. Next, adjust the input loop or probe for best *noise figure,* using whatever diode noise generator you may have. You will generally find this point lies in the direction of greater coupling from the position of maximum signal strength. When the input circuit has been adjusted for optimum noise figure, vary the crystal mixer current from 50 μa. to the maximum available. Make comparative noise-figure measurements for every 20-μa. increase in mixer current. You will probably find the best noise figure occurs between 150-200 μa. with very little change for values between 200 and 500. You are now in business with a 1296 converter.

500-WATT AMPLIFIER FOR 432 MC.

The best tuned circuit for an amplifier in the 420-Mc. band is a coaxial line. To build a good one requires some metal work, but the assembly described here should not be difficult for the advanced worker. Amplifiers of this type have been built and used by W1QWJ and W1RVW, with excellent results. They run up to 500 watts input on f.m. and c.w., and the amplifiers operate very much as they would on much lower frequencies.

Input circuit details are given for both 144 and 432 Mc, permitting the stage to be set up for tripling or straight-through operation. An inexpensive 4X150A running as a tripler will drive any of the 250-series tubes with ease.

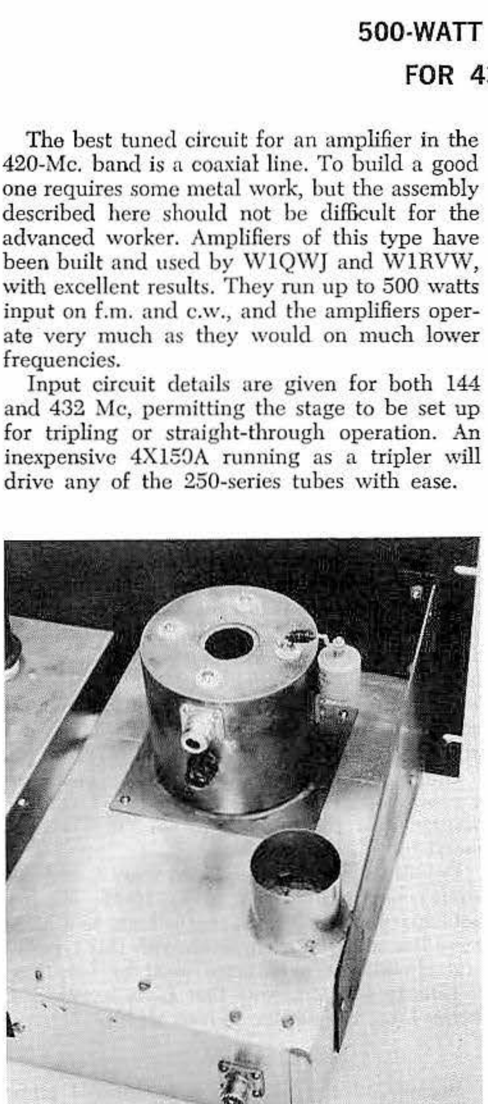


Fig. 10-46—Looking down at the coaxial plate circuit of the 500-watt u.h.f. amplifier. Air fed into the screened intake, lower edge of the picture, flows through the enclosed chassis below, up through the tube socket and out through the hole at the end of the plate line.

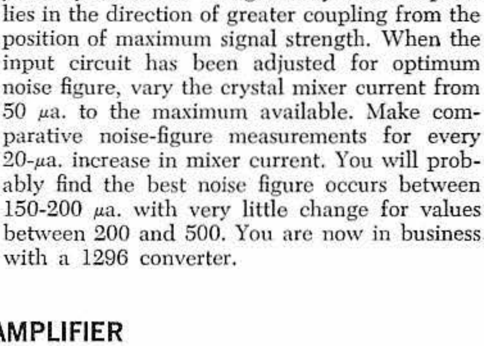


Fig. 10-47—Interior of the plate circuit assembly, showing the center conductor with its ring of finger stock, the output-coupling loop, left, and the disk-type tuning capacitor, right.

Construction

The basic design should be clear from the photographs, Figs. 10-46, 47 and 49. Structural details may be obtained from Fig. 10-50. The straight-through amplifier and the 144-Mc. grid circuit are shown schematically in Fig. 10-48. In the amplifier photographed, W1RVW used two separate 8 by 12-inch chassis, with their 8-inch surfaces fastened to a standard rack panel, 1 inch apart. They are held in firm alignment by an aluminum plate fastened at the back. One chassis carries the amplifier, the other a regulated screen supply.

The amplifier plate circuit is built in a 3¾-inch section of 4-inch copper tubing. This is mounted on a 5-inch square brass base plate. The top is a copper disk with a 1⅜-inch air hole at the center. Inside the cover is a teflon-insulated capacitor plate, soldered to the inner conductor of the plate circuit, L_3 in Fig. 10-48. The latter is 1½-inch copper tubing, 2^3⁄$_{16}$ inch long. A ring of finger stock extends ⅝ inch below the end of L_3, for making contact to the

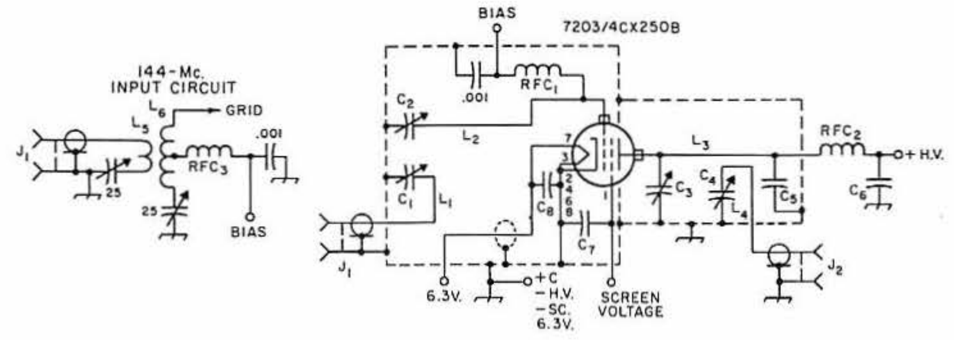


Fig. 10-48—Schematic diagram of the 432-Mc. amplifier, as set up for straight-through operation. An alternate 144-Mc. input circuit for tripling is shown at the left.

C_1, C_2, C_4—9-pf. miniature trimmer (Johnson 160-104 or 9M11).
C_3—Disk-type tuning capacitor, 1½-inch diam. brass.
C_5—Teflon-insulated high-voltage bypass. See text.
C_6—500-pf. 20-kv. TV-type capacitor.
C_7, C_8—Built into socket.
J_1, J_2—Coaxial fitting.
L_1—No. 12 wire loop, 6 inches overall. See Fig. 10-49 10-50.
L_2—1/16-inch brass, 1¼ by 3⅞ inches. See Fig. 10-49.
L_3—1½-inch copper tubing, with finger stock. See Fig. 10-50.
L_4—No. 16 wire loop, ¼ inch wide. Top is ¼ inch from C_5.
L_5—2 turns No. 16 enam., ½-inch diam., coupled to L_6.
L_6—4 turns No. 14 enam., ½-inch diam., 1 inch long, c.t.
RFC_1—8 turns No. 16 enam., ¼-inch i.d., ⅞ inch long.
RFC_2—8 turns No. 20 enam., on 1-watt 1-meg. resistor.
RFC_3—1.4-µh. r.f. choke.

7203/4CX250B anode. Eimac CF-300 Finger Stock, 31/32 inch wide, is used here.

The line is tuned by means of a brass-disk capacitor, C_3, details of which are shown in Fig. 10-50. The method of keeping tension on the lead-screw may be of interest, since this is often a problem with this type of tuning device. Two methods have been used by the builders. The amplifier shown has a piece of brass ½ inch square and ¾ inch long fastened to the outer wall. The screw passes through this, and the lower part of the block is slotted, up to the ¼-inch hole. A tension screw threaded into the block makes it possible to pull the sides together slightly, as required. The other tension system is shown in Fig. 10-50. Here a springy piece of metal is threaded onto the lead-screw, and then put under tension slightly by screws at either end.

The capacitor plate, C_5, at the top of the line is insulated from the cover with teflon sheet, the thickness of which is determined by the type of operation intended. If the amplifier is to be plate-modulated this sheet should be 1/32 inch. For c.w. or f.m. 0.01 inch is satisfactory. Four ceramic buttons insulate the screws that hold the capacitor together. Dimensions are not given for the holes required, as they will depend on the insulators available.

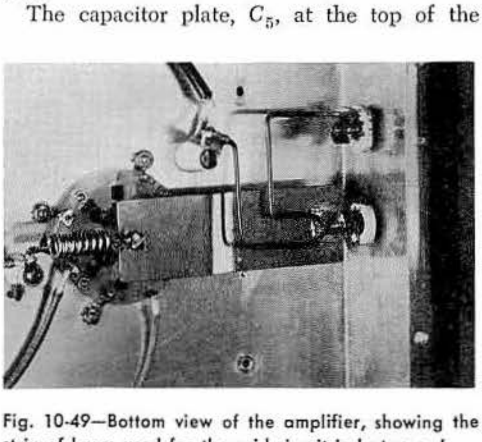


Fig. 10-49—Bottom view of the amplifier, showing the strip of brass used for the grid circuit inductance. L_2.

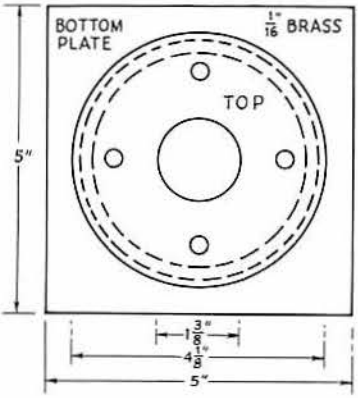
Note that the high voltage is on these screws when the amplifier is in operation. It is fed into one of them through a small r.f. choke, RFC_2, the outer end of which is supported on a TV-type 500-pf. high-voltage capacitor, C_6. The lower end of C_6 is supported on a brass angle bracket fastened to the side of the line assembly.

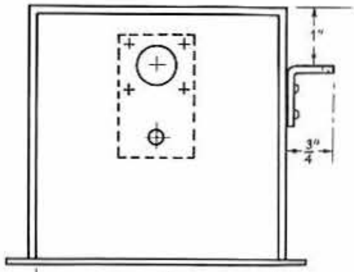
Output coupling from the line is by means of a small loop of wire, L_4, mounted in a vertical position near the top of the line. It is series-tuned by C_4, directly below it.

Details of the 432-Mc. grid circuit and its input coupling are given in Fig. 10-50. The input capacitance of these tubes is high, so a half-wave line must be used. Even with this type of grid circuit, the inductance must be very low to tune to 432 Mc. Note that L_1 is less than 4 inches long, despite its 1¼-inch width.

Operation

Because of the high-efficiency coaxial plate circuit, the amplifier operates almost as it would on lower frequencies. The manufacturer's ratings may be followed, using the maximum figures if desired. It is usually desirable to make provision for lowering the plate voltage in some way, however, as the difference between the maximum rating and something perhaps 25 to



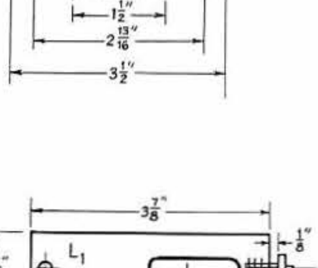


50 percent lower will make only a trifling difference in results, except where contact is being maintained under marginal conditions.

About the only variation from lower-frequency practice is the need for keeping the heater voltage low. The rated voltage for these tubes is 6.0, not 6.3, and at frequencies above 300 Mc. it should be reduced. At 432 Mc. the voltage should be 5.5. With higher voltages the back-bombardment that the cathode is subjected to raises the overall tube temperature and shortened tube life results. The drifting of operating conditions often observed in v.h.f. and u.h.f. amplifiers is likely to be traceable to excessive heater voltage.

Be sure to use plenty of air flow through the socket and tube anode. In the amplifier shown, air is fed into an opening in the top of the chassis. The bottom has a tight-fitting cover, so that the only air route open is through the socket and out through anode and L_3.

Adjustment of the position of the output coupling loop, L_4, with respect to the inner conductor of the line is fairly critical, if maximum efficiency is to be achieved. In one of the amplifiers the coupling loop, the coaxial fitting and the series capacitor were made into a single assembly on a curved plate of copper or brass. This could be removed at will, to permit adjustment of the shape and position of the coupling loop. It is fastened to the outside of the main cylinder with small brass screws, covering a rectangular hole in the cylinder cut for this purpose.



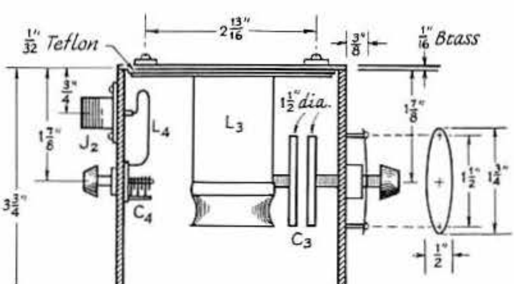
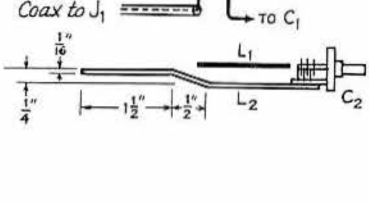


Fig. 10-50—Principal mechanical details of the 432-Mc. amplifier. The coaxial tank circuit is shown in cut-away form at the lower left, and in outline, center. The top view of the assembly and the capacitor plate for C_5 are the other views. .Details of the strip-line grid circuit are at the lower right.

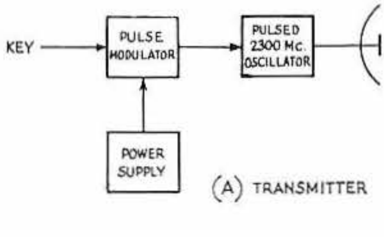


PULSE COMMUNICATION
ON 2300 MC.

Amateur microwave communication with simple oscillator-type transmitters, using amplitude and frequency modulation requires a wide frequency band, as such oscillators are highly unstable under modulation. The receiver must necessarily be broad-band, and therefore inefficient. Transmitter efficiency is poor, and the power output with tubes generally available is very low. While much good work has been done this way, and it is still useful in many instances, something better is needed if the full potential of the microwave region is to be realized.

The usual alternative, crystal control and narrow-band receiving techniques, effects a very great improvement in communications range, but it entails considerable effort and financial outlay. The pulse system described here, developed by John T. Zimmer, W2BVU, and Robert F. Guba, W1QMN,[16] represents a desirable compromise between these two extremes. Stability requirements are no greater than with the simple oscillator approach, yet the communications range approaches that of narrow-band c.w., with cost and complication far below the narrow-band method.

Pulse is a wide-band mode inherently, so it is permitted only in the microwave region, where amateur assignments are wide enough to accommodate it. The technique set forth in this condensation of the W2BVU-W1QMN *QST* series is applicable to the amateur bands from 2300 Mc. up, with the exception of the 10,000-Mc. band, where pulse is not permitted. What follows is merely the how-to-do-it treatment; the complete *QST* series is recommended to anyone interested in the advantages of pulse communication. An earlier discussion by Beers[17] is also highly worthwhile.



Equipment Requirements

A block diagram of a complete amateur pulse system is shown in Fig. 10-51. A code setup is shown; phone can be used with pulse, but keyed pulse is much simpler and more effective. The transmitter consists of a pulse-generating modulator and a simple oscillator for the r.f. Keying is in the pulse generator. The receiver has a conventional front end using a crystal mixer, with a local oscillator similar to that used for transmitting. The i.f. amplifier is broadband. Then come the principal elements wherein the pulse receiver differs from one for a.m. or f.m.: the threshold detector and the p.r.f. filter.

Overall efficiency is high. The oscillator is the sole r.f. component, in marked contrast to the string of multipliers required for microwave crystal control. It uses a 2C43 lighthouse tube, a low-cost surplus item. Though the average-power output is about 2 watts, using a pulse length of one microsecond and a pulse repetition frequency (p.r.f.) of 1000 per second, a peak-power output of 2 kilowatts is possible.

The modulator's three tubes draw about 10 ma. at 1300 volts, and 20 ma. at 300 volts, to produce 6 kw. peak input. The oscillator assembly can be installed at the antenna, to keep down feedline losses, and the modulator at any convenient spot below, with only a coaxial cable between them.

The receiver local oscillator can be similar to that used for transmitting, though this much power is not needed. A 2C40 oscillator similar to one described in *QST* by W2RMA[17] was used by W1QMN and W2BVU. Surplus oscillators can be found by some scrounging, and the possibility of using an APX-6 oscillator with a diode multiplier should not be overlooked. The i.f. amplifier, threshold detector, filter and audio system are described herewith.

The antennas should be parabolic reflectors, preferably 4 feet in diameter or larger. Using two stations as shown, the existing 2300-Mc. DX record of 170 miles was achieved with one station like that in Fig. 10-52 at sea level and the other only 600 feet higher. Typical v.h.f. propagation conditions prevailed at the time.

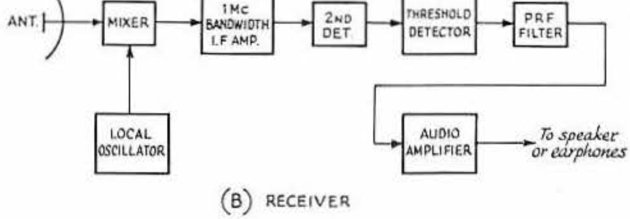


Fig. 10-51—Block diagram of a complete pulse communications system for the 2300-Mc. band. Communications range approximates that obtainable with narrow-band methods, with much simpler and less expensive equipment.

Fig. 10-52—W2BVU lines up his rooftop dish for a 2300-Mc. test beyond the visual horizon. Transmitter r.f. unit is in the weather-proof box at his feet. Reliable operating range approaches that on 144 Mc.

The Transmitter

In the r.f. unit, shown in Fig. 10-53 with its weather-proof cover removed, the oscillator is seen at the rear of the picture. A blower for cooling the oscillator, the 2C43 heater transformer and the pulse transformer are also visible. The only signal connection between the oscillator assembly and the pulse modulator, which may be placed at or near the operating position, is a coaxial cable to carry the high-voltage d.c. pulses from the modulator to the plate of the oscillator tube.

The oscillator is shown in cutaway form in Fig. 10-54, along with detail drawings of its component parts. A cylinder (C) mounted on the grid ring of the 2C43 lighthouse tube forms a coaxial line with the outermost cylinder (D) of the oscillator. This acts as an open-ended resonant tank circuit connected between the grid and cathode of the tube. Similarly, the same grid cylinder forms a resonant tank circuit between the grid and the plate. The feedback necessary for oscillation is obtained through the common opening at the ends of the coaxial tank circuits. Beyond this common opening, the plate line (A) is short-circuited to the outside cylinder, for r.f., by the cup-shaped choke assembly (B) mounted on the plate line. The outside cylindrical surface of this cup forms an open-ended coaxial line which, since it is exactly a quarter-wavelength long, appears as a very low impedance to r.f. inside the oscillator. Because of the way the grid and plate lines open into each other to produce feedback, the over-all circuit is called a re-entrant-cavity oscillator.

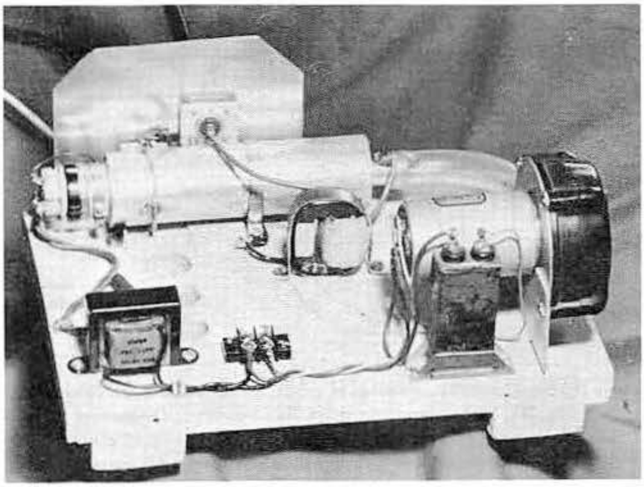


Fig. 10-53—The r.f. portion of the 2300-Mc. station is assembled on a wooden base that forms the bottom of a weatherproof box, for roof-top mounting. Output and modulator cables come in through fittings on the metal plate at the rear. Filament transformer in the foreground, pulse transformer between the blower and oscillator assembly. All holes in the box are screened to prevent entry of insects.

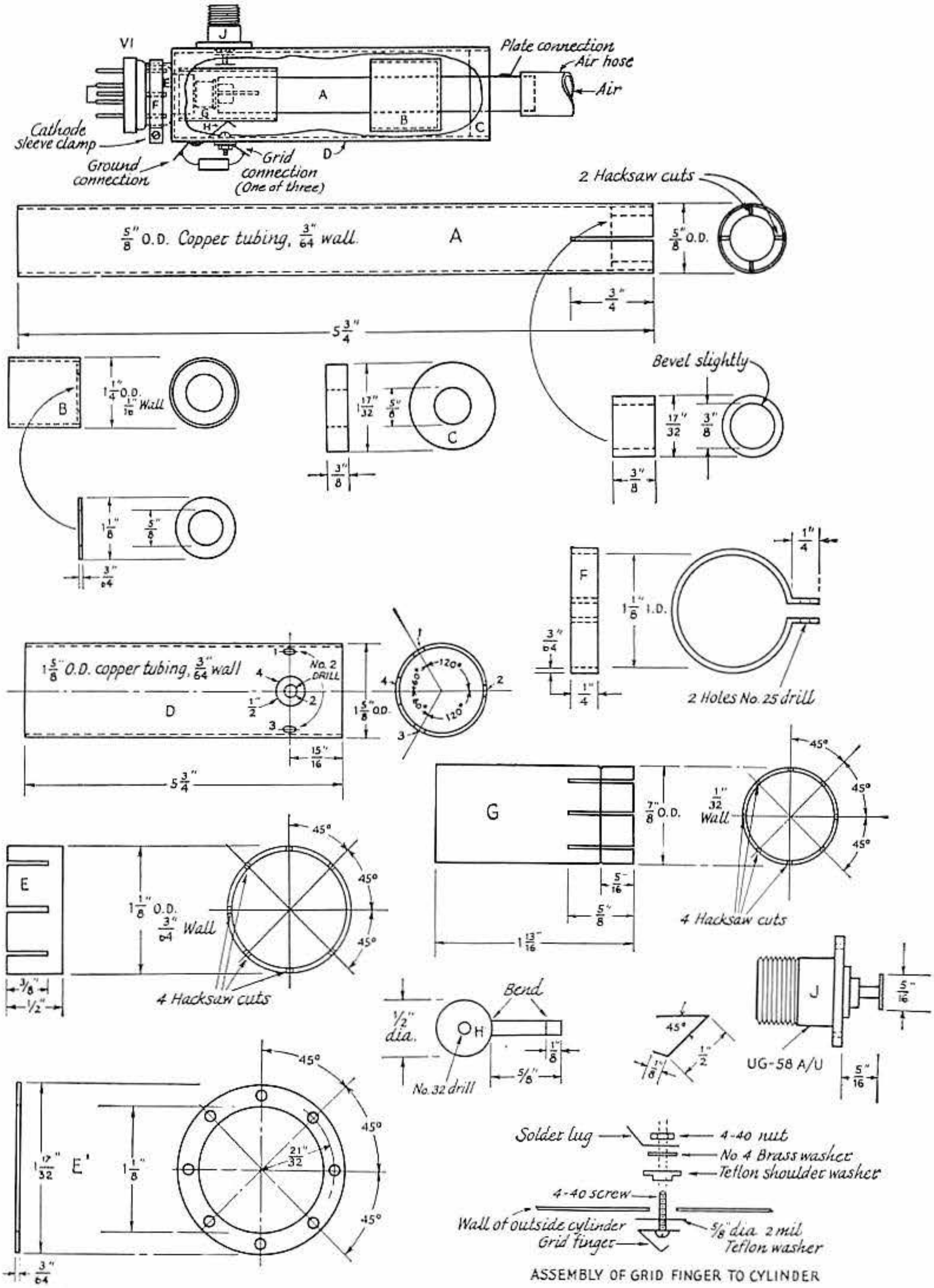


Fig. 10-54—Cut-away drawing of parts of the pulsed oscillator. A—plate line and contact; B—plate choke assembly, 1¼ inches long; C—plate line insulator; D—outside cylinder; E—cathode sleeve; E'—cathode end disk; F—cathode sleeve clamp; G—grid cylinder; H—grid contact finger (3 required); J—output probe (modified UG58A/U receptacle with thin disk attached). All parts except H and C are copper or brass. Cylinders are standard copper pipe diameters. H is thin beryllium copper.

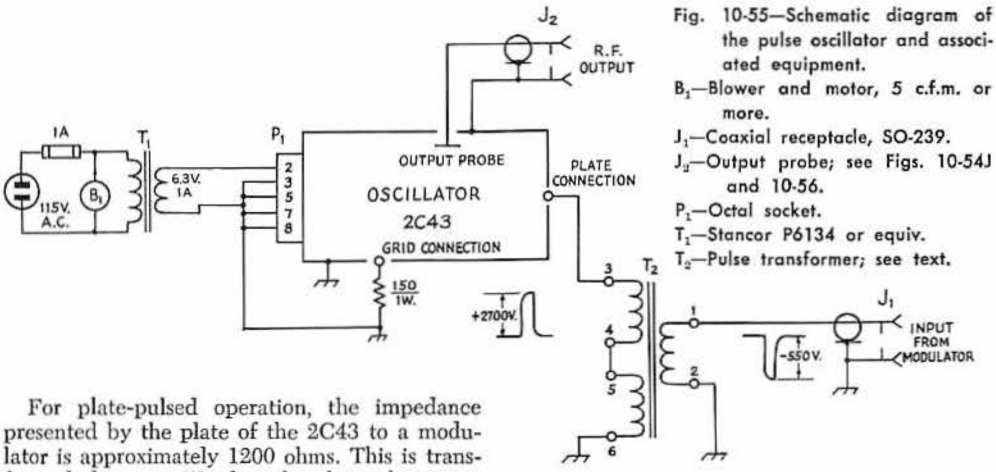


Fig. 10-55—Schematic diagram of the pulse oscillator and associated equipment.

B1—Blower and motor, 5 c.f.m. or more.

J1—Coaxial receptacle, SO-239.

J2—Output probe; see Figs. 10-54J and 10-56.

P1—Octal socket.

T1—Stancor P6134 or equiv.

T2—Pulse transformer; see text.

For plate-pulsed operation, the impedance presented by the plate of the 2C43 to a modulator is approximately 1200 ohms. This is transformed down to 50 ohms by the pulse transformer, T_2 in Fig. 10-56. Since the output pulse of the modulator is negative, the pulse transformer is also required to invert the polarity of the pulses. The voltage applied to the plate of the 2C43 when it is pulsed is between 2500 and 3000 volts. The peak plate current is two to three amperes, depending on the particular tube used.

Construction of a pulse transformer is described below, but it is likely that a suitable transformer can be found on the surplus market. Any pulse transformer rated for a few microseconds pulse length and a secondary voltage of at least 2500 volts, and having a primary-to-secondary voltage or turns ratio of roughly 1:5, should work satisfactorily.

Re-entrant Cavity Oscillator

The oscillator can be made from standard-size tubing available in plumbing-supply stores. Desirable tools are an electric drill, Greenlee punches, a tubing cutter, and a propane torch. All necessary dimensions are given in Fig. 10-54. The following comments indicate the proper sequence for assembling the complete cavity.

Outside Cylinder, D: Be sure to remove all burrs around holes 1, 2, and 3, as the d.c. grid connections are made here. Hole 4 can be made with a ½-inch punch if care is taken not to deform the tubing.

Plate Contact and Plate Line, A: The plate contact is made from ½-inch diameter brass rod. Wrap a 1/64-inch copper shim, ⅜-inch wide, around the rod and press fit the rod into one end of the plate line. Solder in place. Next, drill a ⅜-inch hole in the center of the brass rod and then cut two slots with a hacksaw. Remove all burrs and bevel the inside edge of the plate contact to facilitate insertion of the 2C43 plate terminal.

Plate Choke Assembly, B: The outside diameter of the choke ring should fit snugly within the choke cylinder, and the hole in the ring should fit snugly about the plate line. Insert the ring on one end of the choke cylinder, making sure it is flush with the end of the cylinder, and solder. Position the choke assembly as shown in Fig. 10-54, with the *inside* of the closed end of the choke 4 1/16-inch from the plate end of the line. Make sure that the choke cylinder is concentric with the plate line and then solder in place. To prevent previously-soldered connections from remelting, a dampened rag should be wrapped about these joints before the torch is applied. Position the plate-line insulator, C, on the plate line. The outside diameter of the insulator should make a snug but movable fit inside the outside cylinder.

Cathode Sleeve Assembly, E: The slots should be cut in the cathode sleeve before it is cut from the tubing stock. Before cutting out the cathode end disk, E′, mark the position of the eight air holes but do not drill these holes until after the disk is otherwise finished. Exact inside and outside dimensions of the disk should be tailored to provide a snug fit with the cathode sleeve cylinder and the outside cylinder. Fit the end disk inside the outside cylinder, making sure that the surface is flush with the end of the cylinder, and then solder in place. Fit the cathode sleeve inside the end disk, as shown, making sure the end of the sleeve is flush with the inside surface of the end disk. It is important that the cathode sleeve be concentric with the outside cylinder. Wrap the outside cylinder with a damp rag and solder the sleeve to the end disk. To secure a snug fit to the cathode surface of the 2C43, bend the ends of the slotted sleeve inward or file out the

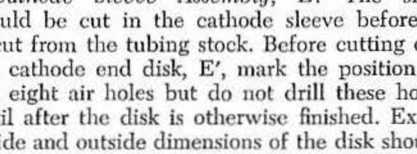


Fig. 10-56—Mounting of the output probe on the outer cylinder of the 2300-Mc. oscillator.

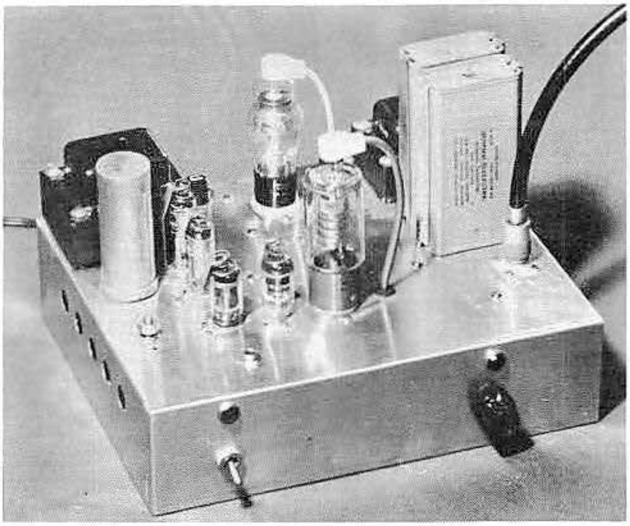


Fig. 10-57—Pulse modulator and power supply. The 3C45 hydrogen thyraton is the large tube near the center of the chassis. The pulse-forming network is located beneath the blank area of the chassis, between the thyraton and the output cable. The latter runs to the remote oscillator assembly, Fig. 10-53.

inside diameter of the sleeve. This will depend on tubing wall tolerances. The sleeve clamp is then made so as to grip the 2C43 firmly after it is inserted into the cavity.

Grid Cylinder, G: The tubing ideally suited for this part is the type used in a hot-water baseboard-heating converter. It is important for proper operation of the cavity that the tubing wall be thin. Score a groove 5/16-inch from one end with a tubing cutter to form a shoulder on the inside of the tube against which the 2C43 grid disk will butt. This operation has to be performed with care so as not to cut through the tubing. Next, cut the slots as indicated. The fingers thus formed are then bent inward slightly until a firm clasp on the grid contact of the 2C43 is achieved.

Output R.F. Probe, J: Mount the probe in the ½-inch hole of the outside cylinder, making sure that it makes a snug connection in the

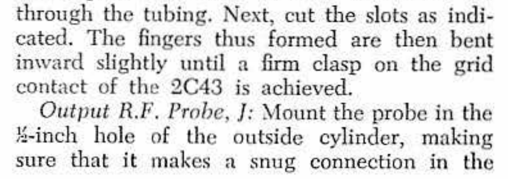


Fig. 10-58—Schematic diagram and parts information for the modulator. Values of parts are plus-or-minus 20 percent, unless specified. Output from J_2 goes to the pulse transformer in the r.f. unit.

C_1, C_2—820-pf., 600-volt, 5 per cent, silver mica.
C_3—20-pf., 600-volt, 10 per cent ceramic
C_4—330-pf., 600-volt, 10 per cent mica.
C_5—0.1-μf., 400-volt paper.
C_6—0.01-μf., 2000-volt mica.
C_7—0.006-μf., 1500-volt mica.
J_1—Closed-circuit jack.
J_2—Coaxial receptacle, SO-239.
L_1—3.8-μh., 28 turns No. 24 enamel, close-wound on ⅜-inch diam. form about 1 inch long.
L_2—13.8-μh., 54 turns No. 30 enamel, close-wound on form similar to L_1.
R_1, R_5—68,000 ohms, 5 per cent, 2 watts.
R_2, R_4, R_{10}—1 megohm, 5 per cent, ½ watt.
R_3—250,000-ohm, 2-watt potentiometer.
R_6—100,000 ohms, ½ watt.
R_7, R_8, R_{11}—47,000 ohms, 2 watts.
R_9—10,000 ohms, ½ watt.
R_{12}—47,000 ohms, ½ watt.
R_{13}, R_{14}—10,000 ohms, 10 watts.

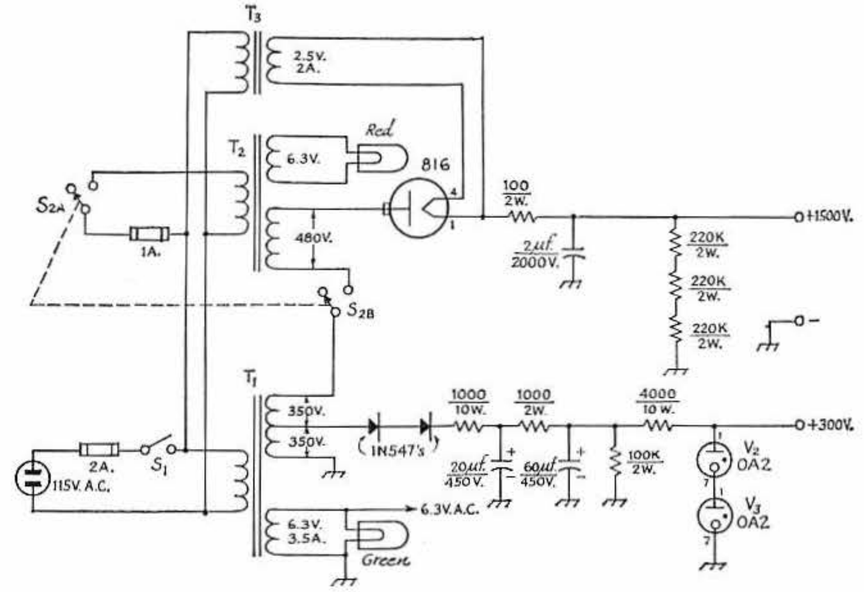


Fig. 10-59—Schematic diagram and parts information for the power-supply portion of the modulator. Components are not critical, and any supply capable of delivering 1500 volts at 10 ma. and 300 volts at 20 ma., and 6.3 volts at 3.5 amp., may be substituted. Capacitors with polarity shown are electrolytic.

S_1—S.p.s.t. toggle switch.
S_2—D.p.d.t. wafer switch.
T_1—Thordarson 24R04-U or equiv.
T_2—Thordarson 24R00-U or equiv. Do not use high-voltage center tap.
T_3—Stancor P4082 or equiv.

hole, flush against the surface of the cylinder. Two small right-angle brackets should be tailor-made to secure the r.f. connector. Two 4-40 tapped holes are made in each bracket for attaching the connector to the brackets. This assembly is shown in Fig. 10-56.

Grid Contact Assembly: The Teflon shoulder washers can be hand cut from ¼-inch rod stock using a sharp knife. When the grid contacts are assembled, check to see that the three grid fingers touch the grid cylinder when it is inserted into the cavity. The wire used to interconnect the external solder lugs of the grid contacts should be well insulated and kept clear of the outside cylinder, since peak grid potentials of several hundred volts are produced during operation.

The Pulse Modulator

Much of the modulator, Fig. 10-57, is conventional power-supply circuitry. Any supply capable of providing 10 ma. at 1500 volts and 20 ma. at 300 volts could be substituted for that shown in Fig. 10-60. The actual pulse-generating circuitry uses three tubes: two 12AU7s and a 3C45. The latter is a hydrogen thyratron available on the surplus market.

The p.r.f. is generated in the modulator, Fig. 10-58, by V_1, in a multivibrator circuit, a square-wave oscillator which has reasonably good frequency stability. It can be adjusted over a small frequency range by means of R_3. Thousand-cycle square waves appearing at the plate of V_{1B} are differentiated by a short-time-constant coupling network, C_3R_6, to produce impulses at the grid of V_{2A}. Positive and negative impulses are produced when the voltage of the square wave is rising and falling, respectively. Since the voltage at Pin 1 of V_{1B} falls faster than it rises during the square wave, a larger negative impulse is generated. This negative impulse becomes the positive trigger for firing the thyratron, V_3, after having been amplified by V_2. The operating conditions for V_2 are arranged to suppress the undesired positive impulses appearing at the first grid. V_{2A} can be keyed in its cathode circuit. When the key is closed, the stage is a conventional pulse amplifier, and the negative impulses on the grid (point B) produce positive pulses at the plate. These are applied to the grid of the thyratron by V_{2B}, a cathode follower. When the key is open, sufficient self-bias is developed across R_9 to prevent the pulses from triggering the thyratron.

The thyratron acts as a high-speed switch, closed by the trigger pulses whenever an output pulse is to be produced. The one-microsecond length of the actual output pulses of the modulator is determined by the pulse-forming network (p.f.n.) in the plate circuit of the thyratron. To create each output pulse, C_6 is first charged to almost 1500 volts by the power supply, acting through resistors R_{13} and R_{14}, L_1, L_2, and the transmitter load resistance of 50 ohms (connected to J_2). When the thyratron is fired, it becomes almost a short circuit from its plate to ground so that the energy stored in C_6 of the p.f.n. begins to discharge through the load, which is then effectively in series. The p.f.n. acts as a delay line in such a way that, one microsecond after the thyratron fires, it causes the voltage across the thyratron to be reduced to zero. When this happens, the thyratron becomes an open circuit again, and C_6 begins recharging in preparation for the next pulse. The charging resistors R_{13} and R_{14} are large enough in value so as not to affect the action of the circuit when pulses are actually produced.

The pulse-forming network has a characteristic impedance of 50 ohms which, when working into a 50-ohm load, causes the output pulses to have an amplitude equal to approximately one half the power-supply voltage. This low impedance permits a coaxial cable, such as RG-8/U, to be used to conduct the pulses to a remotely-located oscillator.

Modulator and Power Supply Layout

The modulator and its associated power supplies are constructed on a $10 \times 12 \times 3$-inch aluminum chassis as illustrated in Fig. 10-58. The power-supply section requires no special layout or critical wiring technique, other than observing insulation requirements in the high-voltage section. Transformers T_1 and T_2 are interconnected as a means of generating high voltage with readily obtainable components. Before wiring the transformer secondary high-voltage leads permanently into the circuit, check the phasing of the high-voltage windings to make sure they are aiding, not bucking.

Winding the Pulse Transformer

The winding of the pulse transformer is similar to winding a heater transformer for 60-cycle operation. The differences are the type of core used and the amount of insulation needed between the windings. The core, specially fabricated for use in pulse circuits to minimize the high-frequency eddy-current losses, can be purchased from Arnold Engineering Co., Marengo, Illinois, or through one of their many sales offices. The full description of the core is Arnold 2-mil. Silectron "C" core, part No. AL-12. Data for winding the transformer and a cross-sectional view are shown in Fig. 10-60.

Since winding wire directly on the core initially is very impractical, a wooden mandrel should be made having the same cross-section dimensions as the core. After clamping the mandrel in a vise, cut a strip of cardboard approximately 0.025 inch thick (such as used in a tube carton), 1⅛ inches wide, and wrap it tightly around the mandrel. Overlap the ends ½-inch and cement the ends together. The cardboard form prevents the windings from collapsing when removing the coil from the mandrel. This permits easy insertion of the core. Before winding, precut 1⅛-inch wide strips of Teflon from 2-mil sheet stock. Wrap two layers of Teflon around the form, securing the ends with short strips of masking tape. Next, center a 2-inch long strip of masking tape, sticky side up, across the form and wind 50 turns of No. 26 enameled copper wire over the tape. Fold the ends of the tape over the winding, thus securing the end turns. Continue with steps E, F, G, H, and J as listed in the winding-data diagram. Start all windings at the same end of the form and wind in the same direction. Label the ends of each winding according to Fig. 10-60 to facilitate wiring the transformer into the circuit.

When the winding is completed, slip the cardboard form supporting the winding stack off the mandrel and insert the core. Tape tightly around the periphery of the core with vinyl electrical tape to butt the ends of the core pieces together. Apply a finishing coat of coil varnish to the windings and the transformer is complete.

Operation

Attach a dummy load, made by paralleling five 270-ohm 2-watt composition resistors, across the modulator output jack, J_2. Set the p.r.f.-adjust control, R_3, in the center of its range, and insert a key in J_1. Turn on the heaters and 300-volt supply with switch S_1, and check the voltages in the circuit against the values given on the schmeatic. The dual voltages on Pins 1 and 3 of V_2 correspond to key-closed and key-open conditions. If the voltages agree approximately, turn on the high-voltage supply with S_2, making sure the key is open. Measure the high voltage at the power supply. It should be between +1300 and 1500 volts d.c. With the key closed, the 3C45 should ionize with a purple glow, indicating that it is being triggered properly. One should also hear a faint 1000-cycle tone. After five minutes of operation,

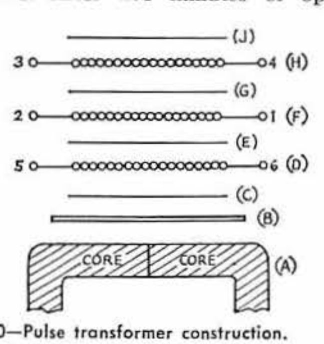


Fig. 10-60—Pulse transformer construction.
A—Core; Arnold Eng. Co. No. AL-12.
B—Cardboard form.
C—2 layers 2-mil Teflon sheet.
D—50 turns No. 26 enam.
E—4 layers 2 -mil Teflon sheet.
F—20 turns No. 22 enam.
G—4 layers 2-mil Teflon sheet.
H—50 turns No. 26 enam.
J—4 layers 2-mil Teflon sheet, followed by 3 layers masking tape.

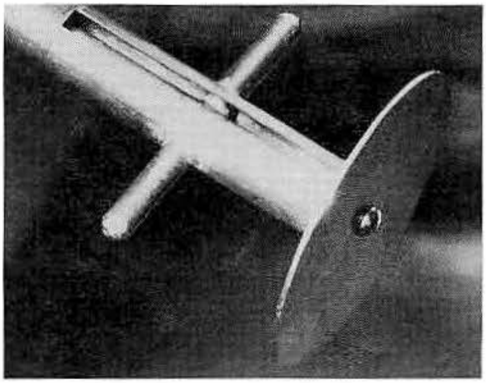


Fig. 10-61—Close-up view of the dipole and reflector assembly. For constructional details see text and Fig. 10-62.

turn off the high voltage and touch the dummy load. It should be hot, as 6 to 7 watts of average power is dissipated. For those amateurs who have a fairly good oscilloscope, waveforms are also given at four points in the circuit as a check-out aid.

Operating the Transmitter

Before connecting the cavity into the circuit, make sure that the 2C43 is seated securely and the plate-line contact is fully engaged with the 2C43 plate terminal. Connect the blower to the plate line with a short length of plastic hose and attach the heater, cathode, grid, and plate connections according to the schematic, Fig. 10-55. Connect the oscillator to the modulator with RG-8/U cable. At K1JIX, a 50-foot cable connects the modulator in the shack to the oscillator on the roof, next to the antenna. Apply 115 volts a.c. to the oscillator unit and check for air flow at the cathode end of the cavity. Allow the cathode of the 2C43 at least 60 seconds to come up to temperature before turning on the pulse modulator. When the modulator high voltage is turned on and the key is closed, one should immediately hear the pulse transformer "sing" at the p.r.f.

A simple check for r.f. output can now be made by touching a neon lamp to the center conductor of the output jack, J_2. A second check is to connect the dipole feed to the oscillator with a 5-foot length of RG-8/U or RG-9/U cable and hold a neon lamp near the dipole. The lamp should glow brightly in both cases. Between 1000 and 2000 watts peak-power output can be obtained, depending on the condition of the 2C43, the degree of coupling by the r.f. probe, and the amplitude of the pulse applied to the 2C43 plate. If output indication such as described above is produced, it is likely that the output power is at least 1000 watts. It was found that silver plating of the entire cavity assembly increased the cost considerably but did not result in any measurable increase in power out.

The remaining and most difficult part of the procedure is checking the oscillator frequency. There are at least three ways that this may be done: (1) a wavemeter, (2) a slotted line or Lecher wires, or (3) the companion receiver. The simplest way is to use a wavemeter. Several types of these are available on the surplus market. The authors acquired a surplus type 402-B coaxial wavemeter. The third method consists of using the companion receiver and calibrating it by means of harmonics from a lower frequency source. Possible sources are a grid-dip meter operating in the u.h.f. range, a 220- or 420-Mc. transmitter, or a 1215-Mc. equipment such as the APX-6.

The operating frequency of the oscillator constructed using the dimensions given here was found to be 2333 Mc. The frequency can be shifted 50 Mc. higher by shortening the grid cylinder length from 1 13/16 inches to 1¾ inches, or lower, by sliding the plate line outward on the plate terminal.

Antenna System

The antenna feed is designed for operation at a center frequency of 2360 Mc., where a wavelength is exactly 5 inches. Details of the feed are shown in Figs. 10-61 and 62. It is basically a dipole radiator, fed from a rigid coaxial line having a characteristic impedance of 50 ohms. The disk is a reflector, so there is little forward radiation from the dipole, and the main lobe of the radiation pattern is centered on the direction from which the coaxial line approaches. A parabolic dish antenna can therefore be illuminated with radiation by using the coaxial line to support the feed at the center of the dish.

The manner in which the microwave energy is coupled to the dipole elements from inside the line is interesting. Two half-wavelength slots are cut in opposite sides of the outer con-

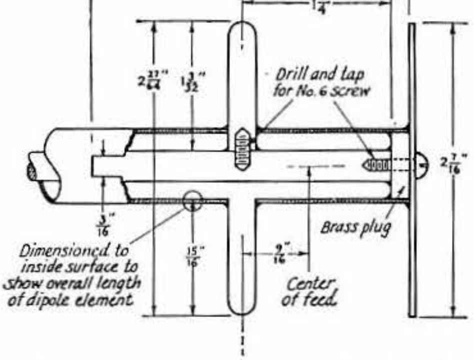


Fig. 10-62—Dimension drawing of the dipole and reflector assembly. Rigid coaxial line, left, extends to the parabolic dish reflector, and is terminated in a coaxial fitting in back of the dish. Inner conductor and dipole are ¼-inch diameter.

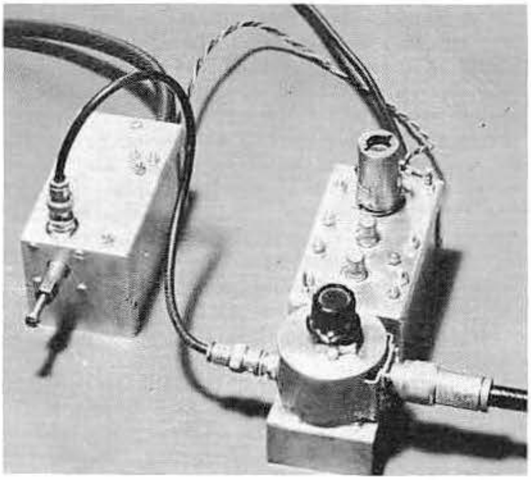


Fig. 10-63—Front end of the 2300-Mc. receiver. The box at the left contains the local oscillator. The mixer is the cylindrical assembly in the right foreground, with the cable from the antenna entering from the right. The i.f. preamplifier is just to the rear of the mixer. Knob on the top of the mixer is a tuning adjustment which was found to be unnecessary.

ductor of the coaxial line, extending back from the closed end. A short circuit is placed between the inner and outer conductors at the mid-point of these slots. This causes an r.f. potential difference between opposite sides of the inside of the line, which is transferred to the outside of the line by the slots. The dipole elements are excited by this r.f. potential difference. Although the short circuit inside the line is formed by an extension of one of the dipole elements, the dipole element and the short are electrically independent.

The effective origin or center of the radiation from the feed is located between the dipole and the disk as shown by the arrow in Fig. 10-62. To properly illuminate a parabolic antenna, the feed should be mounted so that this spot is at the focal point of the parabola.

Although dimensions are given to a 64th of an inch, errors of this amount should not affect performance noticeably. It is important, however, that each dipole element extend exactly the same distance beyond the outside surface of the coaxial line.

The outside conductor of the coaxial line is a standard size of copper water tubing. Its length depends on the size of the parabolic reflector with which it is used. The inside dimensions of the coaxial line make it convenient to mount a "type N" connector (UG-58A/U, as used for the transmitter r.f. probe) on the input end. When this is done, the center conductor is shortened somewhat with respect to the outer conductor (so as to keep the tapped hole in the side of the center conductor lined up with the dipole), drilled in the center to fit over the center conductor of the connector, and tapered or rounded off, as are the ends of the dipole elements.

In assembling the feed, first solder the center conductor to the connector. Two or three washers made from polyethylene foam or Teflon are then placed at intervals along the center conductor, to keep it in the middle of the outer conductor. The outer conductor is then slipped over it, and the longer dipole element screwed on the inner conductor. It may help to solder the No. 6 screw in place in the dipole element beforehand. The disk and brass end plug are then mounted on the end of the center conductor and the shorter dipole element inserted. (The hole in the outer conductor for this element should be made slightly undersize to give a force-fit). At this point, the assembly is complete, and all metal-to-metal joints (including that of the outer conductor with the N connector) should be soldered, using a propane torch. The inside joints of the feed are accessible for soldering through the slots. Finally, the external lengths of the dipole elements should be checked to see that they are the same; if one is longer, it can be filed down to match the other.

Pulse Reception

A block diagram and brief description of the pulse receiver were given earlier. The receiver as shown in Fig. 10-51 is complete; nothing is needed but earphones or a speaker to do an effective job in pulse reception. Reasonably good reception is possible without the special threshold detector and p.r.f. filter included here.

The r.f. portion consists of a cavity mixer and a one-tube local oscillator. These and a 3-stage

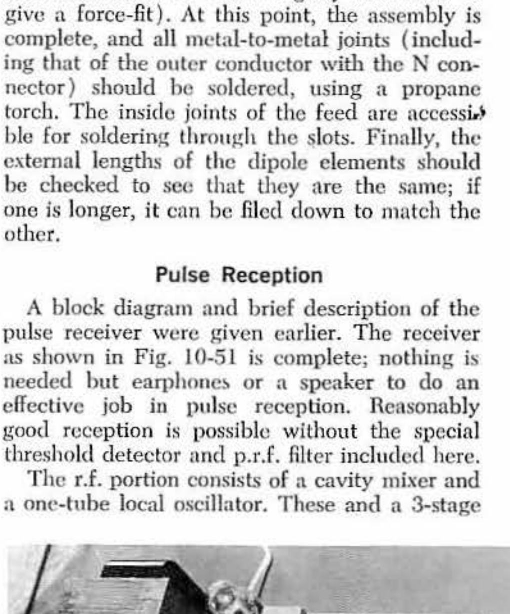


Fig. 10-64—Top view of the main receiver chassis. Controls are for i.f. gain, video gain, threshold stability, audio gain, audio bandwidth and power on-off, reading from bottom up. The first row of tubes are the video and audio stages. Power supply components are at the rear, and the i.f. amplifier stages are the three small shielded tubes in a row at the edge of the chassis. A bottom plate is required to shield the i.f. from stray 30-Mc. signals.

i.f. preamplifier are shown in Fig. 10-63. The mixer is a 1N21-series diode, in a quarter-wave resonant coaxial cavity made from standard-sized copper tubing. The 2C40 local-oscillator design was obtained from a *QST* article by W2RMA[17]

The main i.f. amplifier, the threshold detector, p.r.f. filter, audio amplifier and power supply are constructed on a $10 \times 12 \times 3$-inch chassis. This assembly, Fig. 10-64, can be remotely located from the front-end assembly, and interconnected with it by means of a coaxial cable for the i.f. signal and a power cable for the B-plus and heater voltages.

The over-all i.f. amplifier has a center frequency of about 30 Mc., though any frequency from 20 to 60 Mc. is satisfactory, provided that a bandwidth of about one megacycle can be maintained. A low-noise design with Nuvistor cascode input is used here, but it should be possible to adapt surplus i.f. amplifiers with good results.

The combination of threshold detector and p.r.f. filter is effective in detecting pulses barely exceeding the noise level at the output of the second detector. The threshold detector uses a multivibrator circuit, as shown in Fig. 10-65, and works as follows: V_3 and V_4 comprise a one-shot or monostable multivibrator, producing a single square output wave only when triggered by V_1, a video amplifier which also inverts the polarity of the positive output of the second detector. Negative noise peaks and pulses are therefore applied to the grid of V_4 by way of V_2 and C_4. R_2 is adjusted so that V_3 is normally cut off and V_4 is conducting. When a negative peak from V_1 cuts off V_4, the multivibrator "flips," and V_3 conducts for a time before returning to the original conditions. Duration of the positive pulse at the plate of V_4 is determined by C_4 and the 1.5-megohm grid ressistor of V_4. For values given, the pulse out of the multivibrator is about 35 microseconds long.

The amplitude of the negative pulse required to trigger V_4 can be varied by means of R_2. This control and the video amplifier gain control, R_1, therefore serve as threshold level adjustments. The multivibrator threshold is set so that, in the absence of a pulsed signal, it is triggered by noise peaks several times per sec-

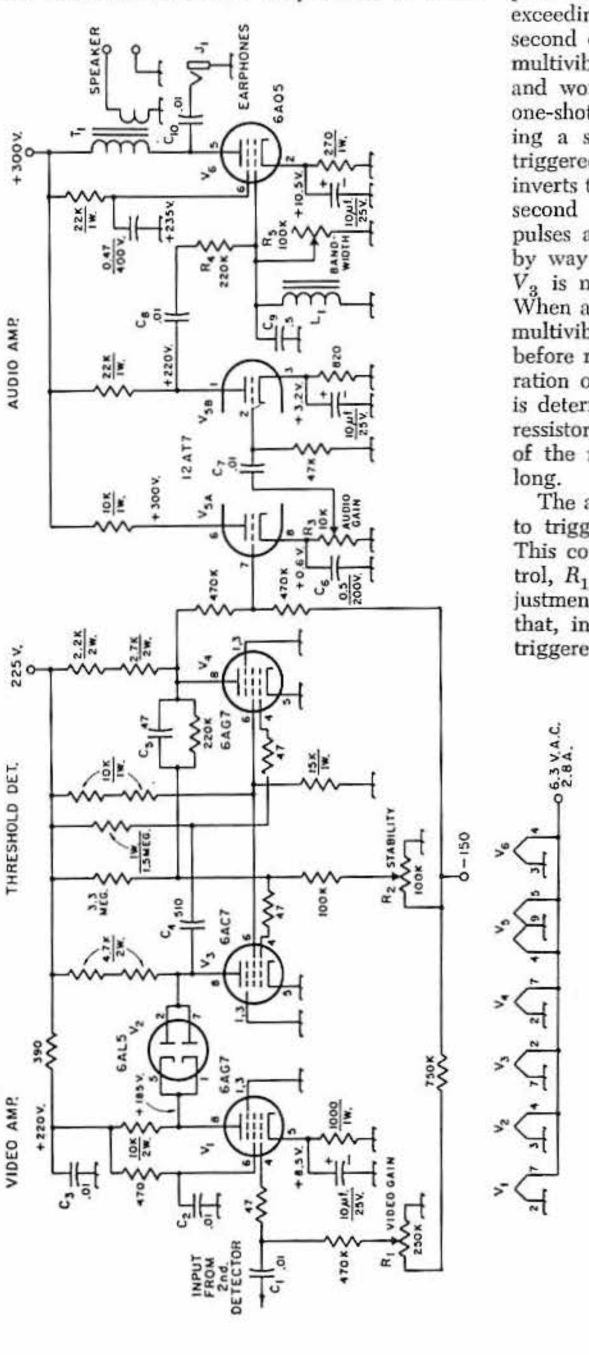


Fig. 10-65—Schematic diagram of the video amplifier, threshold detector and audio stages. Resistors are ½ watt, 10 per cent tolerance, unless specified. Decimal values of capacitance are in μf., others in pf., unless specified. Those marked with polarity are electrolytic.

C_1, C_2, C_3, C_7, C_8, C_{10}—Paper or ceramic, any tolerance.

C_4, C_5—Mica, 10 per cent.

C_6—Paper, 20 per cent.

C_9—0.5 μf.; see text.

J_1—Phone jack.

L_1—50-mh. high-Q toroid (Freed F-1755). See text.

R_1, R_2, R_3, R_5—2-watt potentiometer.

R_4—See text.

T_1—Audio pentode output transformer (Stancor A3822).

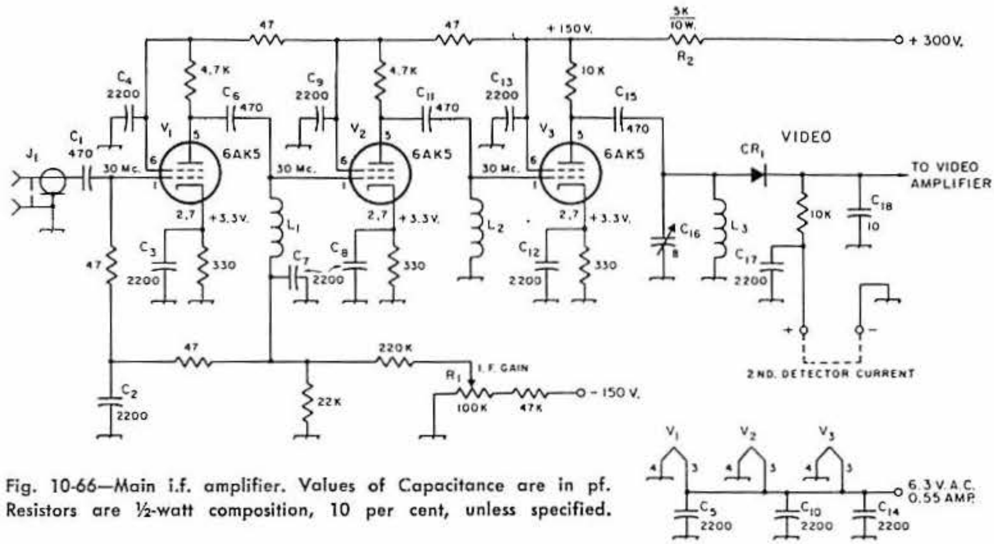


Fig. 10-66—Main i.f. amplifier. Values of Capacitance are in pf. Resistors are ½-watt composition, 10 per cent, unless specified.

C_1–C_{15} incl.; C_{17}—ceramic disk.
C_{16}—1-8-pf. cylindrical trimmer (Erie 532B).
C_{18}—10 pf., 10 per cent, mica.
CR_1—Chystal diode, 1N60, 1N67, 1N295 or equiv.
J_1—Coaxial receptacle, BNC type.
L_1, L_2, L_3—2.2 μh., 32 turns No. 24 enam., close-wound on ¼-inch diam.
R_1—2-watt potentiometer.
R_2—5000 ohms, 10 watts, wire-wound.

ond. When even a weak signal appears, there is a pronounced increase in the triggering rate. When a moderately strong signal with the correct p.r.f. appears, a noise-free tone is produced at the output of the p.r.f. filter. Difference between the signal and no-signal output conditions is quite distinct, and the effect is similar to that of a squelch circuit. With freedom from a constant level of background noise, searching for weak signals is relatively easy on the ears.

The three stages following the threshold detector make up a very narrow-band audio amplifier centered at 1000 cycles. Most of the selectivity is provided by a 1000-cycle filter formed by inductor L_1, resonating with C_9, between V_5 and V_6. A small but important amount of filtering is provided in the cathode circuit of V_{5A}. Capacitor C_6 removes undesired high-frequency components of the multivibrator square wave, so that they will not overload the following triode amplifier. The *LC* filter uses a high-*Q* toroid to obtain an audio bandwidth as small as 10 cycles. The result is that unless the square waves produced by the threshold detector-multivibrator have a p.r.f. of exactly 1000 cycles, there is little output from the audio amplifier. The net effect of these circuits is a signal-to-noise ratio somewhat comparable to that obtained with narrow-band c.w. on lower frequencies.

The Main I.F. Amplifier

The main receiver chassis is seen in Fig. 10-64. It is recommended that this general arrangement be used, but exact wiring details are not important except in the case of the i.f. amplifier circuitry. The three i.f. amplifier stages, Fig. 10-66, are in a line along the lower side of the chassis, with the input stage toward the left where there is little neighboring wiring. The second detector is near the video amplifier. In order to reduce cost, identical fixed inductors are used to tune each stage. The exact frequency is not important, so long as each is tuned to the same frequency. Since the coils resonate with the capacitance of the interstage wiring and the input and output capacitances of the tubes, the wiring layout between each pair of tubes must be identical. Similarly, the coils should be identical, though it is not important exactly how they are wound. The wiring of the plate circuit of the last stage is not critical, as a variable capacitor, C_{16}, allows for differences in circuit capacitance.

The *Q*, and therefore the bandwidth, of each i.f. stage is determined by the plate load resistance (in the output stage, the detector load resistance appears in parallel with the plate load). For a value of 4700 ohms, the *Q* is 10, so the bandwidth of each stage is about three megacycles. Over-all bandwidth, including the preamplifier, approaches one megacycle, but it will depend on alignment.

Variable inductors could be used for each stage, to avoid the need for careful layout and permit exact alignment. A small coil such as the Miller type 5403, 1.6 to 2.8 μh., should work well, or similar surplus slug-tuned coils could be used. An inductance of the proper value will resonate at 30 Mc. with a 12-pf. capacitor.

P.R.F. Filter

The toroid, L_1 in Fig. 10-65, has a *Q* of approximately 200 by itself, and effective circuit

Q of about 100, due to loading by R_4 and R_5. A Freed F-804 ($6.60) would give a Q of 70, and be entirely adequate. Suitable toroids may be available on the surplus market. Qs as low as 20 to 30 would still give good results. Inductance values other than those given can be used by changing R_4 and R_5 in the same proportion. Example: if the inductance is increased 10 times, to 0.5 henry, R_4 becomes 2.2 megohm and R_5 1 megohm.

Capacitor C_9 is actually several capacitors in parallel. A 0.47-μf. capacitor is used with enough 0.01-μf. or 0.0047-μf. capacitors in parallel to resonate the toroid to exactly 1000 cycles. These capacitors should be low-loss types, to preserve circuit Q.

Mixer

Crystal diodes of the 1N21 series have letter suffixes from A to F, each giving progressively lower noise figure and costing more. Very satisfactory results have been achieved with the inexpensive 1N21C. Mounting of the mixer to the preamplifier is important, as the diode forms part of the amplifier input circuit. The mixer cylinder must be well grounded to the preamplifier chassis, so that it will not act as an i.f. pickup loop and cause the amplifier to oscillate.

Construction details of the microwave mixer and a cutaway view of the complete assembly are given in Fig. 10-67. Following is the sequence for assembling the mixer. Attach the center rod to the end disk with the hole, using a 6-32 screw, and solder with a propane torch. Attach the receptacle that receives the pin end of the 1N21 diode to the center rod. This is a demountable base, made for reversible-case cartridge-type diodes, and is available from any manufacturer of microwave diodes. A substitute can be made from ¼-inch brass rod 7/32 inch long, drilled out 3/32 inch at the center. Slot it for most of its length with a thin saw blade, and bend the fingers in slightly so that it will make good contact to the diode pin. Clamp the base in the notch on the center rod, and solder the connection, being careful not to fill in the slots with solder. Attach this assembly to the outside cylinder, and align the 9/32-inch hole in the cylinder wall with the crystal receptacle on the center post, using a crystal diode. Clamp securely, and solder the end disk to the outside cylinder. Clamp the other end disk to the top of the cylinder and solder securely.

To mount the l.o. connector, center and clamp the nut supplied with the UG-1094/U connector over its hole, and solder, being care-

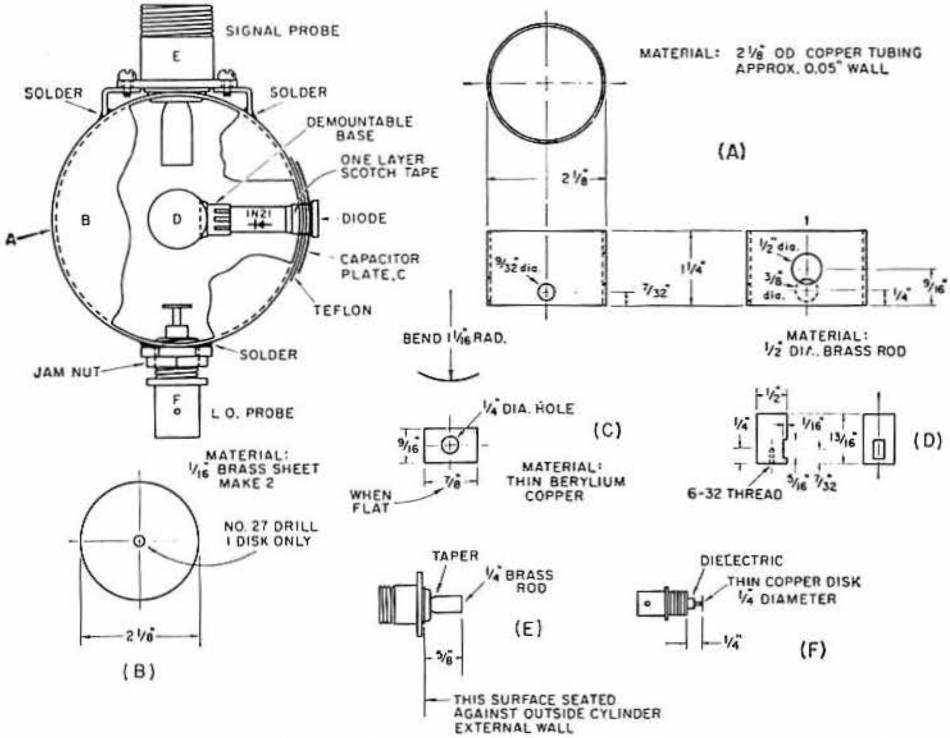


Fig. 10-67—Cutaway view of the crystal-mixer assembly, with details of individual parts. As may be seen from Fig. 10-63, the mixer cavity is mounted adjacent to the end of the i.f. preamplifier. The head of the diode, and its external bypass capacitor, project through a rectangular hole cut in the end of the preamplifier case. A ¾ by ¼-inch spring finger on a standoff insulator makes contact to the diode, to take off the 30-Mc. i.f. output.

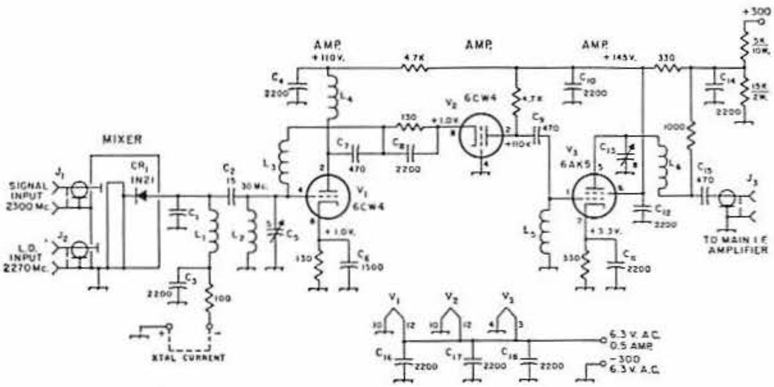


Fig. 10-68—2300-Mc. crystal mixer and 30-Mc. I.f. preamplifier. Capacitors are ceramic disk, (CD Tinymike) values in pf., 20 per cent, unless specified. Resistors are ½-watt composition.

C_1—Bypass capacitor built onto mixer assembly; see Fig. 10-68.

C_3, C_4, C_8, C_{10}, C_{11}, C_{12}, C_{14}, C_{16}, C_{17}, C_{18}—2200-pf. ceramic disk ,CD Tinymike LT0D22).

C_7, C_9, C_{15}—470-pf. ceramic disk (CD Tinymike L10T47).

C_2—15 pf., 10 per cent, mica.

C_5—5-pf. cylindrical trimmer (Erie 532A).

C_6—Critical lead length; see text. (CD Tinymike L10D15)

C_{13}—8-pf. cylindrical trimmer (Erie 532B).

CR_1—1N21-series crystal diode; see text.

J_1—Coaxial receptacle, UG-58/U (Type N).

J_2, J_3—Coaxial receptacle, UG-1094/U (Type BNC).

L_1—22μh., plus or minus 10 per cent (Delevan 1537-44).

L_2—1.3 μh., 21 turns No. 24 enam., close-wound on ¼-inch diam. form.

L_3—22 μh., 65 turns No. 32 enam., close-wound on ⅜-inch diam. form; see text.

L_4, L_5, L_6—2.2 μh., 32 turns No. 24 enam., close-wound on ¼-inch diam. form.

ful not to flow solder into the nut threads. The l.o. probe can then be screwed in or out, to adjust the injection level. A second nut threaded onto the probe can be used to lock it in position, once the desired level is obtained. This and the method of attaching the signal probe to the cylinder are shown in Fig. 10-67.

The mixer is attached to the preamplifier by two brackets. A clearance hole is first cut in the wall of the preamplifier chassis, large enough to clear the crystal bypass capacitor, C_1, which is assembled later. Details of the bracket depend on the chassis used for the preamplifier. Before mounting the crystal diode, cut a piece of 2-mil Teflon sheet slightly larger than the crystal bypass plate. Carefully cut a true ¼-inch hole in the center of the dielectric. Wrap a layer of Scotch tape around the base of the diode to prevent it from shorting against the wall of the outside cylinder. Slide the capacitor plate and the Teflon sheet over the diode. Insert it into the mixer cavity and seat the diode flange firmly against the capacitor plate. Tape the capacitor plate to the outside cylinder so that the crystal can be removed without disturbing the capacitor. A small beryllium spring finger mounted on a standoff insulator holds the crystal in place, and serves as the i.f. signal connection.

Local Oscillator

Both oscillators described by W2RMA have been constructed and used with the receiver described here, with good results. Several other approaches are open, the most obvious being to use the oscillator from a surplus microwave unit. A few such possibilities are mentioned later. The transmitter oscillator and local oscillator of the APX-6 immediately suggest themselves for this application with a diode multiplier. Such an oscillator could be located at the operating position and connected to a varactor mounted in the mixer at the antenna. This arrangement could be quite simple, as only a milliwatt or less power is required for l.o. injection at 2300 Mc.

I.F. Preamplifier

The i.f. amplifier has a noise figure of only slightly more than 1 db., due mainly to the Nuvistor cascode input stages. Low-loss coils at L_1, L_2 and L_3 in Fig. 10-68 are important, as is the small value of coupling capacitance, C_2. This transforms the low impedance of the crystal up to the higher value which is optimum for the 6CW4 at 30 Mc. A critical value of cathode bypass is used for C_6. This should be wired into the circuit with ¼-inch leads, to tune out the cathode lead inductance of the first stage. L_1 can be made like L_3, if desired.

The three tubes of the preamp should be arranged in line, with components mounted close to the tube sockets they serve. Lead lengths should be kept to ½ inch or less. The 6CW4–6AK5 wiring should be like that of the main i.f. amplifier. Since the preamp makes an extremely sensitive 10-meter front end, a shielded chassis such as a Minibox, and a 6AK5 tube shield, should be used. The dropping resistors and bleeder in the B-plus circuit are arranged so that the supply voltage will not be excessive during warm-up periods.

Operation

Before reception is attempted, the l.o. injection, i.f. amplifier and threshold detector should be checked out. Operate the main receiver chassis first, without the preamplifier or local oscillator connected. Before applying power, set the i.f. and video gain controls at maximum (fully clockwise) and the remaining potentiometers fully counterclockwise. After turning on power, the VR tubes should light up in a few seconds, and the voltage after R_2, Fig. 10-66, should read about 150, as soon as the heaters warm up. Other voltages should read about as given.

Connect a speaker or earphones to the output, turn the audio gain and bandwidth controls fully clockwise, and turn the i.f. and video gain controls fully counterclockwise. As the stability control is turned up slowly, a popping sound should be heard, followed by a weaker squeal, whose pitch varies with rotation. Leave the stability control set just below where the popping is first heard.

The main i.f. amplifier can be aligned with a signal source such as a grid-dip oscillator, using a few inches of wire in the input BNC connector as an antenna. Connect a milliammeter in the second detector circuit, as indicated in Fig. 10-66. Place the bottom cover on the chassis, and turn the i.f. gain full on. The 30-Mc. signal source should be coupled to the input so as to give about 1-ma. output current. Without changing input coupling, tune the signal source until a frequency is found which gives the highest current when the output stage is peaked by adjusting C_{16}. This is the center frequency of the response of the three stages together, and is about 31.5 Mc. for the receiver shown.

Next, connect power to the i.f. preamp and the local oscillator, but do not connect the l.o. or i.f. signal cables. Check the cathode voltages of the preamp. With the preamp and mixer completely assembled, connect the preamp to the main receiver with a 50-ohm cable, and tune the signal source to the center frequency of the main i.f. amplifier. With a few inches of wire connected to the signal-input jack of the mixer, peak C_5 and C_{13} to the i.f. signal, with the meter showing about 1 ma. Move the signal source away, or use a 2300-Mc. signal later, to peak C_5 accurately.

To check mixer operation, remove the jumper from the crystal-current meter terminals on the preamp, and connect a 1-ma. meter with the polarity shown. Decouple the l.o. probe, J_2, from the mixer cavity as far as possible, and connect a short coaxial cable from the l.o. Adjust the l.o. output for 0.3- to 0.8-ma. crystal current. Increase injection by threading the l.o. probe farther in, if necessary. When the proper level is obtained, lock J_2 in place with the jam nut. The meter can be removed and the jumper replaced.

The receiver is now ready for use on 2300 Mc. It will be found that front-end thermal noise is capable of continuously triggering the threshold with the video and i.f. gain at maximum. The most sensitive condition for weak signals is with the i.f. gain near maximum and the video gain turned down until the threshold is triggered on noise peaks a few times per second. The i.f. gain control is then used for minor adjustment of the triggering rate. It should not be left turned down for long periods, as the B-plus voltage after the main i.f. dropping resistor, R_2, will be excessive in this condition. No trouble has been experienced with this, but if it becomes a problem it can be eliminated through the use of a VR-150 regulator after R_2.

The final step is to check the tuning of the p.r.f. filter, by varying the transmitter p.r.f. and watching for the peak in audio output.

Conclusions

It should be emphasized that this description is only one of many possible approaches. The equipment could be improved in many ways, and could be adapted to 3300 Mc. with only minor modification. There are also many intriguing possibilities in the large amount of surplus radar pulse equipment now available. Some useful equipments are listed below:

APR-4, with tuning head TN-54 Radar receiver 2150 to 4000 Mc. Noise figure poor. L.o. in tuning head.

APR-5A Radar receiver, 1000 to 5000 Mc., in one assembly. Noise figure probably poor.

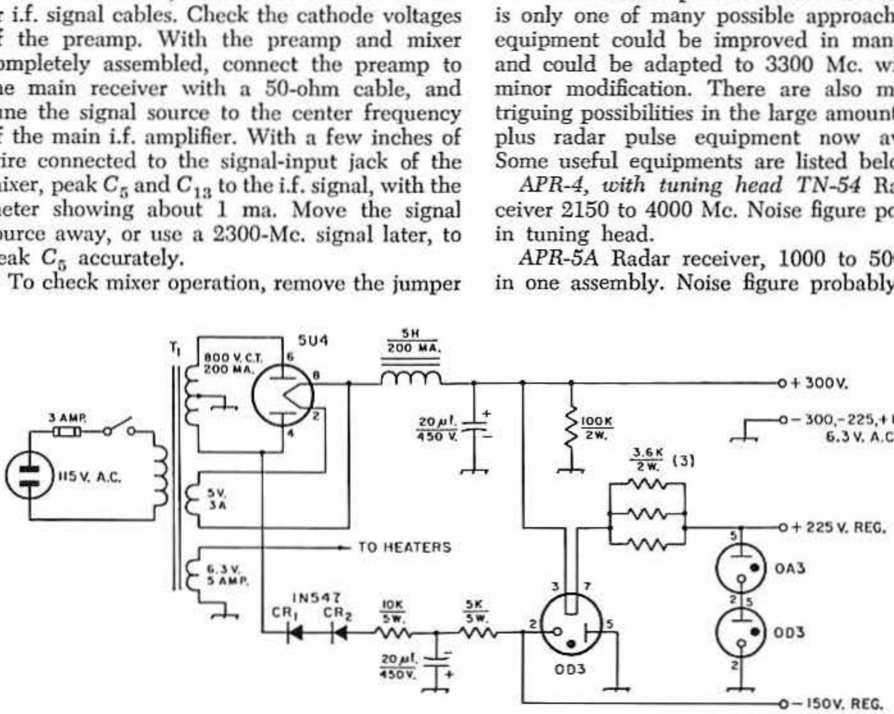


Fig. 10-69—Power supply for the 2300-Mc. receiver. Capacitors are electrolytic, polarity as indicated. T_1—Stancor 8412, or equivalent.

APR-9 with tuning head TN-128 Radar receiver, 1000 to 2600 Mc. Noise figure poor. L.o. in tuning head.

APG-5 or APG-15 Radar for B-29 tail gun, 2700 to 2900 Mc. Has pulsed 2C43 transmitter and 2C40 l.o., in cavities similar to those described in this series. Probably convertible to 2300 Mc. without too much difficulty.

2J39 Integral-magnet magnetron. 9-kw. peak power output, 3267 to 3333 Mc.

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21,000 Mc.:

Sharbaugh and Watters, "Our DX—800 Feet!" August, 1946, *QST*. Same Authors, "World Above 20,000 Mc.," May, 1959, *QST*., includes information on equipment for 50,000 Mc.

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Scott and Banta, "Using the Helical Antenna on 1215 Mc., July, 1962.

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"Wireless Lecher Wires," Cover and p. 10, September, 1948, *QST*.

"W1QWJ 432-Mc. Kilowatt Amplifier," February, 1966.

Transistor Converters for 432: Clark, December, 1965, and 1968 Handbook; Brannin, June, 1966

Jensby, "Stable Microwave Oscillators," July, 1966.

Poland, "Converting Wideband F.M. Equipment for 420–Mc. Service," August, 1968, *QST*.

For those amateurs who wish to acquire a more basic understanding of microwave circuits and techniques, the following low-cost publications are available from the Superintendent of Documents, Government Printing Office, Washington 25, D. C.

Generation and Transmission of Microwave Energy, Cat. No. D101.11:11–673. 204 pp., ill. Price, $1.00.

Microwave Techniques, Cat. No. D211.2:M58. 188 pp., ill. Price, 55 cents.

Microwaves and Waveguides, Cat. No. D211.2 M58/2. 56 pp., ill. Price, 25 cents.

Radar Electronic Fundamentals, Cat. No. N29.2: R11/3. 474 pp., ill. Price, $1.50.

Radar System Fundamentals, Cat. No. N29.2: R11/2. 394 pp., ill. Price, $1.25.

Radar Circuit Analysis, Cat. No. D301.7:52–8. 480 pp., ill. Price, $5.25.

Pulse Techniques, Cat. No. D101.11:11–672. 102 pp., ill. Price, 55 cents.

Most of these are training manuals used by the armed services. They are clearly written, with numerous illustrations.

Test Equipment for the V.H.F. Station

Many v.h.f. men tend to be experimenters at heart. Much of the pleasure and satisfaction to be derived from v.h.f. work comes from striving for the best possible performance from our stations, whether they are lavish or modest in nature. Thus it is important that we acquire suitable test equipment, and the know-how to use it effectively. This does not necessarily mean a laboratory full of expensive gear. Much of what is needed can be provided at moderate cost, particularly if we are willing to take the time and trouble to make some of its ourselves.

All ham gear costs money, and the average amateur investment in equipment is rising every year. Regrettably, this does not always apply to the percentage of the total budget that is devoted to test equipment. Yet whether we buy our operating gear or build it ourselves, only by having adequate testing and measuring equipment, and using it regularly and well, can we be sure that we are getting maximum return from the money spent on transmitters, receivers and antennas. Nowhere is this more true than in the v.h.f. field.

A good deal of amateur test equipment is essentially the same, regardless of the frequencies we intend to concentrate on. Things like volt-ohmmeters, oscilloscopes, meters in general, the audio oscillator, the grid-dip meter—these are standard items that we will not dwell on extensively in these pages. Rather, we will devote our space mostly to simple gadget-type items, tailored to the v.h.f. man's needs. We'll start with the simplest of them all.

LAMPS—POOR MAN'S METERS

Good meters are a very worthwhile investment. Every ham should have a few, but they do not have to be built into everything. Knowing what to test and measure, and being able to make a meaningful interpretation of what the meters tell us, is more important than having shelves full of expensive test equipment. An experienced and knowing ham can do good work with a meter or two, a few pilot lamps, neon bulbs and some simple wavemeters. There is no substitute for experience in this department, and the way to get the experience and know-how is to start right in, building simple equipment and making it work.

Current and Power Indicators

More can be done with pilot lamps as current and power indicators than most hams realize. The information of Table 11-I can be put to use in a number of ways. Most of the lamps listed are available in radio or even hardware stores. The rating in amperes, right column, tells us that these lamps draw 0.06, 0.1, 0.12, 0.15, 0.17, 0.2, 0.25, 0.4 and 0.5 amperes, at normal brilliance. These figures (60 to 500 milliamperes) are close to ranges of milliammeters we would buy if we had an unlimited budget. Obviously, lamps can replace meters in many uncritical circuits where we need merely an approximate indication of current being drawn. Lamps are fuses of a sort, too; they'll burn out if current goes very high for any reason.

Lamps serving as current and power-output indicators are shown in Fig. 11-1A. Lamp I_1 is a plate-current "meter." Connected in the line from the plate circuit, it gives an indication of plate current being drawn. If the stage is supposed to draw 30 to 60 ma., a 2-volt 60-ma. pilot lamp (No. 48 or 49) will do. It can be mounted in a socket, to make it easy to replace if burned out. You can burn out a lot of them for the price of one good meter.

Lamp I_2 is a dummy load and power-output indicator. The coil L_1 is a single turn of insulated hookup wire about one inch in diameter. With the loop coupled to an r.f. circuit carrying power, the lamp will light due to power absorbed. With a very low-power circuit the loop can be hung over the coil in the plate circuit, or inserted between its turns. For looser coupling, needed where the power output is much over 1/10 watt, the lamp can be taped to an insulating rod to be used as a handle to hold the loop at the desired distance from the tank circuit.

We can tell several things from these two lamps. They can be calibrated roughly for both current and power from known d.c. sources. At a current of 50 ma., which it will draw from a 1.4-volt cell connected as in 11-1B, a No. 48 or 49 bulb gives off considerably less light than at its rated 60 ma., but still enough to be clearly visible. From its normal rating, we know that full brilliance indicates 120 milliwatts of power.

Table 11-I—PILOT-LAMP DATA

Lamp No.	Bead Color	Base (Miniature)	Bulb Type	RATING Volts	RATING Amp.
40	Brown	Screw	T-3¼	6–8	0.15
40A[1]	Brown	Bayonet	T-3¼	6–8	0.15
41	White	Screw	T-3¼	2.5	0.5
42	Green	Screw	T-3¼	3.2	**
43	White	Bayonet	T-3¼	2.5	0.5
44	Blue	Bayonet	T-3¼	6–8	0.25
45	*	Bayonet	T-3¼	3.2	**
46[2]	Blue	Screw	T-3¼	6–8	0.25
47[1]	Brown	Bayonet	T-3¼	6–9	0.15
48	Pink	Screw	T-3¼	2.0	0.06
49[3]	Pink	Bayonet	T-3¼	2.0	0.06
49A[3]	White	Bayonet	T-3¼	2.1	0.12
50	White	Screw	G-3½	6–8	0.2
51[2]	White	Bayonet	G-3½	6–8	0.2
53	—	Bayonet	G-3½	14.4	0.12
55	White	Bayonet	G-4½	6–8	0.4
292[5]	White	Screw	T-3¼	2.9	0.17
292A[5]	White	Banonet	T-3¼	2.9	0.17
1455	Brown	Screw	G-5	18.0	0.25
1455A	Brown	Bayonet	G-5	18.0	0.25
1487	—	Screw	T-3¼	12–16	0.20
1488	—	Bayonet	T-3¼	14	0.15
1813	—	Bayonet	T-3¼	14.4	0.10
1815	—	Bayonet	T-3¼	12–16	0.20

[1] 40A and 47 are interchangeable.
[2] Frosted bulb.
[3] 49 and 49A are interchangeable.
[5] Use in 2.5-volt sets where regular bulb burns out too frequently.
* White in G.E. and Sylvania; green in National Union, Raytheon and Tung-Sol.
** 0.35 in G.E. and Sylvania; 0.5 in National Union, Raytheon and Tung-Sol.

We thus have four useful references with which to work: 50 ma. and 60 ma. of current, and 70 and 120 milliwatts of power. Similar references can be set up easily for other lamps in the table. These are rough indications, of course, but there are plenty of places where they are good enough.

Using Neon Lamps

Neon lamps of various sizes are handy as indicators of r.f. fields, as they will ignite when merely held near a hot circuit, if the field is strong enough. Held close to a transmitter tank circuit, or to the end of an antenna carrying r.f. power, they will glow with a brilliance and color that conveys several bits of information. The brilliance is a relative r.f. power indicator, and the color is somewhat of an indication of frequency. At the low end of the r.f. range the glow is a bright orange, changing gradually to purple with increasing radio frequency. In the v.h.f. range the neon glow is markedly more

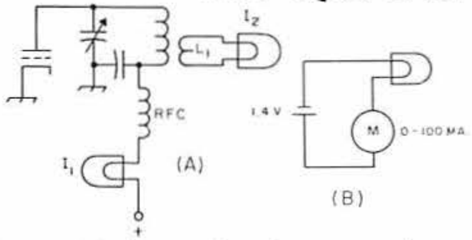

Fig. 11-1—Pilot lamps can be used as current and power output indicators. Lamp I_1 can substitute for a plate-current meter, where only a rough indication is needed. I_2, inductively coupled to the plate circuit, provides an indication of r.f. power output. Lamps can be calibrated roughly for brilliance, as shown at B. Here a 2-volt 60-ma. pilot lamp is connected in series with a single flashlight cell and a 100-ma. meter, for calibration purposes.

toward the purple than the orange. Glow color can thus be an indication of the frequency range of a parasitic oscillation, as an example.

A neon lamp or pilot lamp connected at the center of a half-wave dipole gives an indication of relative power at some distance from a transmitting antenna. The pilot lamp is the better for this purpose, for the neon must be subjected to a fairly high r.f. voltage to ignite the gas, whereas the filament of a lamp glows at once from the heat being dissipated.

Any of the lamps mentioned can be an aid to making a rough check on frequency, when absorption wavemeters or Lecher wires are used. We'll cover these applications later in this section.

Neons or pilot lamps can be used to give a rough indication of standing-wave ratio on open-wire or other balanced transmission line. A lamp with a coupling loop attached, or a neon bulb held close to the line, will brighten or dim in proportion to the r.f. voltage along the line. Two lamps, one at the minimum and another at the maximum point of r.f. voltage, thus pro-

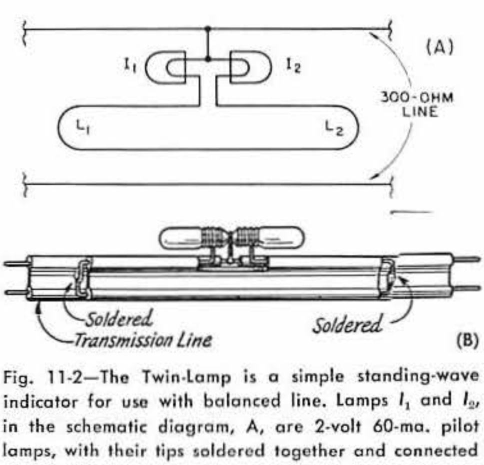

Fig. 11-2—The Twin-Lamp is a simple standing-wave indicator for use with balanced line. Lamps I_1 and I_2, in the schematic diagram, A, are 2-volt 60-ma. pilot lamps, with their tips soldered together and connected to one side of the 300-ohm line. Loops L_1 and L_2 can be made from short pieces of Twin-Lead, as shown at B.

vide a direct indication of s.w.r. that can be roughly calibrated against brilliance indications from a similar lamp in a metered d.c. circuit.

The Twin-Lamp S.W.R. Indicator

A fine example of the use of lamps to do a much-needed job is the Twin-Lamp s.w.r. indicator originally described by v.h.f. enthusiast W4HVV years ago in *QST*.[1] It works on the principle of separating the outgoing from the reflected power on a transmission line, by a combination of inductive and capacitive coupling. The Twin-Lamp is merely two 60-ma. pilot lamps connected to two coupling loops, and to one side of the transmission line, as shown in Fig. 11-2. The schematic circuit is given in A, and the mechanical details in B.

The Twin-Lamp can be made up permanently attached to a short section of Twin-Lead, which can be inserted in the transmission line as needed, or left permanently connected. A similar arrangement can be made for open-wire lines. Length of the loops, L_1 and L_2, depends on the frequency and power level. Two to four inches in length should be suitable for v.h.f. work at average power levels.

The lamp nearest the transmitter end will light when power is fed to the line. If the s.w.r. is low, the lamp at the other end will remain dark. At moderate ratios the lamp on the antenna side will glow, and with a bad mismatch the two will be about the same brilliance. The device should not be thought of in terms of *measurement* of s.w.r., but it does give a usable indication of adjustment, and it tells whether the s.w.r. is high or low, which is all that is really important in most instances.

ABSORPTION WAVEMETERS

One of the most useful items of test equipment in a v.h.f. station is the absorption wavemeter. Fortunately, it is also simple and inexpensive. Anyone who does much work with v.h.f. equipment should have a few of them. Three that take care of basic v.h.f. requirements are shown in Figs. 11-3 to 6.

A wavemeter can be a sophisticated device, but in its simplest form it is merely a tuned circuit whose resonant frequency for various capacitor settings has been calibrated. When the wavemeter is coupled to a circuit carrying r.f. power it will absorb some of the power when tuned through the frequency of the source. Indication that this is happening can be obtained in several ways.

If the stage to be checked is an oscillator, tuning the wavemeter through its frequency may cause a shift in frequency. This can be heard on a receiver tuned to the oscillator frequency. The plate or grid current in an oscillator or amplifier stage will nearly always show some change, and the power output will dip, so any of these indications, or a combination of them, can be employed. The frequency of an oscillating detector can be checked with the wavemeter in a similar manner. The hiss level of a superregenerative receiver will dip when the wavemeter tunes through its receiving frequency. The regenerative detector may go out of oscillation, or the frequency of the beat note will change, due to wavemeter absorption.

The approximate frequency of any unwanted oscillation in a transmitter or receiver can be learned with the wavemeter in much the same way. Check the grid current in the oscillating stage, and watch for a quick change as the wavemeter is tuned. In a receiver, listen to the effects of the oscillation, and note when they change abruptly in any way as the wavemeter is adjusted.

Wavemeters can be made to cover wide ranges of frequency with rough calibration, or narrower ranges with more accurate calibration. We show both types here. With the wide-range type, frequencies go by so rapidly as the tuning capacitor is rotated that the frequency indication is necessarily only approximate, but it is a very useful tool. Two of our wavemeters are this kind. The first tunes from 45 to 155 Mc., giving coverage of the 50-Mc. band, the 144-Mc. band, and other highly useful spots such as 48, 72, 96 and 100 Mc., as well. The other spans 113 to 255 Mc., passing through the 144-

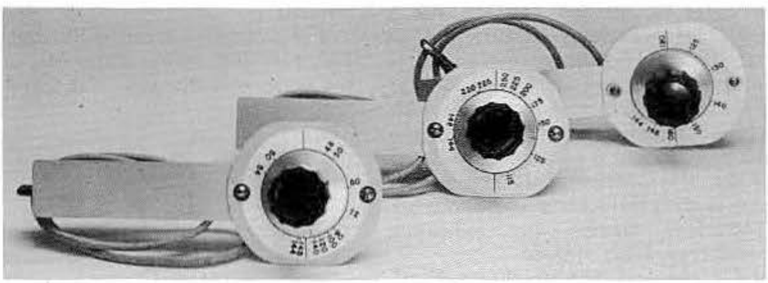

Fig. 11-3—Three simple wavemeters for v.h.f. use. The two at the left have wide-range circuits, while the third is a band-spread version. Each has a clip lead connected to the handle, for grounding to the chassis of the equipment being tested, in the interest of safety.

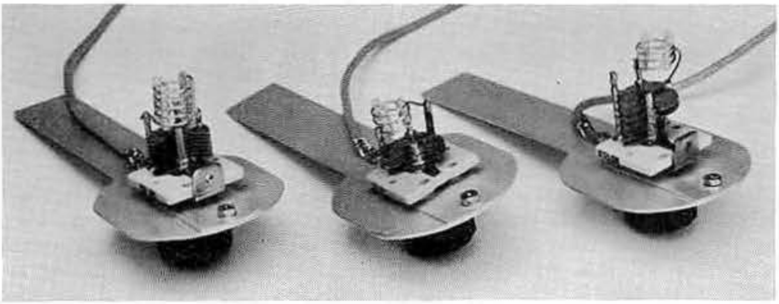

Fig. 11-4—Rear view of the wavemeters, in the same order as in Fig. 11-3.

and 220-Mc. bands along the way. The third is the bandspread type, having a fixed capacitor in parallel with the variable one. Covering 120 to 160 Mc., it spreads the 144-Mc. band enough so that a fairly good indication of actual frequency is possible, if the wavemeter is made, calibrated and used with care.

Construction

The wavemeter should be solidly built, to retain its calibration in normal handling and use. Ready-made coil stock provides the required rigidity, if connected with the absolute minimum of lead length. Some kind of indicating knob and scale arrangement is desirable, so that a permanent calibration can be made. Construction should be such that the instrument can be held in the hands without affecting the calibration, and used with reasonable safety around live circuits.

Many arrangements are possible; those shown are merely one approach to the problem. Each is mounted on two pieces of aluminum, a handle and a panel area. The handle is about 1 by 6 inches and the panel is roughly 2½ inches in diameter. One of the screws holding these pieces together has a soldering lug attached, with a clip lead soldered to it for grounding the wavemeter frame, in the interest of safety. The variable capacitors are the type in which the rotor is grounded directly by the mounting shaft and nut, these being less subject to hand-capacitance effects than variables with ungrounded mounting studs. One type of coil stock (B & W Miniductor No. 3002, having No. 20 tinned wire, ½ inch in diameter, 8 turns per inch) is used in all three.

The wavemeter for 50 and 144 Mc. has a 50-pf. variable (Hammarlund HF-50) and 4¾ turns of coil stock. The bottom of the coil is soldered to the right stator post (as we look at the back of the wavemeter) with less than ⅛ inch of exposed lead. The wire end of the top of the coil is ⅞ inch long, its last ¼ inch being soldered to the rotor lug. The position of the coil is such that it is coaxial with the rotor shaft. Knobs are Johnson 116-222-5, though any indicating pointer or dial-type knob will do.

The unit for 144 and 220 Mc. has a 15-pf. variable (Hammarlund HF-15) and roughly 2½ turns of coil stock. The top edge of the coil lines up with the center of the rotor shaft in this model. The top lead is about ¾ inch long, and the bottom one substantially nil.

The band-spread model uses a double-spaced capacitor, originally 15 pf. (Hammarlund HF-15X) with all but the back two rotor plates removed. The stator is left intact. The coil is practically identical to that in the second unit, except that the top lead is 1⅛ inches long. The coil is connected between the left stator post and the rotor lug. Laid tightly against the ceramic end plate of C_1, and connected with the shortest possible leads between the stator post and rotor lug is a 5-pf. ceramic fixed capacitor, C_2. It is out of sight behind the variable capacitor in Fig. 11-4.

Calibration

Pieces of white card cut to 2¼-inch diameter disks are mounted under the nuts holding the capacitors to the frames. With the capacitor full out or full-in, the knob is set on the shaft so that its indicator points straight up when the wavemeter is held in the left hand. A mark is made on the white card to indicate that this is one end of the tuning range. Swinging the capacitor around 180 degrees, another mark is made to indicate the other end.

A calibration accurate enough for most purposes can be made with a calibrated grid-dip meter. Known frequencies of transmitter or receiver stages may also be used. Calibration points throughout the range can be marked on one half of the capacitor range, giving calibration points at 48, 50, 72, 90, 100, 120 and 140 Mc. on Unit 1. The other half of the full circle of rotation can be marked for 50 to 54 and 144 to 148 Mc., the two amateur bands we're interested in.

The second unit was calibrated at 115, 125, 150, 175, 200, 220 and 250 Mc. on one half of the scale, and the 144- and 220-Mc. bands on

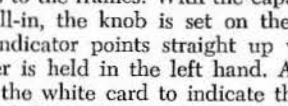

Fig. 11-5—An absorption-type wavemeter can be the simplest form of tuned circuit, as in L_1-C_1, at A. Circuit B has less capacitance in C_1, a fixed capacitor, C_2, in parallel, and a smaller value of inductance in L_2, for a band-spread effect.

the other. The bandspread version has nicely-spread points at 120, 125, 130, 140, 150 and 160 Mc. on one side and 144 to 148 on the other.

If care is used in duplicating the originals, you can copy the dial scales reproduced here, Fig. 11-6, and you'll be close enough for ordinary work. Direct calibration with a reliable frequency source is much better, of course. You very likely can borrow a grid-dip meter for this purpose, if you do not own one.

Using the Wavemeters

We have already described some of the uses for the instruments, but here are practical examples. Suppose you have a 50-Mc. crystal oscillator. You know it is oscillating, but you're not sure whether it is on the right overtone. Or perhaps you want to check a frequency multiplier to see if it is operating on the right harmonic. Connect a meter in the plate circuit, or in the grid circuit of the following stage, if any. Lacking a meter, use a lamp load or a neon indication, as in Fig. 11-1. Now, connect the wave-meter clip lead to a ground point in the equipment, and hold the wavemeter so that it couples to the tuned circuit in question. Rotate the wavemeter dial slowly, watching the meter or output indication. There will be a quick dip downward as the wavemeter tunes through the frequency at which the stage is operating. Hold the wavemeter as far as possible from the circuit being checked, and still get the dip indication. This will show the frequency of the r.f. power with the best accuracy.

Watch out for live circuits. Remember that even low voltages can be lethal. Large amounts of r.f. power absorbed by the wavemeter may cause r.f. burns on sensitive skin. Always be on the alert to keep hands and other parts of the body away from any circuits carrying a.c. or d.c. voltages. The wavemeter circuits can be enclosed in transparent plastic, if desired, but constant attention to the ever-present danger of high voltage is the only real protection.

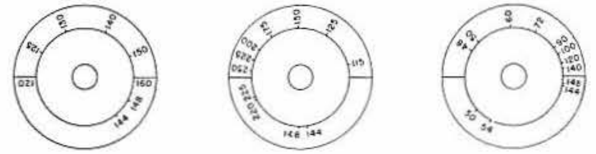

Fig. 11-6—These wavemeter scales may be copied, if construction is exactly like the original units. White cards are 2¼ inches in diameter. Calibration should be regarded as rough only, until checked against a source of known frequency accuracy.

SIMPLE FIELD-STRENGTH INDICATOR AND WAVEMETER

Some sort of field-strength indicator is a must for v.h.f. antenna work, and it can be useful in other ways. The instrument of Fig. 11-7 doubles as a wavemeter. In this role it is more sensitive and versatile than tuned-circuit absorption wavemeters just described. The diode CR_1 rectifies r.f. current, and the meter reads the resultant direct current. You can get visual indications of r.f. field strength and frequency without metering the circuits under examination.

With a pickup antenna connected to J_1, the meter will indicate relative power being radiated by a v.h.f. antenna, for use in evaluation or adjustment. Sensitivity depends on size of the pickup antenna, closeness of coupling between L_1 and L_2, type and condition of the diode, and range of the indicating meter. The lower the meter range the better. With a 100-μa. meter the instrument will give full-scale deflection at 100 feet or more from a v.h.f. beam, if a half-wave pickup antenna is used. With a 1-ma. meter the separation is considerably less, for a given transmitter power and antenna gain.

The tuning range of C_1-L_2 can be calibrated with a grid-dip meter or any other source of r.f. power of known frequency, or the scale shown can be copied with accuracy sufficient for most purposes, if the parts and layout of the original are duplicated. With values given the range is roughly 47 to 175 Mc., covering the 50- and 144-Mc. bands with some leeway on both ends of the scale. The various frequencies from 48 Mc. up, commonly encountered in v.h.f. transmitters, are also marked on the white

Fig. 11-7—A field-strength indicator and wavemeter for 50 and 144 Mc. The coil size is set so that the tuning range is approximately 47 to 175 Mc.

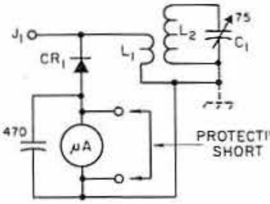

Fig. 11-8—Schematic diagram of the field-strength meter. Ground connection shown in broken lines is for metal-case construction.

C_1—75-pf. miniature variable (Hammarlund APC-75B).

CR_1—1N34 or other v.h.f. diode.

J_1—Insulated tip jack.

L_1—1 turn insulated hookup wire, ½-inch diameter, spaced 1 turn from L_2.

L_2—3 turns insulated hookup wire, ½-inch diameter, ¼ inch long. Adjust turn spacing for desired tuning range on C_1.

Meter is 100 micro-amperes, though up to 1 ma. is usable, with reduced sensitivity.

scale, making the meter handy in checking oscillator and mutliplier stages. Our brethren on lower bands may find this gadget useful for checking the frequency and source of v.h.f. parasitic oscillations or harmonics in their equipment for the "d.c. bands."

A small plastic meter box with a sloping front panel is used for a case. If a metal box is used, the instrument will have less hand-capacity effect than with a case of insulating material, but it will interfere with its use as a simple wavemeter. The ground connection shown in broken lines in Fig. 11-8 is for a metal-box version, and represents the grounding of the rotor of C_1. Parts may be arranged to suit the maker, so long as r.f. leads are kept short. Note the copper-strap short across the meter terminals. This should be kept in place whenever the plastic-box instrument is not in use, to prevent damage to the meter in strong r.f. fields.

Uses

A rough idea of the gain, pattern and front-to-back ratio of a v.h.f. beam can be obtained by setting the field-strength meter up at a distance out in front of the antenna under test and then performing substitutions or adjustments. No diode is linear, so changes in meter reading are only indications, not accurate measurements. A favored method of evaluation, with a device of this kind, is to measure the power input to the antenna that will give a constant reading as conditions are changed. If power in the line can be read accurately the power ratios thus obtained give a fair indication of the results of the antenna adjustments or changes. More on this in the chapter on antennas.

In using the instrument as a wavemeter a pickup antenna may or may not be required. Usually an r.f. indication can be obtained by holding the wavemeter near the circuit to be checked. Tune the capacitor slowly until a meter indication begins to show, then move the instrument away to keep the meter from going off scale. If the diode is in a strong enough r.f. field there will be rectified current regardless of the setting of C_1. In such circumstances it is advisable to use a pickup antenna a few inches long or more and move the instrument farther away, in order to prevent direct pickup by the diode. Varying the coupling between L_1 and L_2 is helpful in eliminating this blocking effect. Be sure to take all due precautions against high-voltage shock and r.f. burns in using this and other items of r.f. test equipment.

A sensitive microammeter is costly and fragile. Especially when working around high power, it is easy to run the 100-μa. meter off scale without realizing it. To prevent damage to the meter it is advisable to keep it shorted at all times when the instrument is not in use. A strip of copper about ¼ inch wide is kept wrapped around the meter terminals for this purpose, and is removed only when work is to be done.

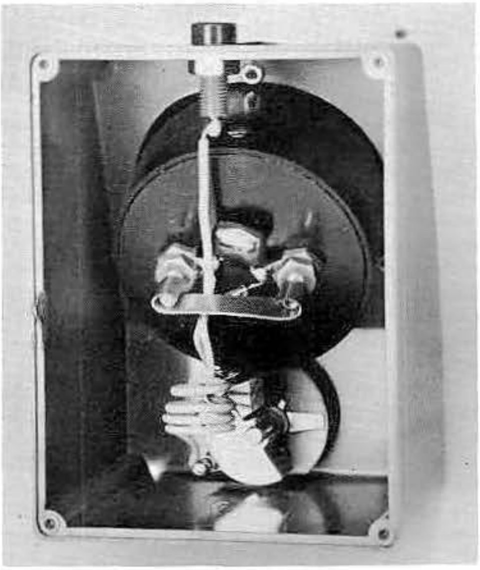

Fig. 11-9—Rear view of the field-strength meter. Note copper strap across meter terminals. This should be kept in place when the instrument is not in use, to protect the meter.

REMOTE-INDICATING FIELD-STRENGTH METER

A limitation on the usefulness of a simple field-strength meter of the type shown in Fig. 11-7 is that when it is in a good position for doing its job the instrument may be far from the man who wants to observe its fluctuations as he makes adjustments. The ideal spot for a field-

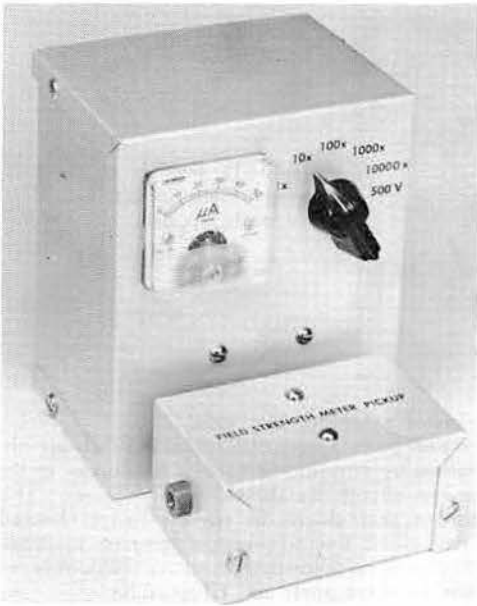

Fig. 11-10—A v.h.f. field-strength indicator and general-purpose test instrument for d.c. measurement. The small pickup unit can be plugged into the indicator as shown, or connected to it by a 2-wire cable of any desired length. Antenna rods or wires plug into jacks at each end of the small box. The meter can be used with test leads for measuring direct current and voltage. Ranges are 0 to 50 and 500 microamperes, 5, 50 and 500 milliamperes, and 0 to 500 volts d.c.

strength pickup device is as far out in front of the antenna being worked on as possible, and preferably in about the same plane. At this distance and height it must be observed by a separate worker, or the man making adjustments will have to use field glasses to read it, if he can see it at all.

The logical remedy is to make the pickup portion of a field-strength indicating device a separating unit, put it up where it will do the job required, and then run a line down to a meter at or near the adjusting position. The field-strength meter of Fig. 11-10 can be used with the meter box and pickup unit plugged together, as shown, or connected by means of a cable of any desired length. The meter box itself is made usable for other measuring jobs by building a switch and appropriate multiplier circuits into the case. As a general-purpose test meter it can read 0 to 50 or 0 to 500 microamperes, or 5, 50, or 500 milliamperes, and 0 to 500 volts, d.c. The multiplier circuit also allows different levels of sensitivity in the field-strength indicating role.

As may be seen from the lower portion of Fig. 11-11, the pickup unit, a 1⅝ × 2⅛ × 3¼-inch Minibox, contains only a diode, a resistor to complete the d.c. circuit, a bypass capacitor, two r.f. chokes, and two "banana jacks," one at each end of the box. These take plug-in wires or rods, which make up a half-wave dipole. Something shorter is usable, if maximum sensitivity is not important. There are no tuned circuits, so the system will work on any frequency. The only selectivity is the very slight amount introduced by the length of the pickup dipole.

On the back of the small case are two pin plugs. These take matching pin jacks on the end of the remote cable, or they can plug directly into the pin jacks on the 3 × 4 × 5-inch Minibox used for the meter case. The two-conductor connecting cable can be any available wire. Ordinary zip cord for home-appliance wiring is convenient. Coax or Twin-Lead that may have seen better days is also suitable.

Layout of parts in both units is generally uncritical, and obvious from the pictures. The lower the meter range used, the more sensitive the instrument becomes. A fairly inexpensive imported 50-microampere meter (AMD MRA-38) is shown. With this meter it is possible to get a usable indication of r.f. at up to several hundred feet away from a good beam, if a half-wave pickup antenna is plugged into J_3 and J_4. It is used with a 100-foot cord by the author.

There is really only one important rule in using the instrument: Be *sure* that you have the multiplier switch, S_1, set in such a position that

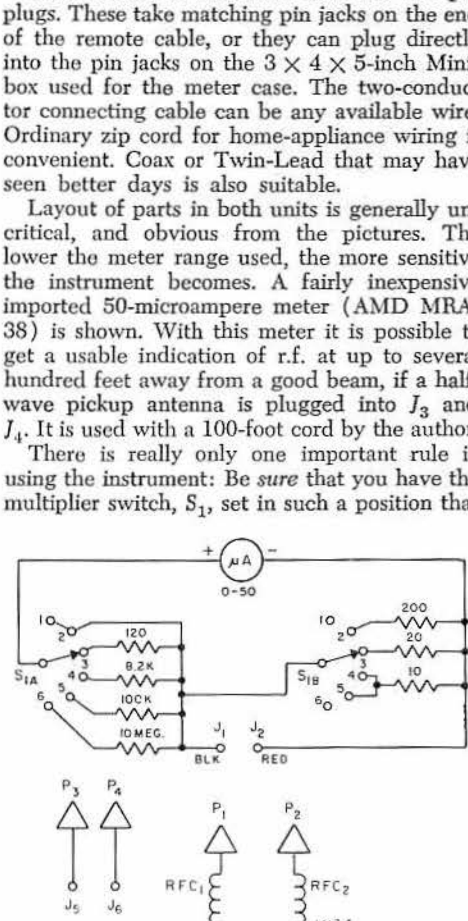

Fig. 11-11—Schematic diagram of the remote-indicating field-strength meter. Resistors in the indicator unit should be 5 percent or better for good meter accuracy.

CR_1—Any small r.f. diode, 1N34 or equivalent.

J_1, J_2—Insulated tip jack (Johnson 105-601 to 611, depending on color).

J_3, J_4—Insulated "banana jack" (Johnson 108-901 to 907).

J_5, J_6—Insulated tip jack and sleeve (Johnson 105-701 to 711).

P_1, P_2—Insulated tip plug, male (Johnson 105-601 to 611, with ⅝ inch No. 12 wire soldered in jack).

P_3, P_4—Insulated tip plug (Johnson 105-301 to 311).

RFC_1, RFC_2—1-watt resistor wound full length with No. 30 enamel.

S_1—2-section 6-position switch.

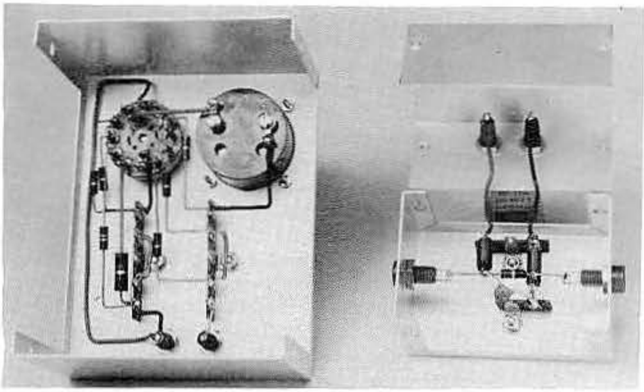

Fig. 11-12—Interiors of the field-strength meter pick-up unit, right, and meter box, left.

the meter movement will not be banged off-scale. When there is the slightest doubt about the magnitude of the reading to be expected, start on the highest current range and work down. This applies whether you are reading plate current in a transmitter or using the instrument as a field-strength meter. A 50-microampere meter can be burned out in a tiny fraction of a second. Play safe!

The meter ranges other than 0-50 μa. are obtained by introducing various resistances in the meter circuit, as shown in Fig. 11-11. This means that shunts in circuits being checked may affect the accuracy of current readings. For more on field-strength-meter techniques see the previous item, and Chapter 8.

A V.H.F. MAN'S MONIMATCH

Many descriptions of v.h.f. antennas mention the desirability of using an s.w.r. bridge in making matching adjustments, yet most bridges do not work satisfactorily in the v.h.f. range. The handy gadget shown in Fig. 11-13 is a v.h.f. version of the popular Monimatch. It will work to some extent on lower frequencies, but it is primarily intended for use at 50 to 450 Mc., at power-output levels from 5 to 100 watts on 50 Mc., and as low as 1 watt on higher bands. It also doubles as a test meter, for use in tuning up the "Two-Band Station" described elsewhere in this book.

The limitation on frequency range is due to the r.f. pickup from the modified coaxial line section that is a fundamental part of this type of instrument. If the loop W_1 of Fig. 11-14 is made long enough to work at, say, 3.5 Mc. the coupling will be so great at 144 Mc. that no null can be achieved in the reflected-power position. Conversely, a short loop like that needed in a v.h.f. version gives so little pickup on lower bands that the forward-power reading is unsatisfactorily low. To develop satisfactory operation in the v.h.f. range also requires special attention to lead lengths and r.f. bypassing.

The bridge can be left permanently in the line from the antenna to the coaxial relay, and it will show relative power output (forward power) as well as reflected power, depending on the switch position, at levels up to 100 watts or so safely. The absorbed power is a negligible portion of the transmitter output.

The meter is a 1¾-inch square plastic-face 1-ma. job (Lafayette TM-400). Connected as shown in Fig. 11-14, it not only serves as an indicator for the bridge, but it also may be used for measuring plate and grid current in the transmitters of the complete v.h.f. station described in Chapters 6 and 7. Resistor shunts are built into the transmitters, so that when the meter with its 1000-ohm resistor in series is plugged into the proper tip jacks the meter reads 10 ma. full scale for grid current measurements and 100 ma. for plate current.

The meter reading when the bridge is in the antenna line indicates relative power only. The sensitivity control, R_1, permits use of the bridge

Fig. 11-13—The s.w.r. bridge is a v.h.f. version of the Monimatch, commonly used on lower bands. Test leads permit the meter to be used for measuring purposes in the transmitters of the v.h.f. station, Chapters 6 and 7.

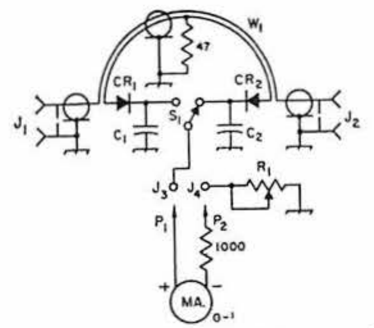

Fig. 11-14—Schematic diagram of the s.w.r. bridge. Fixed resistors are ½ watt.

C_1, C_2—0.001-uf. button-mica (Erie CB21PD501K).

CR_1, CR_2—1N34 diode.

J_1, J_2—Coaxial fitting to suit equipment used. SO-239 in photo.

J_3, J_4—Insulated tip jack.

P_1, P_2—Insulated tip plug.

R_1—5000-ohm control.

S_1—S.p.d.t. rotary switch (Centralab 1460).

W_1—7¼-inch length of RG-58/U, with No. 24 enamel wire inserted as per text. Formvar insulated wire preferred.

at power levels from 1 to 100 watts. It should always be turned down before the meter is used at an unknown power level. The control is then advanced to give a reading that is convenient for the adjustment purpose at hand. In tuning up for maximum power output you may want to set the meter at about half scale, to allow room for improvement. If tuning is completed and you are checking antenna matching, the forward reading should be as near full scale as possible, for maximum sensitivity in the reflected-power position.

The bridge is built in a $2 \times 4 \times 6$-inch aluminum chassis. The input and output coaxial fittings are mounted in the exact centers of the long sides of the chassis. The two button bypass capacitors are 1¾ inches apart, also on the center line of the chassis. Placement of the other components is not critical.

To make the line for the bridge, cut a piece of RG-58/U coax 7¼ inches long, and remove the black covering. Push the braid from the ends toward the center, so that it becomes loose over the inner insulation. At the exact midpoint of the braid, part the strands sufficiently to pass a No. 24 enameled wire. This should be about 10 inches long, preferably Formvar insulated. Clean the insulation from it for about ½ inch at the center, and twist this portion into a small loop. This will be the connection point for the 47-ohm resistor. Now feed the ends into the space in the braid, and bring them out through the opposite ends, pulling them through the braid at about ½ inch from each end. Solder the ends of the inner conductor of the coax to two coaxial fittings. Slide the braid back to its original position and solder the braid ends to grounding lugs at each fitting.

Solder the 1N34 diodes to the outside switch terminals, leaving connections no more than about ¼ inch long. If you have the newer type diode, which is glass-enclosed and color-coded, the end with the black ring should go toward the switch. The other ends of the diodes connect to the ends of the wire that is threaded inside the braid. Make these connections short and direct, and be sure that the exposed leads are the same length on each side. The coax is draped in U shape, so that it just touches the inner end surface of the chassis. At this point the 47-ohm resistor is connected from the enameled-wire tap to a ground lug fastened at the center of the end wall of the chassis.

Placement of the other components is not critical. The sensitivity control, R_1, is mounted in the top end of the chassis, and the meter hole is centered below it, 1¾ inches down from the top edge of the main chassis surface. The two tip jacks are ¾ inch in from the edges of the chassis and 2½ inches down from the top. The switch is 2 inches up from the bottom.

Uses

The primary purpose of the bridge is to determine when the antenna system is properly matched to its feed line, but it also serves other ends. When in the forward-power position, the meter gives a relative indication of the amount of power going through the transmission line, so

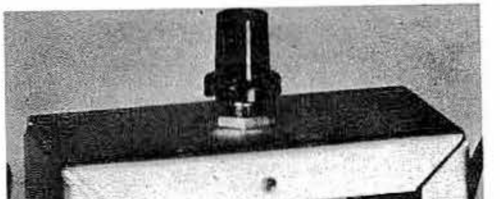

Fig. 11-15—Interior of the s.w.r. bridge. Symmetry and minimum lead length are important, for good results above 50 Mc.

it is useful in tuning up the transmitter. In fact, once the transmitter is operating according to the information given in Chapter 6, tuning can be done merely by watching the bridge meter while adjusting the final-amplifier plate and loading capacitors for maximum indication.

The bridge should be connected between the antenna relay and the line to the antenna, as shown in Fig. 7-2. V.h.f. antennas are usually designed to be fed with 50-ohm coaxial line, or 300-ohm balanced line. Various means for making the antenna present a 50-ohm load are described in Chapters 8 and 9, but whatever the matching system is, it can be adjusted by setting the bridge switch in the reflected-power position and adjusting the antenna matching for minimum indication. Switch to the forward-power position intermittently and check the transmitter adjustments to see that they have not been thrown off by the change in load impedance occurring during antenna work.

If the antenna is fed with a balanced line, some form of balanced-to-unbalanced coupling or balun will be needed. See "Baluns," Chapter 8. It should be emphasized that an antenna coupler or antenna matching device should always be adjusted for minimum reflected power in the coaxial line. This should be zero or very close to it. The bridge is then switched to the forward-power position and the transmitter and loading adjustments are checked to be sure that the rig is delivering maximum power to the line. Adjustments to the transmitter have no effect on the standing-wave ratio on the transmission line. If the transmission line is long (more than a few wavelengths) the bridge will give the most sensitive indication of matching adjustment if it is connected at or near the antenna. Where it is connected in the line will have no bearing on its effectiveness as a forward-power indicator, but a line that is a half wavelength or multiple thereof is desirable, as it will repeat the antenna impedance at the bridge.

The test leads must be plugged into the bridge tip jacks in order to read either forward or reflected power. These leads can be any convenient length as they carry only a very small direct current. The bridge may be left connected in the antenna line while the meter is used for transmitter measurements, as removing the leads from the bridge tip jacks disconnects the meter from the bridge circuitry.

A V.H.F. IMPEDANCE BRIDGE

It is often helpful to be able to measure impedances of input and output circuits of v.h.f. converters, transmitters, cavity filters, dummy loads and antennas. Most impedance-matching devices described in amateur literature are ineffective above about 30 Mc., because of inaccuracies resulting from excessive internal capacitance and inductance. The bridge of Figs. 11-16 through 18 was tailored to the needs of the v.h.f. operator. Minimum lead inductance and components chosen for v.h.f. qualities permit reasonable accuracy in the v.h.f. range.

An s.w.r. bridge can be used to secure a proper match between circuits, or between a circuit and its load, but this does not allow measurement of the terminal impedance when a mismatch is present. This impedance bridge will enable the user to make direct readings of impedance in the 10-to-500-ohm range, thus permitting the solution of a variety of matching problems.

Construction

It is suggested that the builder duplicate, as nearly as possible, the physical layout. Flashing copper ¼ inch wide is used for leads in the bridge portion of the circuit, to keep down lead inductance. A shield of copper, brass or aluminum divides the metering and bridge circuits. The potentiometer R_3 is a 2-watt carbon control, with linear taper, so that the resistance scale will not be cramped at one end. It is mounted on an insulating plate, rather than on the metal box, to reduce capacitance between the chassis and the metal shield on the control. The insulated mounting plate is then fastened to the Minibox, with the control bushing centered in a ⅜-inch hole.

The bridge resistor R_1 is a 5-percent 1-watt carbon. It is made of two 1000-ohm ½-watt resistors in parallel, in the example shown. R_2 should be 50 ohms, a value not readily obtainable. This is within the possible range of 47-ohm 10-percent tolerance resistors, so a stock of

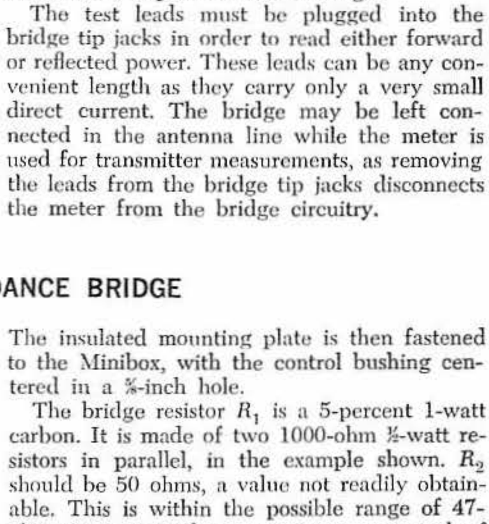

Fig. 11-16—Simple v.h.f. impedance bridge. Settings of the variable control, marked on the front of the case, represent values of impedance corresponding to nulls in meter indication.

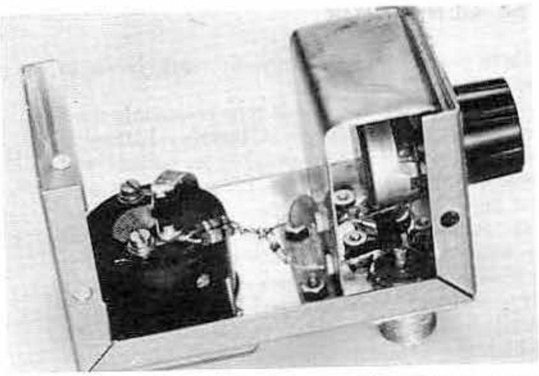

Fig. 11-17—Interior of the impedance bridge. Inside the shield, right, are the coaxial terminals, the variable control, R_3, and the load resistor, R_2. Feedthrough insulators may be improvised from ¼-inch teflon rod. Leads thereto are ¼-inch-wide copper strips. Load resistor R_2 is connected directly between the terminals of J_1 and J_2.

these can be checked for the nearest to the desired 50 ohms. The value of a resistor can also be raised by filing into the carbon element. Be sure that the ohmmeter used for checking the resistance value is reliable, if accuracy is desired.

The jacks J_1 and J_2 are mounted as closely together as their flanges permit. The meter is a 100-microampere type, for maximum sensitivity when using the bridge with r.f. sources of low power output. It is an inexpensive 1⅝-inch square imported model (Calrad).

Checking and Calibration

A pointer knob on R_3 and a paper dial scale pasted to the end of the Minibox provide for calibration. Select a number of 1-watt carbon resistors of values between 10 and 500 ohms, to be used as calibration loads for the bridge. Attach the link L_1 to the input jack J_1. Couple a grid-dip meter or some other low-power source of r.f. power to the link. Use a frequency around 145 Mc., if best accuracy is wanted in the 2-meter band.

Attach a low-value resistor to J_2 with the shortest possible leads. Adjust the coupling between the r.f. source and L_1 for full-scale reading on the bridge meter, or as high a reading as possible if less than full-scale. Next, adjust R_3 for a null in the bridge meter reading, making certain that the null is as deep as possible. Mark the dial scale, and jot down the value of the load resistor used. Repeat this process with various resistor values until a complete set of calibration points has been obtained. A new scale can now be made and marked permanently with India ink, indicating the resistance values at the check points.

Uses

The more nearly a load is to being purely resistive, the deeper the null will be. As a load becomes reactive, as frequently happens in r.f. measurements, the null is less pronounced. Even though the null is poor, the readings remain useful, as they approximate the actual value of the impedance under measurement.

V.h.f. matching networks may be checked by presetting the bridge to the desired value of resistance, and then making adjustments to the matching network until the null is obtained on the bridge meter. Adjust carefully for best null. A Gamma matching system can be adjusted without a transmitter in this way. Insert a length of coax, of the impedance of the line to be used, between the antenna and the bridge. This line should be a multiple of a half wavelength long electrically, so that the antenna impedance will be repeated at the output terminal. This cable permits the operator to take readings without being in the immediate field of the antenna. A 2-wavelength section should be about right for work in the 144-Mc. band. If the coax is solid-dielectric (not foam) RG-8 or similar, a test cable of 107 inches will do.

Other devices such as the coaxial-line filters of Chapter 12 may be checked for input and output impedance by using line sections that are multiples of a half wavelength, and a dummy load of the correct impedance. The filter input or output circuit can then be adjusted for a null at the desired impedance. Transmitter or converter input or output circuits can be adjusted in a similar manner. Unknown values of impedance can be determined by attaching the bridge to the circuit being tested, and then sweeping across the bridge range for a null.

Though a grid-dip meter is mentioned as a power source, a low-powered transmitter can, of course, be used in these various applications. The frequency stability of the transmitter is an advantage, but be sure that it has output only in the desired frequency range. Use of a coaxial filter in the transmitter output may be desirable, in order to prevent harmonics and subharmonics from reaching the bridge.

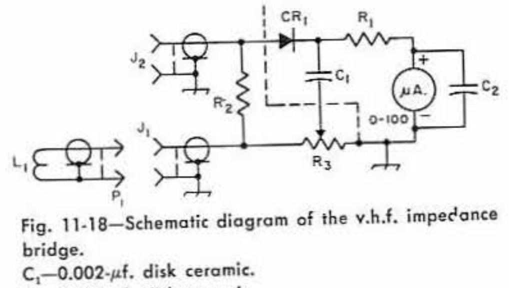

Fig. 11-18—Schematic diagram of the v.h.f. impedance bridge.

C_1—0.002-μf. disk ceramic.

C_2—0.001-μf. disk ceramic.

CR_1—1N82A diode. 1N34 also usable.

J_1, J_2—Coaxial connector, SO-239.

L_1—1 turn No. 12 enamel, 1-inch diameter.

P_1—Coaxial plug, PL-259.

R_1—500-ohm carbon, 1 watt.

R_2—50-ohm carbon, 1 watt.

R_3—500-ohm 2-watt control, linear taper (Allen Bradley). Case is 2¼ by 2¼ by 4-inch Minibox.

SILICON DIODE NOISE GENERATOR

One of the most useful tools in adjusting v.h.f. receivers is a noise generator. In its simplest form a noise generator is a diode drawing current, and therefore making noise. This noise extends all across the r.f. spectrum, up to a frequency determined mainly by the circuitry and the diode used.

Such a crystal-diode noise generator is shown in Figs. 11-19 through 21. Noise figure cannot be "measured" with a device of this kind, but it is handy as a noise source for adjusting a v.h.f. receiver for best noise figure. The lower the diode current for a given margin of diode noise over receiver noise, the better the receiver is working.

Construction

The noise generator is built in an aluminum Minibox 3¼ by 2⅛ by 1⅝ inches in size (Bud CU-3001-A). Only the load resistor, R_1, the diode, CR_1, and the bypass capacitor, C_1, are critical as to mounting position. These should be connected with absolutely minimum leads, if the generator is to be useful above 100 Mc. or so. R_1 is inside the adapter sleeve of P_1. C_1 should be a button-mica or other capacitor having good u.h.f. characteristics. Ordinary disk ceramics are not suitable above 50 Mc.

The coaxial plug should match the connectors on the receiving equipment with which the generator is to be most often used. The PL-259 plug with UG-176/U adapter was used here. The flange on the adapter is only very slightly larger in diameter than the threaded portion, so using it as a means of clamping the assembly to the generator case is not very satisfactory. To give more binding surface, washers of flashing copper were made for both sides of the mounting hole. One is shown in Fig. 11-21. Cut with shears from one edge to the washer hole, and bend the washer at the break slightly, so that it can be threaded onto the adapter sleeve. Use one of these washers on each side of the box, which must be drilled for a ⅝-inch hole to pass the threaded portion of the adapter. One washer can be soldered to the adapter flange and the other to the end of the plug sleeve, to make the whole assembly less likely to work loose in using the noise generator.

The battery is the 9-volt type commonly used in small transistor radios. A mercury battery is worth the difference in price as its voltage will remain practically constant throughout its useful life. This makes possible a reasonably accurate calibration of the generator's noise output in terms of the setting of the series control, R_2.

Connection to the battery is made by means of a terminal block removed from the top of a dead battery. The fiber insulating plate on which the terminals are mounted is fastened to the front wall of the box. The smaller of its terminals is grounded to the case, and the other is backed up by two layers of plastic insulating tape. The assembly thus acts as both connector and mounting plate for the battery.

The large end of the diode is held in a plate clip of the type used for metal tubes. The smaller end contact was removed from an old octal wafer socket. Many different diodes are usable if the upper limit of frequency is not important. Silicon rectifier diodes of the kind used in power-supply work can be used at 50 or 144 Mc., but they draw considerable current, and do not work well at higher frequencies. U.h.f. mixer diodes of the 1N21 series are recommended. Also used with good results: 1N25 and 1N32. Germanium diodes such as the 1N34 are not satisfactory. A good diode will give plenty of noise with no more than about 2 ma. diode current. You may want to put a fixed resistor in series with R_2 to keep the current below this point, if all your work is going to be with receiver front ends known to be in quite good working order.

The tip jacks, J_1 and J_2, are for measuring current through the diode. This will run about 10 ma. maximum with the 1N21 series diodes, but may be higher with other types. The setting of R_2 is not meaningful when batteries other than the mercury type are used, but the diode current is directly related to noise output, and can be calibrated roughly in noise figure.

The load resistor, R_1, should be a value equal to the line impedance of the antenna system to be used. If a 51-ohm resistor is not available, 47

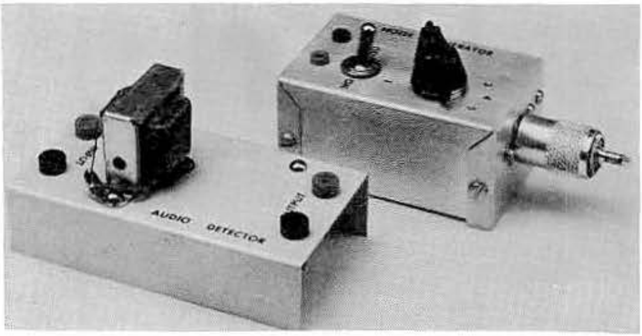

Fig. 11-19—Silicon-diode noise generator for v.h.f. receiver testing, with its audio detector, left, for smoothing out noise readings. Tip jacks in the noise generator permit taking diode current readings.

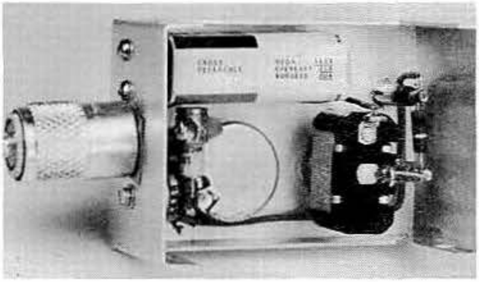

Fig. 11-20—Interior of the diode noise generator. The diode, load resistor and bypass capacitor should be connected with the shortest possible leads.

ohms is close enough for ordinary purposes in work with 52-ohm antenna systems. Use a 75-ohm resistor if the line is 72 ohms. A 68-ohm load may be used if both 52- and 72-ohm lines are to be encountered. The value is not particularly critical, as the noise generator is not a precise test instrument.

Using the Generator

In receiver work the generator is best connected directly to the receiver or converter antenna jack. Run the receiver with its a.v.c. off, if possible. Turn up the r.f. and audio gain controls until receiver noise is heard. Keep the r.f. gain control as low as possible, in order to prevent overload, except where the gain of the first stage in the receiving system is affected by the gain control setting. This is rare in v.h.f. receiving setups, but it may be found in lower-band gear. Note the level of the receiver noise, by ear or by connecting an a.c. voltmeter, db. meter, or the audio detector described later, across the speaker or earphone terminals. Now turn on the noise generator, starting with R_2 at its maximum resistance setting. If no increase in noise is heard, reduce the resistance slowly until noise begins to rise.

Use a noise increase that you can remember or measure. The lower the crystal current required to give this noise increase, the better the

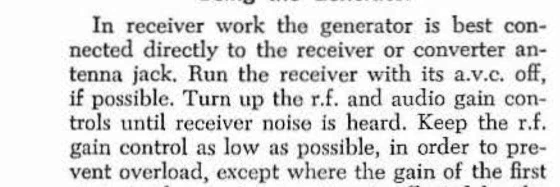

Fig. 11-21—Schematic diagram of the noise generator. Two washers, right, are used to mount the coaxial plug to the case.

C_1—500-pf. button-mica.

CR_1—Silicon mixer diode, 1N21 etc.

J_1, J_2—Insulated tip jack.

P_1—Coaxial plug, PL-259, with UG-176 adapter.

S_1—Toggle or pushbutton switch.

R_1—51-ohm ½-watt carbon, mounted inside adapter sleeve of P_1.

R_2—50.000-ohm control.

receiver. This gives a rough comparison of one v.h.f. converter to another, provided that they are for the same band. Preferably, the converters so compared should be used with the same receiver. Adjustment of antenna coupling, oscillator injection, tests on various tubes or transistors, or checking any other factor affecting receiver performance, can be done with a noise generator of this type. Line loss in a length of coaxial line can also be measured, by connecting the line between the noise generator and the converter, noting the difference in noise through the line, and with direct connection of the generator.

In receivers having no provision for removing a.v.c., some other method of measuring noise output must be made. The receiver S meter can be used, if it responds to the receiver's noise level with the generator turned off. If it doesn't, it may be necessary to run the noise-generator output higher than normally would be the case, in order to get a meter rise indication on generator noise.

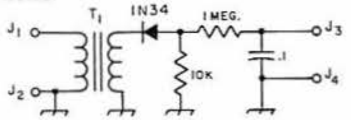

Fig. 11-22—Schematic diagram of the average-type audio detector shown in Fig. 11-20. Parts arrangement is uncritical.

J_1, J_2, J_3, J_4—Tip jack.

T_1—Small audio output transformer. Low-impedance winding connects to J_1, J_2.

Any receiver adjustment that makes it possible to obtain a given noise increase with a lower diode current, or a greater S-meter reading increase with the same level of diode current, is an improvement.

As shown, the noise generator produces in excess of 20 db. of noise at 50 and 144 Mc., and enough to be usable with any fairly good receiver at 220 and 420 Mc. The amount of noise and the upper useful frequency limit depend on the diode used, and on its condition. Avoid subjecting the diode to strong r.f. fields, or to excessive current. If you buy a good diode, it will probably come encased in metal foil, or otherwise shielded. Keep it that way until it is installed in the noise generator, and then use the lowest diode current that will give satisfactory noise output.

A refinement some users like is to substitute a pushbutton switch for the toggle type for S_1. If this is the microswitch type it can be closed with light finger pressure, making it easy to take readings without the likelihood of disturbing the setting of the diode-current control, R_2.

Audio Detector for Noise-Generator Work

In using a vacuum-tube voltmeter or other a.c. output meter in noise-generator work the erratic nature of the meter indication is often a problem. Noise is random in nature, and unless

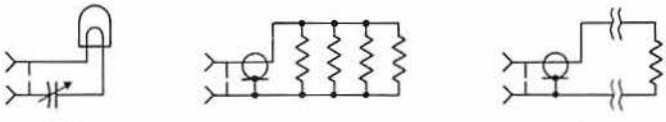

Fig. 11-23—Three dummy loads for v.h.f. use. Lamp in A has variable capacitor in series to tune out reactance. See text. Load B has carbon resistors in parallel. Total resistance should equal impedance the transmitter is designed to work into. A considerable length of lossy coax, C, makes an excellent r.f. load. Since the line dissipates most of the power, the load resistor need not be of high wattage rating.

the meter is highly damped, the needle will fluctuate constantly, making it difficult to establish a reference. The device shown with the silicon-diode noise generator in Fig. 11-19 is a simple audio detector to smooth out such meter readings.

Originally described by KØDJP in *QST*,[4] it is merely connected to the speaker or earphone terminals of the receiver, and the meter is then connected to its output terminals. The detector is shown schematically in Fig. 11-22. Use is the same as if the meter were connected to the receiver directly. The above reference is well worth reading by anyone interested in improving receiver performance.

Though the audio detector is shown with the crystal-diode noise generator, it is helpful with other types of noise generators, wherever the flickering of the meter indication may be troublesome.

DUMMY LOADS

In order to test a transmitter legally, an amateur must use a dummy load. Most of us put our rigs on the air for test purposes for brief periods, but the considerable running of a transmitter that is usually required during construction and trouble-shooting should not be done with the transmitter feeding an antenna. When an antenna test is unavoidable, the operator should be sure that the frequency to be occupied is not being used for communication at the time. If an appreciable transmitter-on time is contemplated, a nonradiating load is the only considerate (and legal) approach to testing.

Once it was the usual thing to hook a lamp of suitable wattage across the output terminals. This may still suffice, but it leaves much to be desired in most instances. Few lamps or combinations thereof come anywhere near to being 50-ohm loads, in the v.h.f. range, and consequently they may be all but useless for any meaningful testing. Exceptions are some small pilot lamps (within their limited power-handling capabilities), and incandescent lamps of around 100-watt rating. Several blue-bead pilot lamps in parallel may make a fair v.h.f. load, and some 100-watt lamps singly or in parallel are usable at frequencies in the v.h.f. range.

Lamps in between these power levels are highly reactive, and the impedances they represent vary greatly with the power being dissipated. Low-wattage incandescent lamps, 15 to 40 watts, are particularly poor. Lamps larger than the 100-watt size nearly always develop "hot spots" in a portion of their filaments, making them unreliable as power indicators. They also tend to go gassy and burn out before reaching their normal power level. Lamps of intermediate wattage can be improved somewhat as r.f. loads by connecting a variable capacitor in series to tune out the reactance they and their leads represent. See Fig. 11-23A.

Put this combination on the output of your s.w.r. bridge, and feed in some r.f. power. Tune the capacitor for lowest reflected power, not maximum lamp brilliance. Readjust transmitter output coupling for maximum output, after the load is matched as well as you can get it. A maximum value of 100 pf. should serve for 50 or 144-Mc. loads. Use 50 pf. if nothing lower than 144 Mc. is to be used, and 25 pf. is enough for higher bands, if the combination will work at all.

The resistive load, B, is much better, within its power capabilities. Two such dummy loads using paralleled resistors are shown in Fig. 11-24. Three considerations are important here: the resistors must be the composition (carbon) type, the inductance of the leads must be kept to an absolute minimum, and the power to be dissipated should be kept below the rated total wattage of the resistors used, except for brief tests.

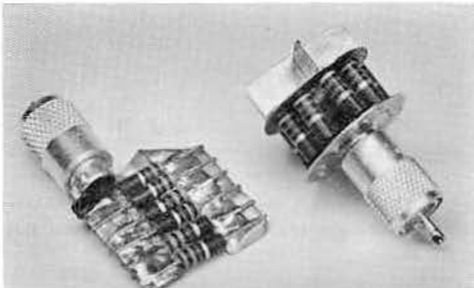

Fig. 11-24—Two v.h.f. dummy loads of the type shown schematically in Fig. 11-23B. Copper fins on unit at right aid in dissipating heat, and disks provide low-inductance parallel connectors.

Any number of resistors can be connected in parallel, so long as they are all the same wattage and the same resistance. Six 330-ohm 1-watt carbon resistors are paralleled by soldering their ends to straps of flashing copper in the unit at the left. One strap solders to the coaxial fitting sleeve, and the other is narrowed down to fit inside the sleeve and make contact to the pin. The portion inside the fitting is covered with insulating spaghetti or plastic tape, to prevent shorts.

A somewhat better load, where higher dissipation is wanted, is seen at the right. Here disks of flashing copper are used, and the resistors mounted in a circle, to keep inductance down. Nine 470-ohm resistors are used in the unit pictured, but thirteen 680-ohm or nineteen 1000-ohm resistors would probably do equally well.

These are not perfect loads, but they are much better than lamps. They show no greater than 1.2:1 s.w.r. at 50, 144 or 220 Mc., with the best match (close to 1:1) at 50 Mc. Don't take the wattage ratings of the resistors too literally. They get warm at a dissipation of 1 watt each, and when they get more than just warm to the touch, the resistance value may begin to change. This is not too much of a problem ordinarily, as the power is usually on for only a fraction of a minute at a time, just long enough to take a reading or make a quick adjustment. The cooling fins on the unit at the right help to keep the heat flowing out of the resistors, and it will dissipate more than 10 watts safely for brief periods; no more than a few seconds at a time.

The load at C is the best of all. In fact, the principle is used in some of the best r.f. wattmeters and dummy loads made. If you've got 100 feet of coax that is too loosy for use on an antenna, don't throw it away; it is an ideal dummy load. Looking back to Table 8-III, we see that 100 feet of RG-58/U (even in new condition!) has a loss of 6 db. per 100 feet at 144 Mc. This means that you can put the circular load of Fig. 11-24 on the end of it, feed 40 watts into the other end, and the resistors will just reach their rated dissipation. At 420 Mc. you could run 60 watts into the line and it wouldn't hurt the load. You can even short the end of the line, or leave it open, and it will make hardly any difference in the s.w.r. reading or the power-handling capability.

A load of this kind is a good match at any frequency where the loss is 6 db. or more, and its power-handling capability is considerable, for short test periods. If the coax can be coiled loosely and subjected to a cooling air blast, it can be made to take just about any amateur power for short periods.

LECHER WIRES

Here is a measuring instrument that is almost as old as radio communication, but it is still a handy item for the v.h.f. or u.h.f. experimenter. The length of an electromagnetic wave can be measured directly on a transmission line, by observing the distance between points of maximum or minimum r.f. voltage. Lecher Wires are a means of doing this reliably and accurately. If you are accustomed to using metric scales you can read wavelength directly, the distance between the voltage peaks or nodes being a half wavelength at the frequency being measured. (A meter is 39.37 inches, which as every v.h.f. man knows is a half wavelength at 150 Mc.)

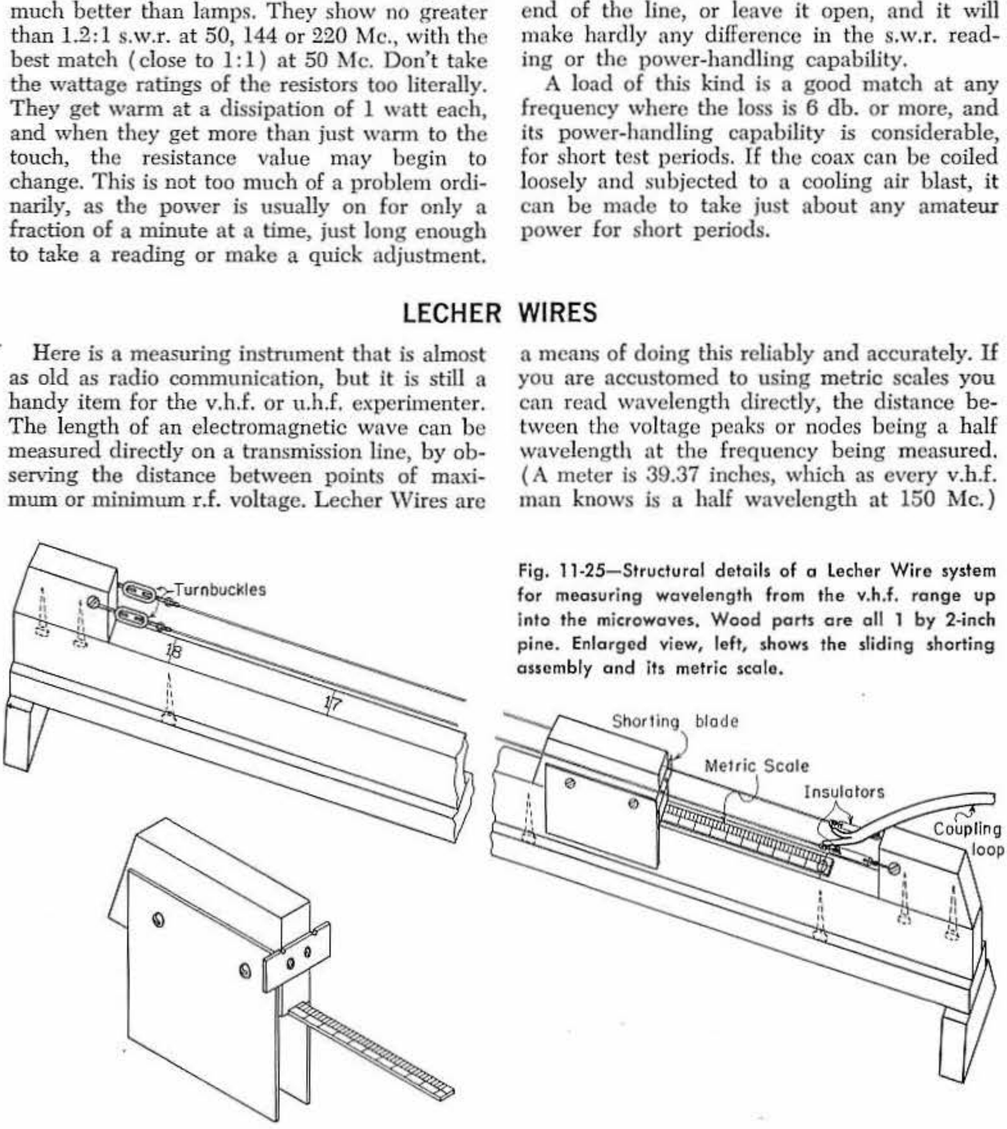

Fig. 11-25—Structural details of a Lecher Wire system for measuring wavelength from the v.h.f. range up into the microwaves. Wood parts are all 1 by 2-inch pine. Enlarged view, left, shows the sliding shorting assembly and its metric scale.

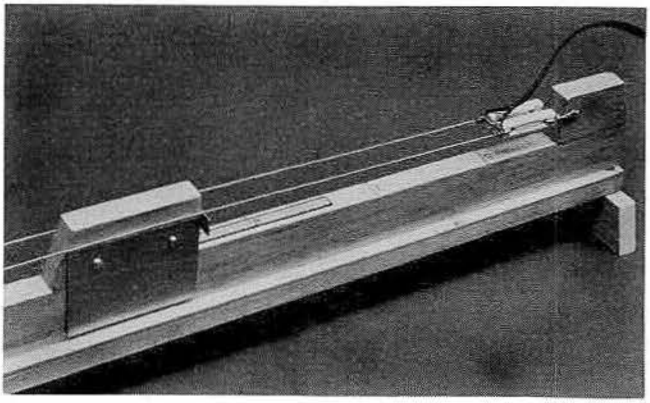

Fig. 11-26—Close-up view of the coupling end of the Lecher Wire assembly. Overall length depends on how low the user wishes to be able to go in frequency.

Construction

The wires in the portion of the instrument to be used for measurement must be without insulating material in direct contact. Provision must be made for holding the wires taut, and in uniform spacing. The shorting device must make firm contact, and the distance between peaks (or nulls) must be measured precisely, if accurate measuring is to be undertaken. These objectives can be met easily and inexpensively in numerous ways, one of which is shown in our practical example, Figs. 11-25 and 26.

The construction requires little explanation, and dimensions are not critical. The base of the assembly is made from two straight pieces of 1 by 2-inch pine, fastened together in a T-shaped cross-section, and supported on two blocks of wood. The anchors for the measuring line are of similar material. The wires are held tight with turnbuckles at the left end, and are supported on insulators at the right end.

How long you make the line depends on the lowest frequency you want to measure. The model shown is 7 feet long, which will take care of measurement from the 144-Mc. band up well into the microwave region. If you want to start at the 220-Mc. band an overall length of about four feet will suffice.

A rough Lecher-Wire measurement of wavelength can be made by running a knife or screwdriver blade along any bare-wire transmission line, but if you want to measure accurately something like the shorting blade and carrier shown here must be incorporated, to give repeatable results. The block rides on the base strip, with two metal side plates keeping it in alignment. These plates need not be metal, but it is convenient that way. At the right end of the travelling block is a notched metal short. The wires are kept tight enough so that a good electrical contact is made by this plate.

The base is marked off in tenths of a meter, beginning at a point directly under the coupling end of the line. The travelling short has a transparent metric scale (most stationery stores have them) fastened to its underside, so that readings can be taken directly in metric units of length. (Inches and feet don't enter into this at all.)

Now we have to couple to the r.f. source in some way. The propagation factor of the coupling line is of no importance, so it can be Twin-Lead or anything in the way of a balanced line that may be handy. From here on the operation is much like that with an absorption-type wavemeter, except that the Lecher Wires are much more accurate if made and used properly.

Measuring Wavelength and Frequency

The energy source can have any of several indicators: a grid current meter, plate current meter, r.f. voltmeter, field-strength meter, lamp load or whatever. The coupling loop of Twin-Lead, at the right end of the Lecher Wires, is

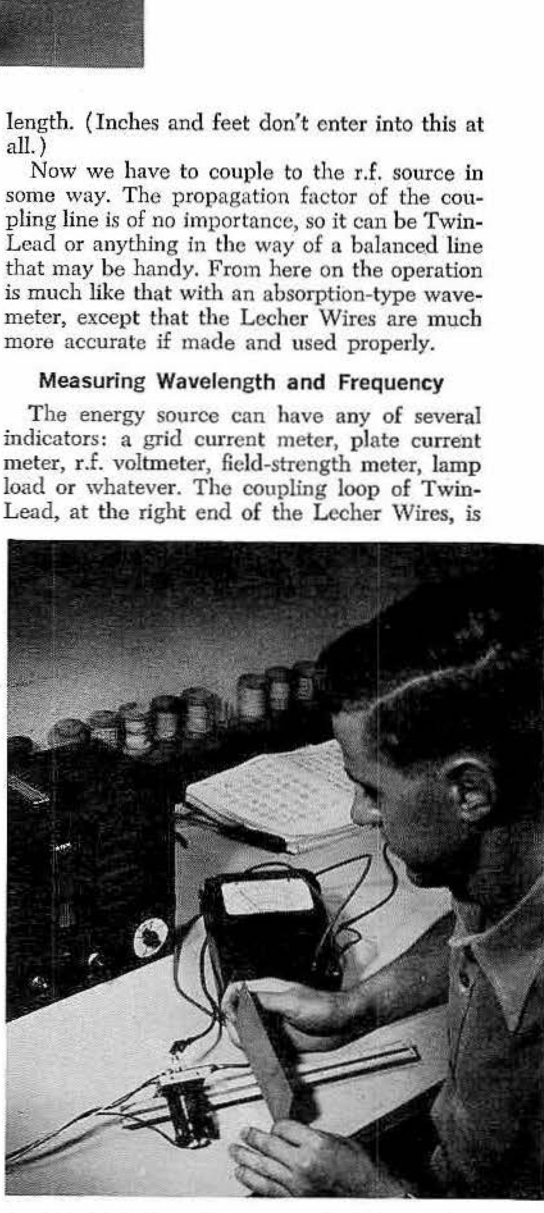

Fig. 11-27—W1QVF demonstrates "wireless Lecher Wires."

shorted at the end. This loop is placed near the r.f. circuit so as to couple some energy from it. The loosest coupling that will work is the best.

Now, run the sliding short along the Lecher Wires, watching for a change in the indicator, whatever it may be. When the change occurs, note the reading on the base and block scale. Let's say it's 0.255 meter. Now move the carrier along until a second dip is found, and note the scale reading. Suppose it is 0.937. Subtract the first reading from the second, the answer in our example being 0.682 meter. This is a half wavelength at the frequency being measured. To convert this to frequency in megacycles, divide 150 by the wavelength just measured. Our answer shows that we have just missed the 220-Mc. band, and we're on 219.9 Mc.

"Wireless Lecher Wires"

The principle of measuring between wave peaks or troughs can be applied in countless ways to estimate or measure frequency. One that is used for rough checks in the microwave region is shown in Fig. 11-27. W1QVF demonstrates the technique with a 723A/B klystron oscillator, in or near the 10,000-Mc. band.

A crystal diode, in this case a 1N34, is connected to a microammeter, and mounted close to the output probe on the klystron. A metal plate a few inches across is then moved toward or away from the klystron, and the distance between current peaks or nulls is measured. This can be done with a metric scale. If you are an incorrigible inches-and-feet man, measure in inches and then convert to megacycles by dividing 5905 by the distance between indications in inches.

Either system can be carried to whatever accuracy you desire, and with a little care it can be as good as it needs to be for amateur purposes.

OTHER TEST EQUIPMENT

The items described in detail are specialized v.h.f. gear. In addition, there are many pieces of equipment that are useful in any amateur's station, regardless of the frequencies he intends to use. They will not be dealt with in any detail here, because the information needed regarding them is available in many other places. We will confine our discussion of them to points of special concern to the v.h.f. enthusiast.

Panel Meters. You can never have too many of these. Most commercial transmitters are under-metered, probably because good meters run up costs. When building gear ourselves, we will do well to provide for metering in every circuit that is likely to require adjustment. One simple way to do this is to put in feedthrough bypass capacitors that can serve as test points. This is shown in the 144-Mc. heterodyne unit, Fig. 6-30. Generous use of tip jacks is another good method. This is shown in the 50- and 144-Mc. transmitters of Figs. 6-8 and 6-10.

Temporary metering is usually enough for low-power stages, where the principal need is for a quick check when something goes wrong. But in a tetrode final amplifier it is desirable to have a constant check on grid current, screen current and plate current. These readings can all be made with a single meter, but doing this is almost like flying blind. There is no really adequate substitute for three good milliammeters in a tetrode amplifier.

Caution, Meter movements are easily damaged by v.h.f. r.f. current. Be sure that meter leads are properly filtered and bypassed for r.f., as even a small amount of r.f. current may get into the meter coil and destroy the meter's accuracy, or even burn it out completely. This is especially likely to happen at 144 Mc. and higher, where conventional bypassing is often ineffective.

Meter styles have changed considerably in the past few years, which means that one can pick up some good buys in the older types. There is no better investment. Desirable ranges are 0-1 ma., for a general-purpose meter; 0-50, for grid or screen metering; 0-100, for low-power plate indication; and 0-500 or 0-1000 for high-power plate current. If you intend to run a Class AB1 linear on s.s.b. or a.m., a zero-center 50-ma. meter is wonderful for the screen circuit. Buy good meters, and don't overlook ham auctions; a two-dollar bid may get you a real bargain there, if you know your meters.

Volt-Ohmmeter. This instrument and the vacuum-tube voltmeter (v.t.v.m.) are often confused in the newcomer's mind. The v.o.m. is a self-contained unit capable of measuring current and voltage, a.c. or d.c. Usually there is a built-in battery and ranges for measuring d.c. resistance. Most instruments also have decibel scales. Such a meter is almost a must for the fellow who is going to do any home-building or trouble-shooting. Here, again, get a good one—a standard make. It is best to look for one that will measure up to 5000 volts d.c., rather than the less expensive models that usually measure only up to 1000 volts or less.

Another point to watch for in the v.o.m. is the ohms-per-volt rating. Usually an inexpensive meter will be the 1000 ohms-per-volt type. This is good enough for d.c. measurements where loading of the circuit by the meter is not a critical matter, but in measurements such as grid voltage developed by the r.f. drive in a transmitter amplifier stage, a high-resistance measuring device is a necessity. The better v.o.m.s are

20,000 ohms-per-volt, and this sensitivity is very desirable.

Vacuum-Tube Voltmeter. If you have a good v.o.m., you can live without the v.t.v.m., though this instrument enables you to make r.f. voltage readings that you can't get with the self-contained job. The v.t.v.m. usually operates from a.c. power, and consequently the user has the inconvenience of being tied to the a.c. line. It is fine for voltage readings, a.c. or d.c., and usually the v.t.v.m. has good provision for resistance measurements—but it is not usable for reading current directly.

Oscilloscope. Regarded as mainly a laboratory instrument not too many years ago, the scope is getting to be commonplace in ham shacks today. And well it may be, for here is one of our most versatile tools, once we take the trouble to learn how to use it. If you intend to use s.s.b. or a.m. linear amplifiers, the scope is a *must.* It's a big help in any voice work, and good for innumerable other purposes too. There are some good ones made especially for amateur applications.

The scope has one weakness in v.h.f. work: it is hard to keep the r.f. where you want it, with the result that scope patterns are often distorted, even at 50 Mc., and at higher frequencies they are almost certain to be. At least one scope designed for in-the-line monitoring of amateur signals, the Heath HO-10 or SB-610, can be made to do the job on 50 and 144 Mc. Mount a coaxial T fitting on the back of the scope, and run the coaxial line from the transmitter through this, instead of using the through-line connection provided on the instrument. See Chapter 13 for another approach.

Audio Oscillator. Useful for checking out speech amplifiers and modulators. Used with a scope, it may turn up some things about your phone equipment that will surprise you—like useless frequency responses well above the voice-frequency range. These are often responsible for the excessive sideband width observed on amateur signals, and they do nobody anything but harm.

Signal Generator. Helpful in receiver work, but expensive, if you pay enough to get one that is really useful. Maybe some surplus bargains. Look for coaxial termination and a good attenuator, calibrated in microvolts. You can do without it, but a good signal generator is nice to have if you can afford it.

S.W.R. Bridge. Absolute must for everyone; you can't do any good work on antennas without one. Get the power-indicating type if you can, but don't do without if you feel that you can't afford the good ones. You can make a Monimatch for peanuts, and the one in Fig. 11-13 will do a good job up through the 420-Mc. band. Most commercial jobs won't work well that high in frequency.

Dipper. Very useful, if you plan to do any building, other than kits. Good, even if you never build anything, for it can be a good aid in trouble-shooting.

Sweep Generator. Growing in amateur interest, though not an absolute necessity. Helps in aligning converter and receiver circuits for any desired passband—almost impossible by other methods.

To summarize, whether you buy or build your station, don't be trapped into the all-too common assumption that equipment for talking and listening is all that is needed to make a good ham layout. This is a *technical* pursuit; to engage in it properly requires some test equipment and the ability to use it. Right from the very first, set aside part of your ham radio budget for test gear. Then *use* it. This is a learn-by-doing process, and the earlier one starts, the better.

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[2] McCoy, "The Monimatch," *QST*, October, 1956, p. 38.

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A more sensitive instrument of the Monimatch family was described by DeMaw, "The Varimatcher," May, 1966, *QST*.

[3] Tilton, "Noise Generators—Their Uses and Limitations," *QST*, July, 1953, p. 10. Detailed information on vacuum-tube noise generators is also in the *ARRL Handbook*, 41st edition and later.

Noise generator information was brought up to date and equipment for 420 Mc. and higher frequencies was described in a three-part symposium, *QST*, February, 1964, pp. 23-35.

[4] Frye, "Adjustment Procedures for V.H.F. Converters," *QST*, October, 1958, p. 24.

Interference Causes and Cures

In one respect amateur radio is vastly different from most other hobby-type activities: from its earliest days it has existed in competition with other services. Because we occupy frequencies that are under constant pressure from other users of radio, amicable and successful solution of our interference problems is vital to our very existence.

With occupancy of the radio spectrum rising daily, and population density increasing almost everywhere, interference problems inevitably tend to multiply. Interference is a two-way affair. We both cause it and suffer from it, but the first is our major concern in these pages. Fortunately some of the steps we take to cure our neighbor troubles are beneficial in receiving situations as well.

In handling TVI and related interference problems two cardinal points should be kept in mind:

> Interference is primarily a public-relations problem, not a technical one. Every form of it can be cured; it is getting the job done amicably that is difficult.
>
> Being able to demonstrate that the transmitter is not at fault is not enough. The amateur *must* understand the factors involved, and be able to take or recommend corrective measures. Nobody is going to do this for us.

THE NATURE OF TVI

With these facts of community life established, let's look at the causes and cures of TVI, the v.h.f. man's major interference problem. The principal forms of TVI from v.h.f. transmitters are listed below, in the approximate order of their importance:

1) *Blocking* Every user of the 50-Mc. band in a Channel 2 area knows about this. It may run all the way from a light cross-hatching of the picture to complete blackout. Nearly always the visual effects change with transmitter modulation. Usually there is audio interference along with the picture trouble. The v.h.f. man's worst interference problem, it has been responsible for a high percentage of all TVI complaints reaching FCC in recent years. It is much worse in Channel 2 than on higher channels. but it is possible on all channels, 2 through 6, in receivers near to a 50-Mc. station. Blocking of Channel 13 by 220-Mc. energy is similar, but by no means so severe.

2) *Audio Troubles* R.f. pickup by the detector or audio circuits of a receiver (TV or radio) results in your voice riding through almost regardless of the tuning of the receiver, or even the setting of the audio gain control. Picture reception may be clear. This problem is not confined to TV receivers; it is common in all devices having audio amplifiers: hearing aids, p.a. systems, record players, musical instrument amplifiers, and so on.

3) *Image Response* This basic weakness of all superheterodyne receivers is explained in Chapter 3. As most commonly encountered in amateur v.h.f. circles it is responsible for reception of 2-meter signals in Channel 2, in TV receivers having the currently-used high intermediate frequency. If an old TV set with a 21-Mc. i.f. does not show interference from your 144-Mc. transmitter, and a newer one does, this is likely to be the cause. Along with it there will probably be some audio trouble (2), if you are using amplitude-modulated phone.

4) *Harmonics of Oscillator or Exciter Frequencies* Usually this shows up as a crosshatch pattern, independent of modulation, changing or disappearing when the transmitter frequency is shifted. Usually trouble develops only when the harmonic falls in a sensitive part of the TV channel. See Fig. 12-1 and 2. Examples of common combinations are the 9th harmonic of 6-Mc. stages and the 7th harmonic of 8-Mc. ones, falling in Channel 2; the 10th harmonic of 8 Mc. in Channel 6; 7th harmonic of 25 Mc.

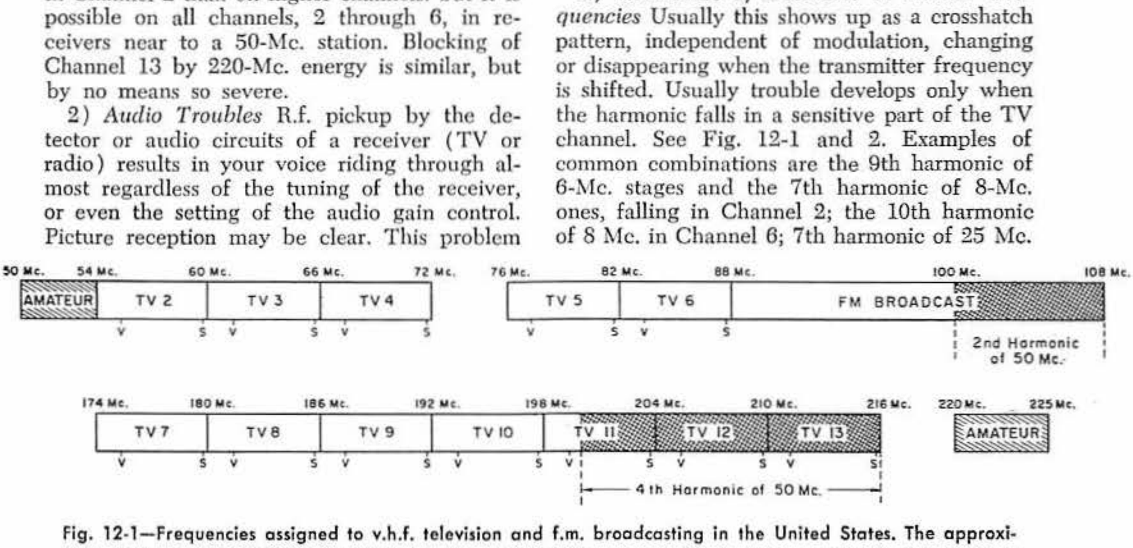


Fig. 12-1—Frequencies assigned to v.h.f. television and f.m. broadcasting in the United States. The approximate positions of the video and sound carriers are indicated on each TV channel. Crosshatched areas show second and fourth harmonics of the 50-Mc. band. Positions of the amateur 50- and 220-Mc. bands with respect to the TV channels are also indicated.

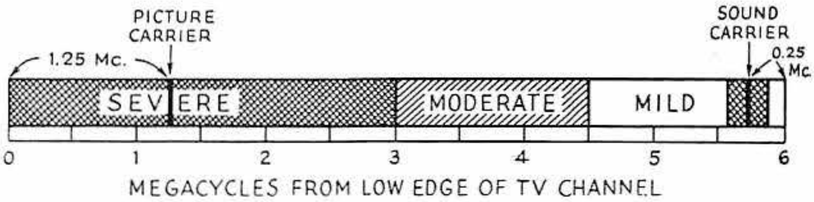


Fig. 12-2—Locations of the video and sound carriers in a black-and-white TV signal, showing relative severity of interference caused by harmonics falling in various parts of the channel. This information can be put to good use by v.h.f. amateurs in instances where mild harmonic interference is encountered. The trouble may be corrected, or at least alleviated considerably, by shifting the operating frequency so that the offending harmonic is moved out of a sensitive frequency range.

in Channel 7; 4th harmonic of 48 Mc. in Channel 9 or 10, depending on the operating frequency. There are others, but these are the most common sources of trouble. Exciter frequencies may also get directly into the receiver's i.f. system, in some instances.

5) *Final-Stage Harmonics* The 4th harmonic of 50 Mc. falls in Channels 11 to 13, depending on the operating frequency. Various harmonics of 50, 144 and 220 Mc. fall in the u.h.f. TV band, though usually they are not strong enough to cause much trouble. The 2nd harmonic of 50 Mc. falls in the f.m. broadcast band, and while this has not given us much trouble in the past, increasing use of f.m. could change this picture.

There are many other possible sources of TVI from the operation of amateur v.h.f. transmitters, but it is safe to say that at least 95 percent of our problems are covered by the above list. Items 1 and 2 are by far the most common, and certainly they are the most troublesome. We will look into each in some detail, but first go back over the list and be sure that you understand each one. Then remember, despite all you may have heard, these interference problems *can* be solved.

The frequencies assigned to television and f.m. broadcasting in the v.h.f. range are shown in Fig. 12-1, together with the harmonics of the 50-Mc. band that fall in this range. Prevention of radiation in these assignments, Items 4 and 5, is the amateur's responsibility. Items 1, 2 and 3 are receiver defects, which must be corrected at the receiver.

It is of utmost importance that the amateur know how to recognize the nature of the interference, so that he can correct the trouble if it lies in his transmitter, or recommend the corrective measures to be taken at the receiver. The earlier the amateur gets into the matter the better, for friendly relations between him and the set owner are vital, if a solution is to be reached.

CORRECTING TV RECEIVER DEFICIENCIES

Fundamental blocking (1) is a receiver problem. The r.f. circuits of a TV set are broad in frequency response. If they were not, picture quality would suffer, since the television picture and its sound occupy a channel 6 megacycles wide. It is not easy to build a receiver that will pass 54 to 60 Mc. and reject r.f. from a nearby amateur station operating on 50 to 54 Mc. It is unlikely that TV manufacturers will ever mass-produce receivers that do it effectively.

Though blocking troubles are much more severe in Channel 2, if the 50-Mc. signal is strong enough it may block the receiver on all low-band channels, 2 through 6. Clearing any channel above 2 is usually done quite readily. The simple stub to be described later will nearly always handle it. Why clearing Channel 2 of 50-Mc. interference is more difficult is obvious from Fig. 12-2, which shows the locations of the sound and picture carriers in a TV channel, as well as the interference potential of any signal falling in the channel. In Channel 2 the picture carrier is at 55.25 Mc., which is just too close to the 50-Mc. amateur signals to make it a simple matter to keep the latter below the overloading level.

There is a related problem in connection with 50-Mc. interference to the TV sound in Channel 2, in receivers of the intercarrier sound type, which nearly all TV sets are. With such receivers a signal 4.5 Mc. below the picture carrier can cause severe sound interference, the severity depending on the selectivity and alignment of the TV receiver's i.f. system. 55.25 Mc. -4.5 Mc. $= 50.75$ Mc. This makes 50.75 Mc. the worst possible spot on which to operate in the 50-Mc. band with a.m. phone, from the standpoint of sound interference. Proper alignment of the TV set makes a big difference with this trouble, but a practical fact of life in a Channel 2 area is that staying well away from 50.75 Mc. is very desirable. Any operating frequency above 50.4 Mc. makes interference very much more likely.

The approximate range over which 50-Mc. signals are likely to overload TV receivers in Channel 2 was shown graphically in QST by

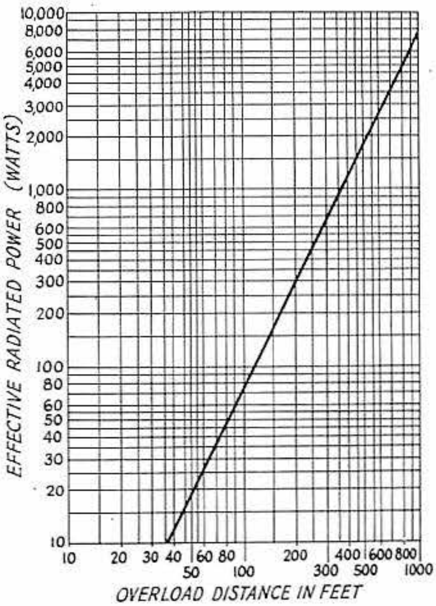


Fig. 12-3—Average overload distance for a TV receiver on Channel 2 and an amateur station between 50 and 51 Mc. The effective radiated power is the transmitter output multiplied by the antenna gain (*not* in decibels). The supersensitivity of misaligned inter-carrier-type receivers to 50.75-Mc. signals is not included.

50-Mc. pioneer W2IDZ. Fig. 12-3 is from his now-classic treatment of the 50-Mc. TVI problem which paved the way for today's thousands of 50-Mc. enthusiasts who now manage to live with Channel 2.[1] This is a matter of effective radiated power. If your transmitter puts out 100 watts, and your antenna has a gain of 10 db. (10 times) your e.r.p. is 1000 watts. If the antenna's main lobe fires into the TV antennas your sphere of evil influence will be roughly 400 feet in radius.

You can cut this in several ways without touching a TV installation. Raising the 50-Mc. antenna to the point where its main lobe of radiation is well above the TV receivers and antennas can knock the e.r.p. down by 20 db. or more. This would mean using 10 watts instead of 1000 at the left side of Fig. 12-3. The net effect of a 20-db. reduction in signal level at the TV antenna is a reduction in interference radius by roughly a factor of 10 in distance. This could make a very large difference in a built-up residential area. Blocking interference has a sort of threshold; raising the antenna may put you on the safe side of it.

Where the receivers and antennas are close to the transmitter the latter should be well-shielded and the transmission line nonradiating, if the high amateur antenna is to pay off to the greatest extent. This is particularly important in the multi-family dwelling. The indoor dipole (often the timid soul's last resort) may be the worst possible approach in such circumstances.

Getting the radiated power well above the TV sets also helps the effectiveness of any corrective measures used on the receivers. A high-pass filter installed on the receiver may be relatively ineffective if there is a strong r.f. field around the receiver itself. Complete shielding of the receiver, a difficult and seldom-taken step, may then be the only interference cure.

Though what has been said thus far is mainly concerned with 50-Mc. blocking-type interference, the principles apply equally to image and audio problems, regardless of the amateur transmitting frequency.

Using Stubs and Traps

If the fundamental interference is mild and the TV signals are strong, a simple quarter-wave stub of Twin-Lead cut to the transmitter frequency and connected to the TV receiver antenna terminals will take care of it. The stub is a good first step in any case, as it costs almost nothing, is easy to try, and ordinarily has little or no effect on TV reception. Such a stub is an electrical quarter-wave-length at the transmitter frequency, open at the far end. If it is fitted with open-end lugs at the other end it can be slipped under the receiver antenna terminals readily.

Start with a piece a bit more than $0.82\frac{\lambda}{4}$ in length, about 50 inches for 50 Mc. or 17½ for 144. Connect the stub at the antenna terminals, and trim it for length while watching the interference. When the interference level drops trim in small increments until interference disappears. The stub should have a negligible effect on the TV reception where a reasonably strong TV signal is available, and it will be effective for any but the worst cases of interference.

Another type of stub, this one tunable and requiring no electrical connection to the TV set, is shown in Fig. 12-4. It is a double stub, connected sandwich-fashion either side of the line to the TV receiver, and tuned for resonance at the transmitting frequency. It is thus a tuned trap, coupled to the line of the receiver. It is somewhat more effective than the self-resonant stub just described, and it may have less effect on the TV reception. It is convenient to make one up on a section of line that can be connected between the receiver antenna terminal board and the line to the TV antenna. The setup can be pretuned to the transmitter frequency, and thus be ready for a quick test.

The stub is shorter than the self-resonant type, to allow for the capacitive loading. About 38 to 40 inches is suitable for 50 Mc., and 10 to 11 inches for 144. A worthwhile refinement for a stub that will be used for test purposes is substitution of a split-stator variable capacitor for the mica trimmer shown. This allows adjustment of stub resonance without introducing

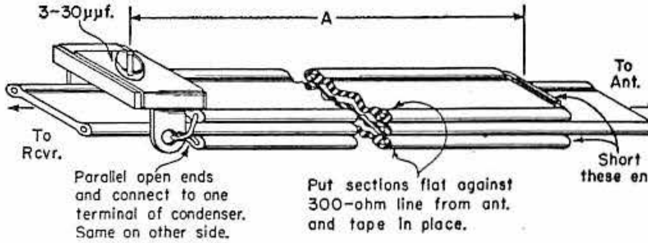


Fig. 12-4—Sandwich-type trap for installation in the 300-ohm line to the TV receiver. Approximate lengths (dimension A) are 40 inches for 50 Mc. and 11 for 144. Two traps are in parallel, one on each side of the TV line.

hand-capacity effects, and interference can be nulled out much more effectively. Once it is determined that this type of stub does the job, an inexpensive trimmer like that shown can be put on. It will work just as well, but is harder to tune accurately.

Traps tuned to the transmitter frequency can be inserted in the receiver line. Often a single trap in one side of the line will do the trick, or one can be connected in each leg. These can be resonated with the aid of a grid-dip meter to the transmitting frequency. For highest selectivity use the smallest amount of inductance that will tune to the transmitting frequency, with the capacitor available. Typical tuned circuits in v.h.f. equipment in this manual can serve as models.

In this discussion the accent has been on 50-Mc. applications but the principles apply to any v.h.f. problem where the interference is coming in on the antenna or transmission line to the receiver. Direct pickup of r.f. by the receiver, or by its a.c. line, will not be affected by stubs, traps or filters at the antenna terminals. All these devices are more effective if connected right where the antenna line enters the receiver chassis, rather than at the terminal board on the back of the cabinet, if there is an unshielded run of Twin-Lead from the terminal board to the tuner input of any appreciable length.

It is well to have any treatment ready for quick application, and to have the actual work done by the owner's serviceman, or by the local TVI Committee representative. Even when the neighbor is friendly he may be a little nervous about your working on the receiver. The quicker and more effectively the job is done, the better.

If you have a functioning TVI Committee, they probably already have a demonstration filter for this purpose. If none is available, you are not required to supply it. The set owner should be encouraged to take up the matter with the dealer from whom the set was purchased, as many manufacturers make provision for supplying filters where needed.

Correcting Audio Troubles

The stub is effective mainly in connection with blocking of the receiver by the transmitter's fundamental frequency (Item 1), and spurious receiver responses such as Item 3. It may help on audio troubles (Item 2) but these are more likely the result of direct pickup of r.f. by the receiver audio circuits. Filtering the antenna input of the TV set will probably do little good; some work on the receiver audio circuits is called for.

First it must be determined where the r.f. is getting into the set. The audio gain control can provide some clues. If adjusting it has no effect on the level of the interference (and this is usually the case) the pickup point is after the control in the circuit. Usually the first audio grid is the offender, in both radio and TV receivers, and the cure is simple: bypass or filter the interference out.

This can be done in several ways, some of which are shown in Fig. 12-5. If the audio grid resistor, R_1 in A, is a high value, reduce it to 2 megohms or less, and bypass the grid pin right at the socket with a value that will be large enough for r.f. bypassing without affecting the audio quality. Something around 500 pf. in a button-mica or ceramic stand-off or feed-through type is recommended. Disk ceramics are likely to be useless. The principle of series-resonant bypassing is good, though it will work for one band only. See Chapter 13.

The r.f. can be kept out with a resistor, R_2 in B, or an r.f. choke, RFC_1 in C. The resistor is the better, if it will work, as it is not frequency conscious. The function of C_1 is the same in B as in A. Where an isolating choke or resistor is used it should be right at the grid pin. Remove all the normal connections, and

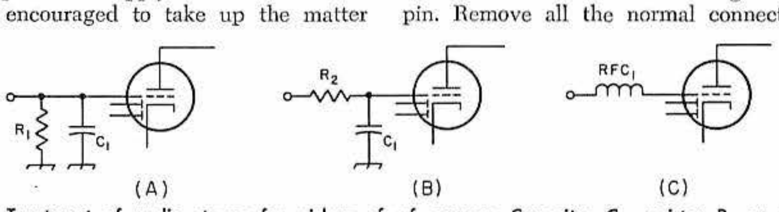


Fig. 12-5—Treatment of audio stages for pickup of r.f. energy. Capacitor C_1, resistor R_2 or r.f. choke RFC_1 should be connected right at the grid terminal. All circuits normally connected to the grid should be connected to the left side of these filtering devices.

reconnect them at the left end of the choke or resistor.

Decoupling of the heater circuits of the audio stages may be necessary, though this is unlikely. Series-resonant bypassing or ferrite-bead chokes are fine for heater decoupling, if needed.

Shielding long audio grid leads may be necessary, particularly where the gain control is mounted at some point remote from the amplifier stages. In this connection, some manufacturers are using a wire having conductive plastic shielding for audio grid leads. This may be ineffective at high radio frequencies.

Increasing use of transistors in audio circuits may call for different methods. Isolation procedures used for tube circuits should work, but shielding the complete amplifier may be easier in some instances, since the physical size may be smaller and the heat problems less, or nonexistent.

Clearing up audio problems is simple in principle, but the set-owner may not take kindly to the amateur's digging into his equipment. If there is any doubt, the wise approach is to give the owner's serviceman the necessary information, and have him do the job. Servicemen who are not amateurs may have little knowledge of the problem, so you may have to use diplomacy in two directions. Thus it is doubly important that you know precisely what you are talking about, in recommending corrective measures.

Pickup by the receiver's a.c. line may be a factor, though not too often in v.h.f. work. R.f. filtering of the a.c. line where it enters the receiver is the answer here. Heavy-wire chokes (No. 18 or so) and good r.f. bypassing are the treatments. Filters on the plug end of the a.c. cord, where it plugs into a wall outlet are almost never of any value.

There is one practically certain cure for all audio-rectification problems: give up operating with amplitude-modulated phone. Where people live close to each other, and especially in apartment houses so common in our larger cities, use of a.m. on the v.h.f. bands is an invitation to trouble. You can't fix 20 to 50 or more broadcast receivers, TV sets, record players and hearing aids, and your a.m. audience can easily run this high or higher. Going over to f.m. or c.w. is easy, and it works wonders. It has made ham radio possible for many city dwellers, and it could solve neighbor problems for thousands more.

KEEPING HARMONICS AT HOME

So far we've been concerned mainly with troubles that arise as a result of receiver deficiencies. With v.h.f. TVI, at least, they are in the vast majority. The possibility that the transmitter may be at fault should not be overlooked, however, and every possible check should be made on this before operating extensively in an area where there are TV sets nearby. The importance of doing this before the TVI complaints begin to roll in cannot be over-emphasized. If you have demonstrated to your own complete satisfaction that your transmitter is "clean," you can face your neighbors with confidence and good humor. These personal attributes are of inestimable value, for this TVI business, remember, is a public-relations problem.

Checking for Harmonics

The first order of business is to be sure that all available channels can be received clearly on your own TV receiver. If there is interference from your transmitter in any of them, don't wait for the angry phone calls. Find the trouble, and fix it—right now! If the various treatments already outlined do not clear up the interference, find out why, at once. Just because you have a 500-dollar superwhatsis transmitter does not guarantee that it is free of the troubles described in Items 4 and 5. Running it indiscriminately on the air can only bring down the righteous anger of the neighborhood around you. By then you may have lost the war, but the fighting will drag on and on.

If your own TV receiver does not respond to treatment for fundamental-frequency and audio-rectification ills, Items 1, 2 and 3, you've got harmonic problems. The first step is to find out where the offending harmonic energy is coming from. Put a non-radiating dummy load on the transmitter, and check again. Use a good load (see Chapter 11), preferably shielded. There are some good ones available ready-made and in kit form, if you don't want to make one from scratch. Do not use a lamp load; it can radiate plenty of energy to cause interference.

If there is no interference with the dummy load on, the harmonic radiation is from the antenna, and your problems are well on the way to solution. There are several simple and practical corrective measures. One of the best is a tuned antenna coupler, particularly if your antenna is fed with balanced line of any kind. Details in Chapter 8. Also very worthwhile is a high-Q coaxial or strip-line filter. More on these later in this chapter, and in *QST*.[2]

A low-pass filter connected in your antenna line is good harmonic radiation insurance, but such filters are rather difficult to make and adjust properly. You can buy them ready-made, and there have been good 50-Mc. designs in every edition of the ARRL *Handbook* for many years. There is little point in repeating such information here, when it is so widely available already.[3]

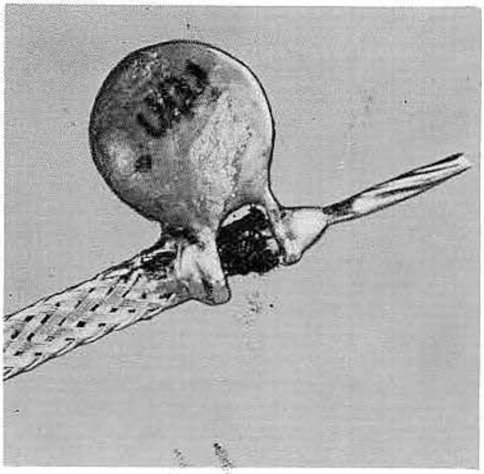


Fig. 12-6—Method for bypassing the end of a shielded power lead. Leads to the 0.001-uf. disk capacitor should be soldered as close as possible to the capacitor body. Shield over the wire should be grounded to the chassis at frequent intervals. This method is suitable for harmonics only up to about 100 Mc.

Both the antenna coupler and the high-Q filter have an important advantage over the low-pass filter: they protect the receiver more effectively, preventing overloading from strong signals below the amateur band in use, as well as above it. They might be quite helpful if you have a near neighbor who runs high power on the lower-frequency ham bands, or on some higher one.

Harmonic Sniffing

Harmonics that get out by way of the antenna disappear when the dummy load test is made. If the interference persists you have work to be done on the transmitter or its power circuits. Some kind of harmonic "sniffer" is now required. The simple field-strength indicator of Fig. 11-7 may be enough. It will cover TV Channels 2 through 6 as it stands. A smaller coil for L_2 will permit it to tune up through Channel 13, if need be. Plug a stiff wire or rod into J_1 for a pickup antenna.

With the transmitter running into a dummy load, place the pickup antenna close to the a.c. leads, power cable, any unshielded tubes or circuits, exposed meters, variable capacitor shafts, or any other part of the transmitter that could be radiating harmonic energy. If you find some you have a shielding or filtering job ahead; perhaps both.

Another effective harmonic radiation detector is the TV set itself. Cut a piece of Twin-Lead long enough to reach from the TV set to any part of the transmitter you want to check. Connect one end to the receiver antenna terminals. Short the other end to make a coupling loop. Tape bare wires so that there will be no shorting of high voltage into the TV set. Now use the Twin-Lead as a probe, coupling it to any suspected part, wire or circuit. If there is harmonic energy present the interference level will increase markedly as the probe is placed near the guilty component.

We used to build in complete harmonic protection into every transmitter. Experience has shown that so much TVI is the result of receiver deficiencies that we no longer do this. The chances are that any reasonably well-designed v.h.f. transmitter will be practically TVI-free, and that the receivers will be the culprits—but you cannot rely on it. If your own TV set shows evidence of harmonic interference, particularly with the transmitter on a dummy load, the chances are good that some of your neighbors will see the same evidence, unless it is visible in your own receiver only when in very close proximity to the transmitter.

Harmonic Suppression

Curing TVI is not a black-magic operation, whether it is the fault of the receivers or your transmitter. All transmitters generate harmonics. Yours is a veritable Pandora's Box full of them; you just have to keep the lid down—tight. Shielding is relatively easy. Most transmitters already have it, but adding it is no great chore. Just be sure that the shielding completely encloses every part of the r.f. portion of the rig. Then, if the harmonics still come out, you can find the leaks and stop them. Here are the common leaky spots:

Power Cabling Even with complete shielding, leads coming out of the r.f. portion of a transmitter are likely to have harmonic r.f. on them. Getting rid of it is no great problem. Shielded wiring in the transmitter is good insurance. Where the lead comes out of the transmitter

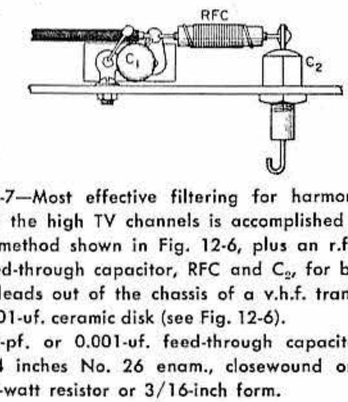


Fig. 12-7—Most effective filtering for harmonics up through the high TV channels is accomplished by use of the method shown in Fig. 12-6, plus an r.f. choke and feed-through capacitor, RFC and C_2, for bringing power leads out of the chassis of a v.h.f. transmitter.
C_1—0.001-uf. ceramic disk (see Fig. 12-6).
C_2—500-pf. or 0.001-uf. feed-through capacitor.
RFC—14 inches No. 26 enam., closewound on high-value 1-watt resistor or 3/16-inch form.

housing it should be filtered. The simple device of Fig. 12-6 will take care of harmonics and other spurious radiations in all the low TV channels, 2 through 6. Ground the shield on the wire at intervals inside the rig, and at the point where it leaves the enclosure.

If exciter or final-stage harmonics, such as 4×48 or 4×50, are radiated by power leads, the 12-6 method may not work, since disk ceramics are ineffective above about 100 Mc. Bringing out leads on feed-through capacitors is much better. See Fig. 12-7.

Chassis Leaks Harmonics, especially those in the upper v.h.f. and u.h.f. TV bands, can leak out of strange places. One exciter for 50 and 144 Mc. built by the author showed harmonic interference in Channels 10 through 13. This got no worse when the exciter drove a kilowatt amplifier. Some harmonic energy was found on the power leads. Decoupling as shown in Fig. 12-7 helped, but there was a faint pattern left.

Using a TV set for the visual indicator, it was found that the metal rings on the exciter tuning knobs were hot with harmonic energy. The receiver blacked out when the Twin-Lead probe was brought near to them. The variable capacitors tuning the exciter stages were the type having small rectangular studs for mounting, providing no way of grounding the rotors directly to the panel. Substituting variable capacitors having threaded bushings on the rotor shafts, permitting direct grounding to the panel or chassis cleared this trouble completely.

Long cracks in a chassis, or between the chassis and its cover plate, can act like slot radiators for harmonics. This is why transmitter shielding is fastened with so many screws.

Harmonic Generators

A transmitter with perfectly clean output can still have harmonic troubles, for harmonics can be generated in strong r.f. fields. Crystal diodes can do it. Look out for them, wherever they may be. Check for corroded connections in the antenna system, in your own array, or in the TV antenna. Watch for poor metal-to-metal contacts not directly connected to either your antenna system or that on the TV set being interfered with. This condition is found fairly often on apartment house roofs, where the litter from years of erection and decay of TV antennas may be strewn, and metal oxides are turned into harmonic generators by the impact of appreciable amounts of transmitter r.f. power. Try for shipshape installation of the amateur antenna, and for antenna height that puts the main lobe of radiation completely above the TV antennas and rooftops.

Designing Around Harmonic Problems

Where radiation of harmonics of oscillator or exciter frequencies is causing trouble in the high v.h.f. TV channels, as in the 4th harmonic of tripler stages working from 48 to 144 Mc., it is often possible to use a different frequency multiplying sequence and avoid the problem. This is not an ideal cure, since radiation on anything but the wanted frequency should be held to the practical minimum, but it can be an easy solution to a local problem. There is nothing sacred about common frequency multiplying practices, and many 2-meter men in Channel 10 areas have found relief by changing the order of frequency multiplication from 8-24-48-144 to 8-24-72-144. There can still be energy in Channel 10 (8th harmonic of 24 Mc.) but it is almost certain to be far lower than when the 48-144 sequence is used.

Another example of taking the easy way out is the elimination of the 10th harmonic of 8.4 Mc. in Channel 6 in 50-Mc. transmitters by going to 6.3 Mc. or 12.6 Mc. in the oscillator stage. Again, this is not the best solution, but it may be a practical one in some circumstances. The right way to do the job is to fix the installation, so that the offending harmonics are not allowed to get through to the transmitting antenna, or to the TV receiver.

A sure cure for most of these troubles is a high starting frequency in the exciter. With a 72-Mc. oscillator in a 2-meter rig there is no chance of a harmonic in Channel 10. (A 48-Mc. oscillator would be no help.) In a 50-Mc. transmitter many of the troubles can be cleared by starting with a 50-Mc. oscillator. It should be emphasized that this does nothing for the fundamental-overload problem in the low TV channels, however, and most 50-Mc. TVI is of this nature. The 50-Mc. oscillator also does not prevent radiation of a 4th harmonic in Channel 11, 12 or 13. This must be suppressed by techniques discussed a few paragraphs back.

Antenna-Mounted TV Boosters

Antenna-mounted boosters using bipolar transistors, currently popular in TV fringe areas, overload very readily, adding greatly to the v.h.f. man's interference problems. This is especially troublesome, as it results from the viewer having purchased an expensive antenna system, usually when a color TV receiver is installed. Conversion to field-effect transistors, raising the amateur v.h.f. antenna, and installation of a stub on the booster input are known cures. This problem, currently one of the most difficult encountered by v.h.f. men, is likely to be with us for some time.

COAXIAL AND STRIP-LINE FILTERS

If harmonics or other spurious frequencies appear in the output of an amateur v.h.f. transmitter they can be kept out of the antenna by a high-Q tuned circuit inserted in the line between the transmitter and antenna. Such a "filter" will pass only a very narrow band of

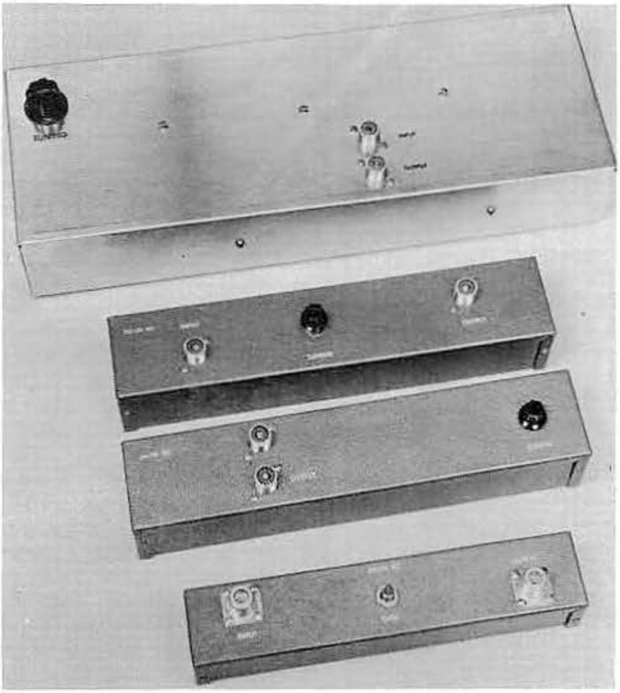


Fig. 12-8—High-Q strip-line filters for 50 Mc. (top), 220, 144 and 420 Mc. Those for the two highest bands have half-wave line circuits. All use standard chassis.

frequencies, offering a substantially impassible barrier to all others. The tuned filter can be helpful in receiving as well, since it will reject energy on frequencies other than the desired ones, and thus prevent overloading from out-of-band signals.

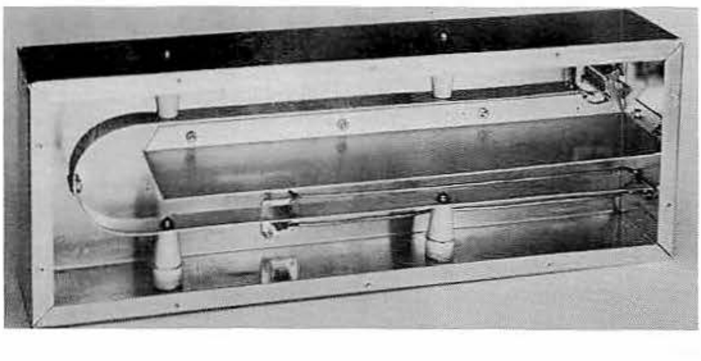
Antenna couplers described in Chapter 8 (Figs. 8-23 and 8-24) perform this function, but higher rejection of unwanted frequencies is possible with the tuned-line filters of Fig. 12-8. Examples are shown for each band from 50 through 450 Mc. Construction is relatively simple, and the cost is low. Standard boxes are used, for ease of duplication. Coaxial filters, also using low-cost components, may be found in *QST* for October, 1964.[2]

Quality of the filter elements is important for best results. Use large conductors and the best possible connections, particularly in high-current areas. Copper or brass, preferably silver-plated, is fine. Aluminum is satsifactory, and even fruit juice cans can be used, if all metal-to-metal contacts are clean and solid. Insulation should be kept to a minimum, especially at or near the high-impedance end of the line. A movable-disk capacitor, requiring no supporting frame or insulating material, is good for line tuning. If conventional variable capacitors are unavoidable, use types with high-quality insulation, and preferably no metal frame other than the minimum needed to support the plates. The type with threaded shaft bearing, permitting direct grounding of the rotor, is preferable.

The filter is not a magical device. To get high selectivity and rejection of unwanted frequencies it should not be loaded too heavily. A properly-adjusted filter will have some insertion loss, and its tuning will be critical. If the rejection need not be extremely high the coupling into and out of the filter can be adjusted to broaden response and reduce insertion loss. Two filters can be used in series, for very high rejection of unwanted frequencies. What you want to do with a filter determines how you adjust and operate it.

A typical use for a coaxial or strip-line filter is to prevent radiation of unwanted harmonics of the exciter frequencies in a 50-Mc. transmitter. The filter of Fig. 12-9 is selective enough to pass 50-Mc. energy and attenuate the 7th harmonic of an 8-Mc. oscillator, that falls in TV Channel 2. With an insertion loss at 50 Mc. of about 1 db., it can provide up to 40 db. of attenuation to energy at 57 Mc. in the same line. This should be more than enough attenuation to take care of the worst situations, provided that the radiation is by way of the trans-

Fig. 12-9—Interior of the 50-Mc. strip-line filter. Inner conductor of aluminum strip is bent into U shape, to fit inside a standard 17-inch chassis. Coupling is by L-shaped loops about ¼ inch above and below the tuned line, lower right.



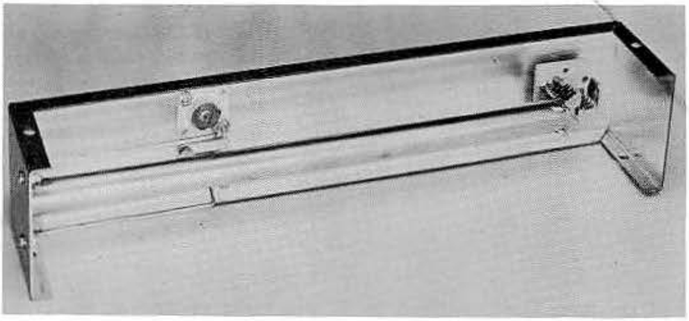


Fig. 12-10—The 144-Mc. filter has an inner conductor of ½-inch copper tubing, grounded to the left end of the case and supported at the right end by the tuning capacitor.

mitter output coax only. The filter will not eliminate interfering energy that gets out from power cables, the a.c. line, or from the transmitter circuits themselves. It also will do nothing for TVI that results from deficiencies in the TV receiver, such as the various problems we have already discussed.

Building the Filters

The 50-Mc. filter, Fig. 12-9, uses a folded line, in order to keep it within the confines of a standard chassis. The case is a 6 by 17 by 3-inch chassis (Bud AC-433) with a cover plate that fastens in place with self-tapping screws. An aluminum partition down the middle of the assembly is 14 inches long, and the full height of the chassis, 3 inches. Construction should be clear from the photograph.

The inner conductor of the line is 32 inches long and 13/16 inch wide, of 1/16-inch brass, copper or aluminum. In the model shown this was made from two pieces of aluminum spliced together to provide the 32-inch length. Splicing (visible at the left end of the U-shaped inner conductor) seemed to have no ill effect on the circuit Q. The side of the "U" are 2⅜ inches apart, with the partition at the center. The line is supported on ceramic standoffs. As may be seen from Fig. 12-9, these were shimmed up with sections of hard wood or bakelite rod, to give the required 1½-inch height.

The tuning capacitor is a double-spaced variable (Hammarlund HF-30-X) mounted 1½ inches from the right end of the chassis. Input and output coupling loops, visible on each side of the line, lower right of Fig. 12-9, are of No. 10 or 12 wire, 10 inches long. Spacing away from the line is adjusted to about ¼ inch. This may be increased for higher rejection, but this will result in increased insertion loss. The position of the input and output coaxial connectors is shown in Fig. 12-8.

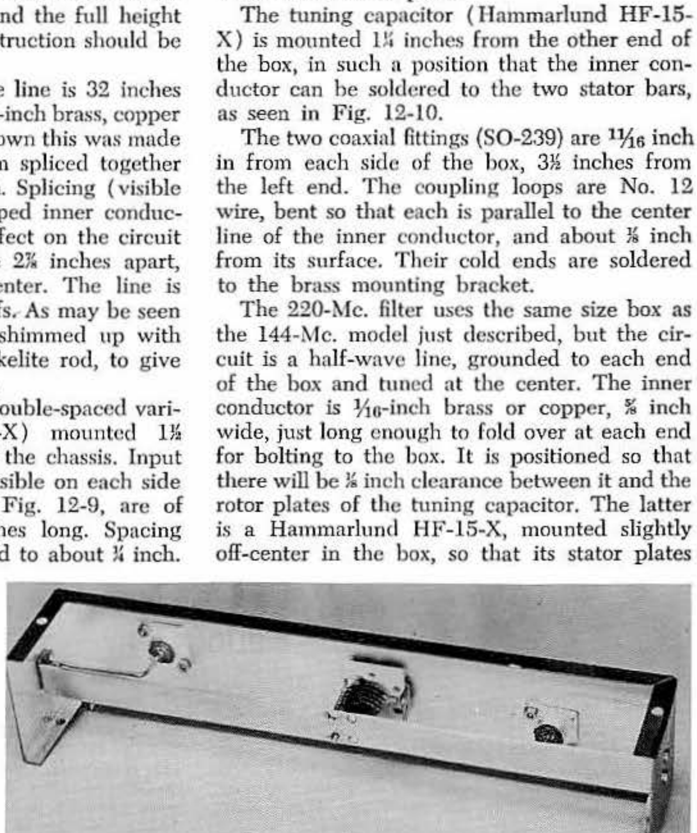
The 144-Mc. model, second from the bottom in Fig. 12-8, is housed in a 2¼ by 2½ by 12-inch Minibox (Bud CU-2114-A). The inner conductor (see Fig. 12-10) is ½-inch copper tubing 10 inches long. One end is slotted ¼ inch deep with a hacksaw. This slot takes a brass angle bracket 1½ inches wide, ¼ inch high, with a ½-inch mounting lip. The ¼-inch lip is soldered into the tubing slot, and the bracket is then bolted to the end of the box, so as to be centered on the end plate.

The tuning capacitor (Hammarlund HF-15-X) is mounted 1¼ inches from the other end of the box, in such a position that the inner conductor can be soldered to the two stator bars, as seen in Fig. 12-10.

The two coaxial fittings (SO-239) are 11/16 inch in from each side of the box, 3½ inches from the left end. The coupling loops are No. 12 wire, bent so that each is parallel to the center line of the inner conductor, and about ⅛ inch from its surface. Their cold ends are soldered to the brass mounting bracket.

The 220-Mc. filter uses the same size box as the 144-Mc. model just described, but the circuit is a half-wave line, grounded to each end of the box and tuned at the center. The inner conductor is 1/16-inch brass or copper, ⅝ inch wide, just long enough to fold over at each end for bolting to the box. It is positioned so that there will be ⅛ inch clearance between it and the rotor plates of the tuning capacitor. The latter is a Hammarlund HF-15-X, mounted slightly off-center in the box, so that its stator plates

Fig. 12-11—A half-wave strip line is used in the 220-Mc. filter. It is grounded at both ends and tuned at the center.



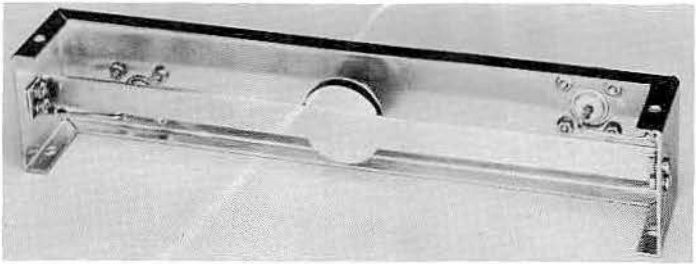


Fig. 12-12—Construction of the 420-Mc. filter is similar to the 220-Mc. one, except that it is shorter, and a disk-type tuning capacitor is used.

connect to the exact midpoint of the line. The 5/16-inch mounting hold in the case is 5⅛ inches from one end. Two small holes drilled in the inner conductor allow it to slip over the stator posts, for soldering in place.

The links for input and output coupling are at opposite ends of the box, as seen in Fig. 12-11. The SO-239 coaxial fittings are 1 inch in from opposite sides of the box, 2 inches from the ends. Their coupling links are No. 14 wire, ⅛ inch from the inner conductor of the line.

The 420-Mc. filter is similar in design, using a 1⅝ by 2 by 10-inch Minibox (Bud CU-2113-A). A half-wave line is used, with disk tuning at the center. The disks are 1/16-inch brass, 1¼-inch diameter. The fixed one is centered on the inner conductor, the other mounted on a No. 6 brass lead-screw. This passes through a threaded bushing, which can be taken from the end of a discarded slug-tuned form. An advantage of these is that some kind of tension device is usually included. If there is none, a lock nut can be used.

Type N coaxial connectors were used on the 420-Mc. model. They are ⅜ inch in from each side of the box, and 1⅜ inches in from the ends. Their coupling links of No. 14 wire 1/16 inch from the inner conductor are visible in Fig. 12-12.

Adjustment and Use

If you want the filter to work on both transmitting and receiving, connect up your system as shown in Fig. 12-13. With this arrangement you need merely adjust the filter for minimum reflected power reading on the s.w.r. bridge. This should be zero, or close to it, if the antenna is well-matched. The bridge should be used, as there is no way to adjust the filter properly without it. If you insist on trying, adjust for best reception of signals on frequencies close to the ones you expect to transmit on. This works reasonably well only if the antenna is well matched.

When the filter is properly adjusted (with the s.w.r. bridge) you may find that reception can be improved by retuning the filter. Don't do it, if you want the filter to work best on the job it was intended to do: the rejection of unwanted energy, transmitting or receiving. If you want to improve reception with the filter in the circuit, work on the receiver input circuit. To get maximum power out of the transmitter and into the line, adjust the transmitter output coupling, not the filter. If the effect of the filter on reception bothers you, connect it in the line to the transmitter only.

Don't expect the filter, or any other device you can connect onto your station, to be a TVI cure-all. There is no such magic box available, at any price. Curing TVI calls for some understanding of what goes on in transmitters, antennas and TV receivers. There is no easy way out, but by the same token, there is no completely hopeless situation. Every form of TVI *can* be cured.

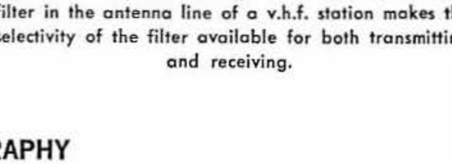


Fig. 12-13—Preferred method of connecting a tuned filter in the antenna line of a v.h.f. station makes the selectivity of the filter available for both transmitting and receiving.

BIBLIOGRAPHY

[1] Ladd, "50-Mc. TVI—Causes and Cures," June and July, 1954, *QST*.

[2] Tilton, "Coaxial-Tank V.H.F. Filters," October, 1964, *QST*.

[3] ARRL *Handbook*, Chapter 23, all modern editions. Also Tilton, "TVI Hints for the V.H.F. Man," April, 1953, *QST*.

Other references of interest to the v.h.f. worker include the following *QST* items:

U.H.F. Strip TVI—November, 1953, p. 45; December, 1953, p. 62; March, 1954, p. 28.

"TVI-Proofing the ARC-5 V.H.F. Transmitter," Johnson, November, 1950, *QST*.

Techniques for dealing with various forms of TVI were included in scores of *QST* articles of the early 1950s. They are of historical as well as technical interest, since they tell the month-by-month story of the TVI battle that was eventually won by the amateur.

Bits and Pieces

Most v.h.f. enthusiasts are experimenters at heart. Their first projects when they are bitten by the v.h.f. bug may be kits or duplicates of *QST* or *Handbook* items, but soon the urge develops to custom design and build v.h.f. gear. This can take the form of studying published constructional articles for ideas, and then adapting them to one's own needs. We like to think that much of the material in this book will be used in this way. Eventually, with accumulated knowledge and experience, most v.h.f. men get to the point of designing for their own requirements, rather than merely duplicating to the last nut and bolt what someone else has already worked out.

This last section of our book is for these amateurs. It will be something of a hodge-podge of ideas and techniques that might have been worked into other chapters, but which more logically fall into the "Hints and Kinks" category. The *QST* section under that title has been a most-read feature for generations. We hope that our version of it will find equal acceptance. Our thanks go to the scores of v.h.f. men who supplied the items you will find here and elsewhere throughout this manual.

IMPARTING THE "COMMERCIAL LOOK"

Well-built ham gear of good design usually works at least as well as equipment purchased ready-made, but it seldom looks the part. Even the simplest equipment can be given a quality look, if the builder will devote a little time and thought to appearance of the final product of his building efforts. Expensive cabinets are not necessary; even simple chassis-mounted units having no front panel or cabinet, in the usual sense, can be made neat and attractive in appearance with the use of a little paint, decals, and care in layout.

Painting

Every hardware store today carries aerosol-spray enamels, in a wide variety of colors. Black, grey and white are favored for ham gear, but other colors have their uses. Matching or contrasting colors can give many nice effects, especially for mobile equipment. Examples of home-sprayed units in this book include the superregenerative receivers in Chapter 4 and the 2-meter transceiver of Chapter 7, built by W1CER, who contributed the following suggestions.

The best paint job is usually possible if the metal parts are drilled, ready for assembly, but the equipment is not actually put together. This is practical on all but the most experimental items, and even these can usually be rebuilt in finished form, once the bugs are taken care of. Clean the metal with fine steel wool, to remove rough spots and dirt. With new aluminum this treatment is desirable to remove the high gloss, making a better base for paint. For exceptional durability, spray a first coat of zinc chromate, an undercoating finish also available in spray cans.

After the rub-down, clean with a grease solvent. Avoid touching the metal with the bare hands, as skin oils and acids can cause blemishes in the finished work. Prop the work up with a large area of newspaper or other protective covering under it. Spray paints dry quickly so dust is no problem, but select a place that is well-ventilated and clean.

If you've not used aerosol sprays before it may be well to practice a bit with some metal scraps. Read the directions; don't assume that you know how to handle these sprays. The manufacturer probably knows more than you do, and he wants your results to be good.

Shake the can thoroughly. Keeping the nozzle at least 12 inches from the work, spray with a sweeping motion, using just enough to cover. More will surely cause runs of paint, destroying the appearance. Allow several minutes for complete drying, and spray again, evenly and lightly. Now put the work aside for at least 24 hours. This allows the paint to age, and greatly lessens the chance of damage in handling.

Two-tone finishes can be made neatly by masking off any area that is to be painted a different color, or, in the case of aluminum, left its natural finish. Wherever metal is to be natural in color, a coating of clear lacquer will keep it looking its best much longer than if it is left uncoated. Large areas can be masked off with newspaper, with masking tape only at the edges. Press the tape firmly along the paint boundary desired, to prevent seepage under the edge of the tape. Keep the tape on until the paint has dried thoroughly.

After using spray paint, turn the can upside down and press the nozzle for a few seconds to clean out the spray jet. This simple precaution, often ignored though it is included in the directions, will make the next job much simpler than if it is omitted.

When using two different colors on a surface be sure that they are compatible. Test them in advance; some different paint bases may react on one another.

Old transformers and chokes can be made to look like new by painting. Clean them thoroughly of rust, loose paint, grease, etc. before spray painting. Sanding or scrubbing with steel wool may be needed.

Highlighting can be applied to cabinets and panels by painting with a base color such as grey or black, and then spraying over lightly with gold, silver or copper. For an effect of depth, use clear spray over the finished product. Take plenty of drying time between these operations.

Applying Decals

Neat labelling provides the final touch, and it is of practical value when other amateurs may want to use your equipment. Typewritten or hand-inked labels pasted onto equipment make it look like the work of a rank beginner, regardless of how skilfully the electrical and mechanical work has been done. Decals are easy to apply. A book of them with enough to last through many projects costs less than $2.00. They are available in black, white and gold.

The label desired is cut from the sheet and then soaked in water to separate it from the paper backing. Slide the decal onto the metal surface and move it into the desired position. If you get it slightly awry, dampen it and move it with a small brush. When it is lined up properly blot the moisture with absorbent cloth or paper towel. The label can be moved again by moistening, until it is permanently dissolved with Tekni-Solv or lacquer thinner. This should be applied with a small brush, using just enough to moisten the label area.

The solvent should be tested on a paint sample, as some solvent-paint combinations cause wrinkling and peeling. Paint should be allowed to dry for at least 48 hours before applying the solvent.

Other Appearance Factors

Choice of knobs can make or break the appearance of homebuilt gear. Occasionally an amateur will devote a lot of time and effort to building a neat outfit, and then spoil the whole effect with a random collection of knobs. Parts arrangement is important, too. Controls don't have to be perfectly balanced in their distribution over a panel area, but pleasing arrangements nearly always can be made without resorting to string drives, remote controls and various other mechanical devices.

Speaker grills offer an opportunity for appearance highlights. A grill of perforated aluminum painted black, mounted against a grey panel, will give a pleasing effect. With some thought and advance planning, and the expenditure of a little extra time, the final product of your handiwork can be something you'll be proud to show off to your friends.

SILVER PLATING—WHAT IT DOES, HOW TO DO IT

Silver is one of the best conductors known. Where very high conductivity is important, silver plating will improve almost any other metal. In addition, silver has a special attribute: it remains a good conductor when oxidized, whereas few other metals do. For these reasons many items of military u.h.f. gear are silver-plated throughout, and copious silver plating has come to be almost synonymous with quality in the minds of v.h.f. and u.h.f. workers.

But silver is expensive, so it is not so widely used in amateur applications. Just how much does it do for us, and is it worth the cost? There is no single answer, except that silver plating probably never did any harm, and it may be helpful. It makes soldering much easier, and it certainly improves metal-to-metal contacts, especially sliding ones. It is well worthwhile in the portions of circuits where r.f. current is high, as in the shorted end of a coaxial or parallel-line r.f. circuit.

Silver plating makes a measurable improvement in the Q of a v.h.f. circuit; 5 to 10 percent increase in 200-Mc. coils wound of copper ribbon resulted from before-and-after measurements in the ARRL Lab. It is probable that copper and brass tank circuits of the type used in the 144- and 432-Mc. amplifiers described in this book would be slightly better after plating. Tests on typical items have shown no measurable improvement in transmitter efficiencies through plating, but these have not been made on enough circuits to be sure that no benefit is obtainable. Certainly the long-term conductivity of silver-plated items, as compared with copper or brass counterparts that oxidize quickly with handling and use, should have some bearing on the value of plating to the amateur.

Plating can be done in several ways. First, you can take your parts to a plating shop. This costs money, but assures a good job. There are at least three do-it-yourself methods now available, including a home version of the process the plating shops use.

For this you need a silver anode and a quart of concentrated plating solution. Both are available from distributors of plating materials. They cost $6.00 each from Hoover & Strong Co., Tupper Bldg., Buffalo, N.Y. Other items required before you set up in the plating business are a voltage source, 1 to 3 volts d.c.; a 2-quart plastic dish, a 5-quart rinsing bucket, degreasing solvent, a pair of clip leads, and some fine steel wool. The plating solution will enable you to plate with other metals as well as silver. The plastic containers can be obtained from any hardware store.

Preparing the Work

Copper, brass and bronze are most suitable for silver plating. Steel can be plated, if it is first plated with copper. Whatever the metal it should be cleaned and polished before immersion in the plating bath. Rub it down with fine steel wool, and clean in a degreasing solution. Chemical houses supply degreasers, or you can boil the work in a mild solution of laundry detergent. Rinse thoroughly in clean hot water. Handle only with rubber gloves; finger oils and acids will prevent the metal from plating properly.

Plating

Use distilled water to dilute the plating solution, usually 3 quarts of water to 1 of solution. This must be at room temperature. Too warm a bath will cause discoloration, and too cold will make for spotty plating. Connect the metal to be plated to the negative side of a 1½-volt cell, and slide it into one end of the plating tank. Connect the silver anode to the positive terminal, and submerge it at the opposite end. Maintain a spacing of at least 6 inches between anode and work. Too close spacing causes excessive current flow and discoloration. Agitate the work frequently to prevent bubble formation on it.

Immersion time is usually 5 to 10 minutes. Longer will give heavier coating, and it is best to err on this side, as far as the r.f. quality of the plating is concerned. The higher the voltage the rougher the finish. Something between 0.5 and 1½ volts is best.

After plating is completed rinse immediately in fresh clean water, preferably lukewarm. Do not touch with the bare hands if you want a clean surface. To preserve the finish, spray with clear lacquer after the work is thoroughly dry. A lacquer spray does not affect the ability of the surface to take solder. If incomplete plating is found near solder areas it is probably due to the presence of flux. Such areas can be scrubbed with a stiff brush and xylol or alcohol. Replating can be done as needed, in the manner already outlined.

Caution: Silver plating solutions contain cyanide. Avoid breathing the vapors from the bath. In mixing, pour the plating solution into the water, *not* vice-versa. Wash hands thoroughly after any contact with the fluid. Do the plating in a well-ventilated room. Store the chemicals in clearly-marked containers, out of the reach of children.

Other Methods

Plating kits are available in several forms. An inexpensive one is made by Miniplating, Box 161, Middleboro, Mass. This consists of a plastic cylinder for holding two penlite cells, an electrode that fits in the end of the cylinder, clip leads, and a jar of plating solution in jelly form. The electrode is covered with a sponge-like plastic. The silver kit is $3.95, postpaid; extra jars of plating concentrate are $1.30 each.

To use the kit the electrode is dipped in the jelly, the clip lead connected to the work, and then the surface to be plated is rubbed with the coated electrode until silver is deposited.

Another method, very simple to use, involves a plating powder. It is applied with a damp cloth dipped in the powder, and then rubbed onto the surface to be plated. Because some rubbing is required, the resulting surface comes out very nice and smooth. The material, called COOL-AMP, is made by a company of that name, 8603 S.W. 17th Ave., Portland, Oregon. The powder is sold only in jars, minimum order 1 pound, $13.50, but a little goes a long way. Several would-be platers could do quite a bit of work each with one pound, which covers about 6000 square inches!

Both the above methods are best used with rubber gloves. The plating materials are a little rough on the skin otherwise, and neater work is possible if the fingers are kept from direct contact with the work or the plating substances. Several of the items described in this book were plated using the kit or the powder.

VARIABLE-FREQUENCY CRYSTAL HOLDER

The frequency at which a crystal oscillates is affected by the pressure on the crystal in the holder, in mounts such as the common FT-243, shown in its original form at A in Fig. 13-1. In B, a flexible top electrode is substituted, and provision is made for varying the pressure this exerts on the crystal.

In this system by W4RMU for swinging the crystal oscillator frequency, a spring electrode of 0.004-inch brass or 0.003-inch steel shim stock is used as the top plate, in place of the usual top electrode of the holder. The top cover of the crystal holder is drilled and tapped to take a ¼-inch screw (fine thread preferred, but ¼-20 is usable) which provides the pressure adjustment. The copper plate that made contact with the original electrode in the holder serves the same purpose in the revised one. The spring plate is bent around a ⅝-inch rod to give the curvature needed.

It is important that the corners of the pressure plate be completely free of roughness or burrs.

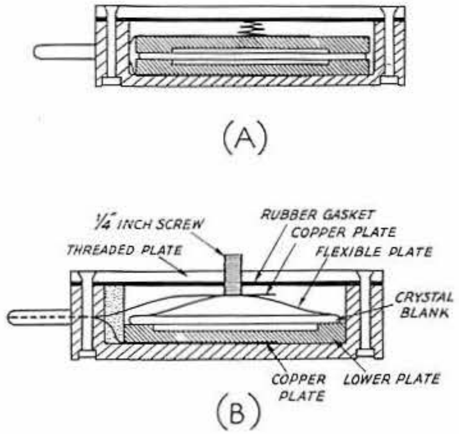


Fig. 13-1—Cutaway views of the FT-243 crystal holder, in its original form, A, and modified, B, for variable frequency control. In the latter a spring with adjustable pressure is substituted for the upper electrode. Spring tension is adjusted by means of a small screw in the cover plate.

Polish them carefully with an emery stone or very fine file. The range of frequency shift will be from the point where the spring plate touches the crystal at its center (low end of the range) to the point where the pressure no longer holds the crystal firmly in place.

A modified FT-243 crystal is shown in Fig. 13-2. The amount of frequency shift will vary from one crystal to another, with the types that are convex-ground giving the greatest usable swing. Activity varies somewhat over the frequency swing, dropping off quickly at each end. A typical surplus crystal on 6006.667 kc., which originally gave a 2-meter frequency of 144.16 Mc. now gives coverage from that frequency to 144.26 Mc., with substantially no change in final grid drive. This is in a circuit similar to the 6CX8 oscillators in several units shown in this book.

Fig. 13-2—An FT-243 surplus crystal, modified for variable frequency control.

CAPACITOR ROTORS—TO GROUND OR NOT TO GROUND?

The question of grounding the rotor of a tuning capacitor, whether to do it, and if so, how, bothers many builders of v.h.f. gear. In single-ended circuits, Fig. 13-3-A and B, grounding of the rotor, as in A, is usually the preferred method. The bypass capacitor, C_1, may be far from perfect, with the result that the rotor will have some r.f. voltage on it, and it may radiate harmonics, or make the tuning sensitive to hand capacity.

Choice of the tuning capacitor, C_2, may be important, too. Some variables have metal studs embedded in the ceramic end plate for mounting. With these a connection, usually to a rotor spider, must be made for grounding. The resultant lead inductance may be enough to leave the rotor above ground for r.f. voltage, at frequencies above 100 Mc. or so. The capacitor having a threaded rotor bushing, for direct grounding to the panel or chassis, is much better for grounded-rotor circuits.

In the push-pull circuit, C, or any other where a split-stator capacitor is used, grounding the rotor or not depends on several factors. If the center-tap of the plate coil is bypassed to ground there is no need to ground the rotor, and it may not be necessary in other circuits. Especially in the upper v.h.f. and into the u.h.f.

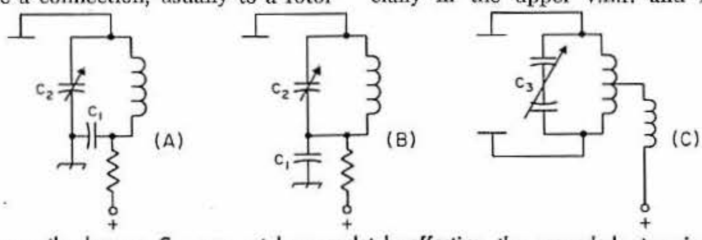


Fig. 13-3—Because the bypass, C_1, may not be completely effective, the grounded-rotor circuit, A, is preferred to that in B, which bypasses both the rotor and the low end of the plate coil. In the push-pull circuit, C, the rotor is best left ungrounded, unless the design of C_3 is such that good balance to ground is assured.

range, grounding the rotor of C_3 may unbalance the circuit severely, though this depends to some extent on the capacitor construction. With small butterfly types, as in the 6360 amplifier stages of some of our units in this book, it is almost impossible to avoid rotor grounding. In these circuits, with their well-balanced miniature capacitors, there is no reason for doing otherwise, but the coil center-tap should not be bypassed if the rotor is grounded. In split-stator capacitors with two sections in line on a single rotor shaft, some unbalance almost always results from rotor grounding. In one 432-Mc. tripler-amplifier formerly in the ARRL *Handbook*, running the tuning capacitor rotors above ground was a necessity. So great was the unbalance with the rotors grounded that neither tripler nor amplifier stage would operate at all in that condition.

SERIES-RESONANT BYPASSING

It is well-known that the inexpensive disk-ceramic and "dog-bone" types of capacitors are relatively ineffective for bypassing above about 100 Mc. or so. This is due mainly to their considerable lead inductance, even when they are connected as close to the elements to be bypassed as possible. Actually this lead inductance can be used to advantage, by selecting lead lengths that make the capacitor series-resonant at the frequency to be bypassed.

This approach is recommended by WA2-KYF, who supplied the information in Table 13-I, showing capacitor and lead-length combinations for effective bypassing of r.f. energy at frequencies commonly encountered in v.h.f. work. The values are not particularly critical, as a series-resonant circuit is broad by nature. The impedance of a series-resonant bypass is very close to zero ohms at the frequency of resonance, and it will be lower than most conventional capacitors for a considerable range of frequency.

A high-capacitance short-lead combination is preferable to a lower value with longer leads, because the former will be less likely to allow unwanted coupling to other circuits. For example, a 100-pf. capacitor with ¼-inch leads is a better bet than a 25-pf. with 1-inch leads, for bypassing at 144 Mc. The series-resonant bypass is worth a try in any circuit where instability is troublesome, and conventional bypassing has been shown to be ineffective. Screen, heater and cathode circuits are usually good candidates.

TABLE 13-I

Values of capacitance in pf. required for resonance at frequencies commonly encountered in amateur-band v.h.f. work, for leads of ¼, ½ and 1 inch in length.

Frequency Mc.	*¼-Inch Leads*	*½-Inch Leads*	*1-Inch Leads*
48-50	800	400	200
72	390	180	91
96	220	100	56
144	100	47	25
220	39	20	10

CUTTING AND TAPPING PREPARED COIL STOCK

Ready-made coil stock such as Air-Dux and Miniductor is a great aid in making neat and mechanically-stable v.h.f. circuits, but cutting and tapping it bothers many workers. The fine wire type with 32 turns per inch is particularly troublesome.

Such stock can be cut almost to the turn desired by pressing a knife between the desired turns and heating it with a soldering iron. With only light pressure on the blade, it will slice through the plastic insulating material with ease.

Tapping is easily done by pressing the wire down toward the axis of the coil, using a thin screwdriver blade. This makes a loop that can be reached readily with the tip of a small soldering iron. Tin the loop, and the wire that is to be connected to it. Put the two together and a light touch with the iron will make a good joint.

Pushing the wire down in this way also makes it easy to cut the wire without breaking the plastic strips. This is handy when a coupling loop (or another coil, closely coupled) is wanted. Merely cut the wire where it is bent in, and then unwind turns to give the necessary spacing between windings, without cutting the insulation. Heating the wire slightly will make it unwind easily.

STIFFENING ANTENNA TUBING

Lightweight tubing used in antenna work can be made much stronger by inserting a wooden plug at the point where strain is expected. This is particularly effective for a light vertical support for beams, where it runs through the tower bearing. This is the point where the tubing is most likely to collapse. Stuffing it with a wood dowel that is a close fit will strengthen it greatly. A foot or two either side of the tower bearing is all that is needed. Rug-pole stock is available at most lumberyards. Get hard wood for greatest strength. This may not be so readily available, but it is worth shopping around for.

Also good for this application, though heavy: ordinary thick-wall water pipe. It may bend, but it will not collapse or break. The so-called 1-inch water pipe (inside diameter) is fine. Lightweight steel tubing is probably the least desirable of all, because of its susceptibility to collapsing. Aluminum or dural tubing offers the best compromise between strength and light weight, of readily-available materials.

LAYING OUT A PARABOLIC CURVE

For effective work above 1000 Mc. a parabolic reflector is almost a must. Usually this means a trip to a surplus depot or junkyard, but parabolas of moderate size are not difficult to make. The antenna shown in Fig. 13-4 was used in the first amateur communication ever achieved on 2300 Mc., but the basic ideas it employs are still useful, if you want to make your own dish. the 5-footer shown has a gain of 200 (23 db.) in the 2300-Mc. band, yet it is light enough to be carried easily.

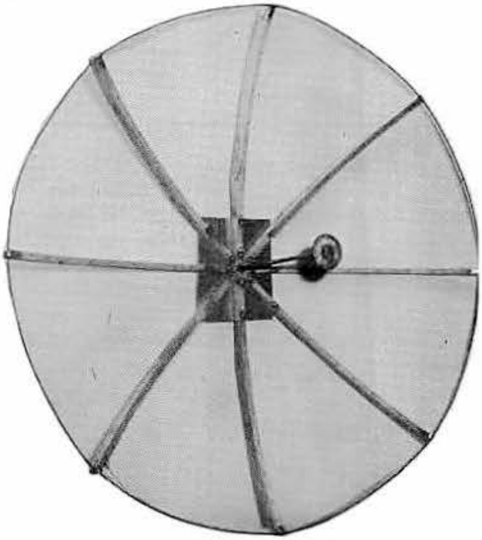


Fig. 13-4—A homemade parabolic reflector for microwave work. The model shown is 5 feet in diameter, but proportions shown in Fig. 13-5 may be used for any desired size.

The dimensions and shape of the wooden supporting frame are given in Fig. 13-5. Clamps of metal at the outer end of each of the 8 wooden arms take a hoop of heavy copper wire or small tubing that comprises the outer ring of the dish. The mesh can be window screening, hardware cloth, or any other mesh having openings less than 0.1 wavelength across in the largest dimension. The wooden supports can be made of plywood or pine shelving. Described originally by Koch and Floyd, in *QST* for July, 1946, p. 36.

The measurements need not be in inches. By using the proportions given, an antenna of any desired diameter can be made. Dimensions are not particularly critical. It is not absolutely necessary that the reflector be a true parabola, and it does not have to be round. A 12-foot parabolic-curve reflector essentially square in shape was described in detail by W1TQZ, in *QST* for April, 1961.

Details of the dipole feed and reflector for illuminating a parabolic reflector at 2300 Mc. are given in Chapter 10, and in the July, 1946, *QST* reference, above.

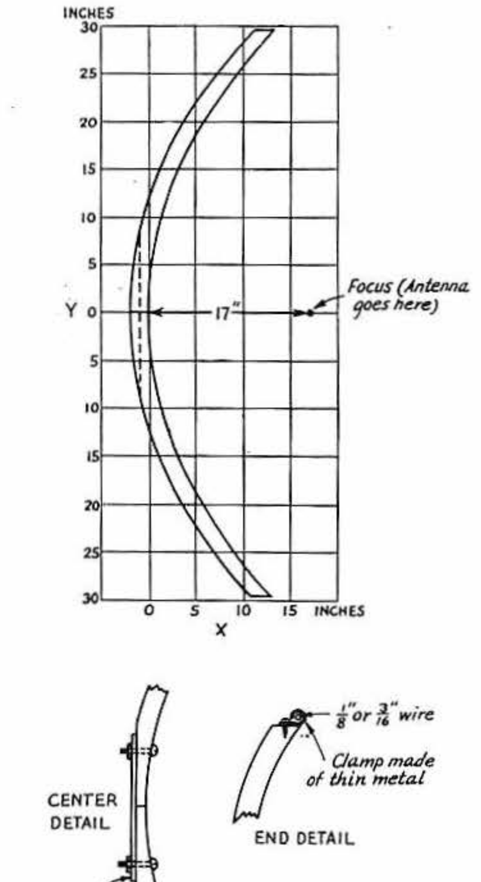


Fig. 13-5—The frame for a parabolic reflector can be cut from plywood or pine shelving, using the dimensions or proportions given above. The shape of the curve can be computed from the formula $Y^2 = 4AX$, where A is the distance from the center to the focal point, in this case 17 inches.

CLEAN SCOPE PATTERNS WITH V.H.F. TRANSMITTERS

Use of an oscilloscope is a must if you want the best possible signal quality with sideband or a.m. linear amplifiers. But when a scope is connected in the manner used on lower frequencies, stray r.f. on the scope plates produces all manner of pattern distortion.

This can be corrected to some extent by connecting a tuned circuit, resonant at the transmitting frequency, between the vertical plates of the scope, and then link-coupling some energy to it from the transmitter. The question of how to couple to the transmitter output still remains. W6GDO has a neat way of taking care of this, and a simple method of connecting to the scope as well.

A modified 83-1T coaxial T fitting is connected in the line between the transmitter and the load or antenna. The contact that normally is the common one in the fitting is removed, and a small capacitance substituted, for coupling into the line to the scope. With high power it may not be necessary to put in any coupling capacitor at all, and for intermediate power something up to about 5 pf. may be needed. This line can then be run to a link around the tuned circuit connected at the scope plates, or it can be terminated in a balun, the two inner conductors of which connect to the scope plates.

W6GDO reports "beautiful textbook patterns" with this system, on all bands from 50 through 450 Mc. A single balun will probably serve for both 144 and 432, because of the third-harmonic relationship, if the coupling in the T fitting is right for the two bands. Several T fittings can be modified for various power levels, and baluns can be made for all bands used, in order to provide flexible monitoring with the scope on all voice transmissions.

There are several ways to modify the coaxial fittings. If an 83-1J adapter (through-connector) is used with the 83-1T T fitting, the coupling capacitance can be included in the adapter, just as well as in the T unit. Where the power level is low the balun can be fed from an unmodified T fitting. This is safe for transmitters of less than 5 watts output.

MAKING AND USING R.F. CHOKES

General-purpose r.f. chokes can be bought almost as cheaply as they can be made, but winding your own has its points. If choke efficiency is important, you can probably make a better one than you can buy, and for random applications a few turns of wire on a resistor, or self-supported, may suffice.

Applications

A choke is used to keep r.f. out of a circuit, or in another. For v.h.f. applications something between a quarter and a half wavelength of wire wound up into a coil will do. It should preferably be of small diameter (usually ¼ inch or less) and 3 to 6 times as long as it is wide. In some circuits a carbon resistor will work just as well; maybe better, if there is not a heating problem. Fig. 13-6 shows resistors used for decoupling the plate-power leads of familiar r.f. amplifier circuits. R_1 and R_2 can be anything from a few hundred to several thousand ohms. They can be used for voltage-dropping as well, as in the case of Nuvistor stages that must run at around 70 volts, with a power source that may be as high as 200 or 250 volts.

For decoupling the heater circuits chokes must be used. RFC_1 and RFC_2 can be wound on small resistors, preferably of 10,000 ohms or more. RFC_3 and RFC_4 prevent signal loss to ground, when low-value cathode resistors are used. This is not a critical application; almost any choke will do.

The shunt-feed plate choke in a transmitter amplifier, RFC_6 in Fig. 13-7, is a more critical

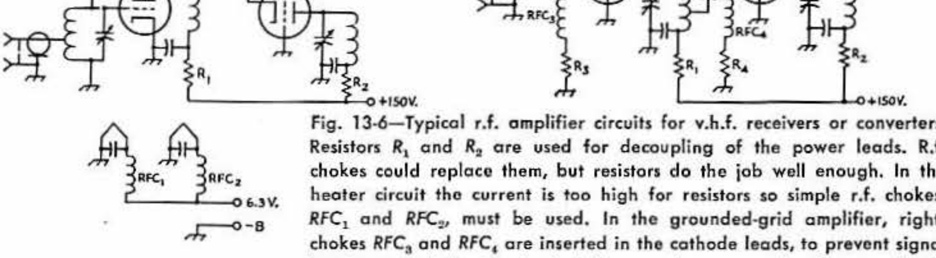


Fig. 13-6—Typical r.f. amplifier circuits for v.h.f. receivers or converters. Resistors R_1 and R_2 are used for decoupling of the power leads. R.f. chokes could replace them, but resistors do the job well enough. In the heater circuit the current is too high for resistors so simple r.f. chokes, RFC_1 and RFC_2, must be used. In the grounded-grid amplifier, right, chokes RFC_3 and RFC_4 are inserted in the cathode leads, to prevent signal loss to ground. Heater circuit chokes are the same as for the cascode circuit at the left. The quality of the chokes is not particularly critical in either application.

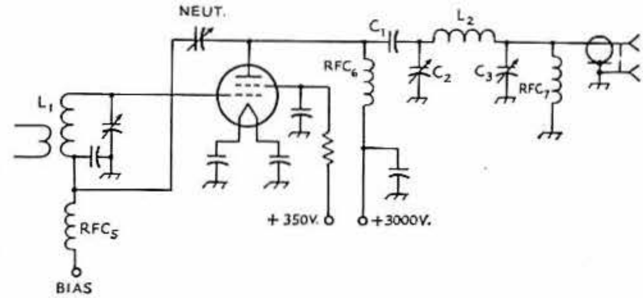


Fig. 13-7—Transmitter applications for r.f. chokes vary markedly in regard to the quality of choke needed. In the grid circuit, RFC_5 has no very difficult job to do, and any choke suitable for low-power use is suitable. The shunt-feed choke, RFC_6, must meet severe requirements, especially in high-powered amplifiers. It is effectively across the transmitter tank circuit, and is subjected to high temperature, current and voltage. The output choke, RFC_7, is mainly a safety device and it operates under much less stringent circumstances.

matter. This choke has the whole r.f. power of the amplifier across it, at high impedance, and it had better be a good one. On the other hand, the output choke, RFC_7, never sees much over 50 ohms impedance, so long as the transmitter is working into a well-matched load. Its quality is not a matter for great concern, and it is mainly a protective device, to prevent d.c. voltage from appearing on the antenna line, should C_1 break down. The grid choke, RFC_5 is not particularly critical.

Placement of an r.f. choke may have considerable bearing on its performance. Fig. 13-8 shows right and wrong positions for the r.f. choke on a 2-meter plate line. The pipe-line amplifier of Chapter 6 was built originally as shown at the left of Fig. 13-8. When power was applied the r.f. choke went up in smoke. Moved to a position outside the "U" of the plate line its replacement has run coolly ever since.

There is no "good" place to put a choke used in the manner of RFC_6 in Fig. 13-7. It has to be close to the tube, so it is subjected to considerable heat, as well as to high r.f. voltage and heavy d.c. flow. Consequently this r.f. choke must have large built-in safety factors in all categories.

Designing for the Job

To handle the d.c. load without overheating, No. 28 wire is about as small as it is safe to go in an r.f. choke for heavy-duty transmitting applications. Larger is better, if there is room. Space-winding the turns increases the heat-dissipating qualities, making it possible to use smaller wire than when the turns are closewound. Most heat trouble in r.f. chokes develops from their being used in hot places, and being subjected to high r.f. voltages, rather than to excessive d.c. flow alone.

Wire size is important in heater chokes, especially where the current to several tubes runs through a single choke. No. 22 or 24 wire is about as small as should be used in heater leads, ordinarily. These or larger sizes can be used for self-supporting chokes for the higher bands.

Distributed capacitance limits the range over which an r.f. choke will work. This makes the space-wound choke superior to the close-wound one. A minimum of cement on the windings

TABLE 13-II

R.f. Chokes for 50, 144 and 220 Mc. service.

Frequency	*Inductance*	*Description*
50 Mc.	7.8 to 9.5 μh.	B & W Miniductor No. 3004, 1⅜ to 1-9/16 inch long.*
50 Mc.	8.3 μh.	No. 28 d.s.c., spacewound on ½-inch Teflon rod. Winding 1¾ inch long. See text.
50 Mc.	7.2 μh.	No. 28 d.s.c., closewound on ¼-inch Teflon rod. Winding 1-7/16 inch long.
144 Mc.	2.15 μh.	No. 22 Nyclad, closewound 1-3/16 inch on ¼-inch Teflon rod.
144 Mc.	1.42 μh.	31 turns No. 28 d.s.c., spacewound on ¼-inch Teflon rod. Winding 1-1/16 inch long.
144 Mc.	1.3 μh.	29 turns No. 22 Nyclad 1⅛ inch long, ¼ inch diam. self-supporting.
(Above 144-Mc. chokes work well on 220 Mc.)		
220 Mc.	0.6 μh.	13 turns No. 22 Nyclad on ¼-inch Teflon rod.
220 Mc.	0.75 μh.	17 turns No. 28 d.s.c. spacewound on ¼-inch Teflon rod. Winding ⅝ inch long.
220 Mc.	0.52 μh.	22 turns No. 22 Nyclad closewound on No. 24 drill, self-supporting.

*Excellent for use except where high temperatures are involved.

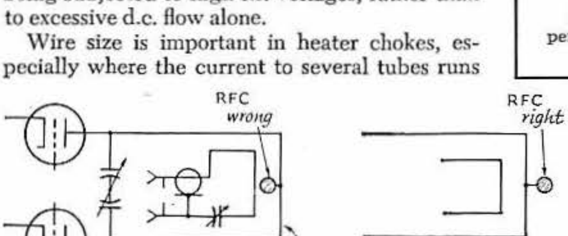


Fig. 13-8—How a choke is positioned with respect to other circuits may be important. The choke at the left is coupled to the plate line of the transmitter tuned circuit. Outside the loop, as at the right, makes the choke far less subject to r.f. breakdown.

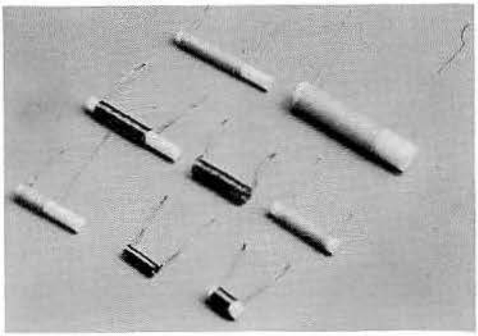


Fig. 13-9—Typical handmade v.h.f. chokes. At the rear are closewound and spacewound chokes for 50 Mc., wound on ¼-inch and ½-inch Teflon rod, drilled and tapped for end-mounting. Three 144-Mc. chokes are seen in the center row, the two at the left being excellent for high-current applications. Similar types to these, but for the 220-Mc. band, are in front.

is also desirable. The space-wound 50-Mc. choke in Table 13-II, shown in the upper right of Fig. 13-9 is as good as you can make for that band, and better than most chokes you could buy. It is good at 144 Mc. as well, and even serviceable at 220 Mc. A closewound choke of fine wire, heavily doped with lacquer, might be usable on only one v.h.f. band, and very likely it would not be too good, even there.

Making Your Own

To set up in the r.f. choke business we need some wire: No. 22 enamel (Nyclad or Formvar preferred); No. 28 enamel, silk or cotton-covered; and No. 30 or 32, of any similar insulation. Silk or cotton-covered wires take cement nicely, but enamel is OK otherwise, and it is usually most readily available.

High-value ½ or 1-watt resistors make good winding forms for use at 144 Mc. and higher. A 2-watt resistor is big enough for a 50-Mc. choke, but Teflon or Nylon rod stock is better. Do not use polystyrene or lucite, if any heat is to be involved. These materials will melt in the heat of an average transmitter enclosure. Teflon rod can now be found in plastics supply houses, in ¼ and ½-inch diameter. It drills and taps nicely, it won't melt, and its insulating quality is excellent. Bakelite rod or even wood dowelling is good enough for the less-critical choke applications. The smallest-diameter prepared coil stock is usable for 50-Mc. chokes, but it won't stand much heat.

Space-winding r.f. chokes is easy. First drill through the rod at spacings indicated under *winding length* in Table 13-II. Now measure off slightly more than a *half* wavelength of wire. Double it back on itself and feed the end through one of the holes in the rod. Now wind the coil as if it were to be bifilar. If you clamp the other end of the double wire in a vise, or tie it down firmly otherwise, this can be done easily. Keep the wires under tension, and be sure that they are not twisted at any point. Wind tightly and then feed the end through the other rod hole.

Now remove one of the wires by unwinding carefully, keeping it under tension throughout. The remaining wire will be space-wound as neatly as if done by machine. Apply a thin coating of polystyrene cement, using a bit more around the lead holes, and your choke is done. It will be dry and ready for use in a few minutes. If having all those wire scraps left over runs against your Scotch instincts, make chokes for the lower end of the range first. The pieces unwound will be useful for higher-frequency production later.

Self-supporting chokes of excellent quality can be made by winding No. 22 or 24 wire tightly on various drill sizes, and then slipping the drill or other winding form out. If wound under tension the coil will hold its shape when slipped off the form. Turns can be spaced by running a thin knife blade between them. You can't make a better choke than this.

You can tell a good choke from an inferior one easily enough. Connect it across your driver-stage tuned circuit, and see what it does to your final-stage grid current. Also note how much you have to retune the driver circuit to restore resonance. A perfect choke would have no harmful effects, and it would not heat up. You won't find one that good, but a well-designed choke will come close. If the choke is not a good one, don't run the test too long at any appreciable power level, or you won't have to *look* for indications—you'll smell them!

A recent development in the r.f. choke field is the ferrite bead. These are small beads of ferrous materials that can be slipped over wires wherever the effect of an r.f. choke is needed. They are particularly effective for heater decoupling purposes, as they can be placed directly on the heater lead, adding no d.c. resistance to the circuit.

"JUST LIKE QST, EXCEPT . . ." *

These words are voiced or written almost daily in telling the owner's sad story of a rig that doesn't work as it should. Investigation nearly always shows that the builder made his own troubles, as a result of common misconceptions about bypassing and grounding in tetrode or pentode amplifiers, transmitting or receiving. Every new project is to some extent a design problem, and a certain amount of debugging is almost inevitable, but there are right

* From a longer article under the same title, in March, 1959, *QST*.

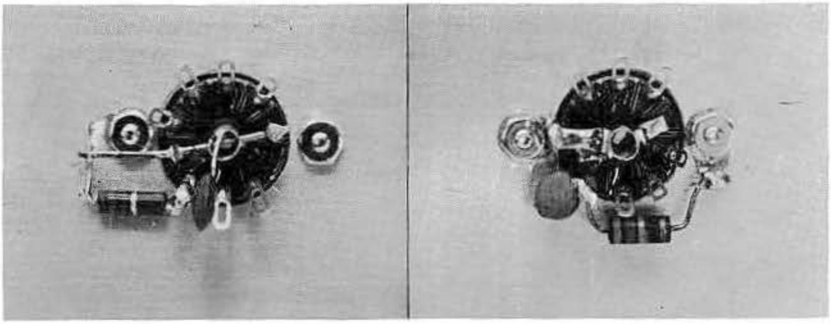


Fig. 13-10—Models showing right and wrong methods for bypassing and grounding. In these 9-pin sockets, Pins 4 and 9 are supposedly grounded, and Pin 3 is hopefully kept at r.f. ground potential by the disk ceramic bypass. In critical v.h.f. circuits, the layout at the right stands a much better chance of achieving these objectives. Common coupling, and resultant instability, are inherent in the arrangement at the left.

and wrong ways to do the basic jobs encountered in amateur equipment construction.

Take the matter of sockets and bypassing. Fig. 13-10 shows wrong and right methods. The wrong, left, has Pins 4 and 9 supposedly grounded by a small wire running through the socket center ring (which is supposed to be a shield) to a lug under the mounting screw at the left. Pin 3 is bypassed to "ground" by running the low side of a disk ceramic capacitor to the center ring. What happens here? The only place that is really at ground potential is the portion of the lug that is under the mounting nut. The rest of the lug, the wire, the socket pins and the bypass lead could all be above ground for r.f. The path to ground from Pin 3 is a long one, and the other pins are common to it. This is an invitation to feedback, due to common-lead coupling.

Now look to the right. The same circuit is wired by bending the socket lugs against the center ring and soldering them there. The lug under the mounting nut is also bent over to the center ring. This puts Pins 4 and 9 very much closer to ground potential than the wire method at the left. The capacitor from Pin 3 is grounded to the lug, right close to the point where it is under the nut. If the disk ceramic is capable of being an effective bypass, this wiring arrangement will give it a chance. There is almost no common path to ground between Pin 3 and the other circuits, and the center ring is much nearer to being a shield than before.

Another demon for the v.h.f. man is the socket shown at the left of Fig. 13-11. The manufacturer very kindly provided an elevated mounting ring on this gem. Fine for use on lower frequencies, or in a hi-fi amplifier, perhaps, but it caused all manner of trouble in a 6146 amplifier for 50 Mc. built by the author of this book.

This socket makes reliable contact to the chassis only at the mounting ears. If you accept the maker's invitation to use those ring extensions for the cold (you hope!) end of your bypass capacitors, you can very easily build in common coupling through the ring's inductance. The socket at the right, with no such conveniences, forces the user to bypass to ground, at the mounting nut, which is the right place to do it.

Our best tetrodes and pentodes, both transmitting and receiving types, have very high gain and power sensitivity. Only a little bit of feedback can make them take off. If this feedback is within the tube itself, or if it is due to coupling between tuned circuits, you can neutralize or shield it out. If the coupling is built in through common ground leads and ineffective bypassing, as illustrated in these typical examples, it can grow you a lot of grey hairs.

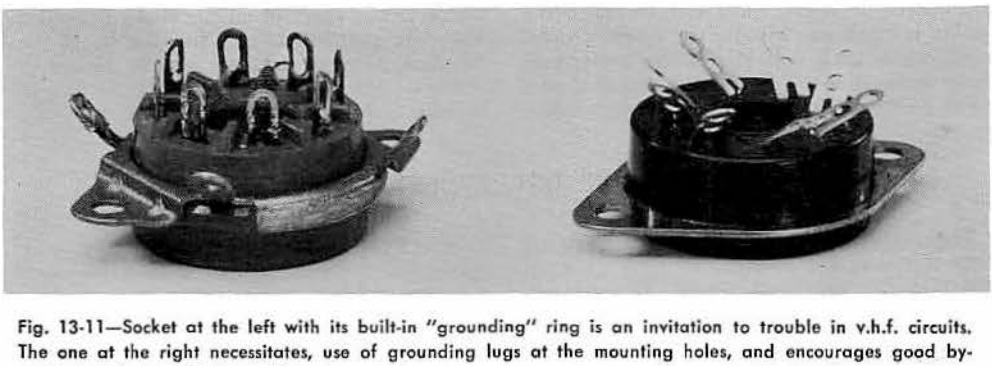


Fig. 13-11—Socket at the left with its built-in "grounding" ring is an invitation to trouble in v.h.f. circuits. The one at the right necessitates, use of grounding lugs at the mounting holes, and encourages good bypassing and grounding practice.

MORE VERSATILITY WITH THE HEATH SIXER AND TWOER

The well-known "Benton Harbor Lunchbox" is a mainstay of activity on 6 or 2 meters in many localities. Here are several modifications of these popular little rigs that will add to their versatility. They are the work of Lew McCoy, W1ICP.

Adding A2 for Code Communication

The 50- and 144-Mc. bands are ideal for code communication—for practice or for improved signal readability when the going is rough. Unfortunately very few readymade a.m. transceivers now available provide for code work of any kind, so much interesting potential of the v.h.f. bands is lost to owners of such equipment. The superregenerative receivers in the Sixer and Twoer make it impractical to copy keyed c.w., but they are fine for reception of tone modulation.

The simple transistor tone oscillator of Fig. 13-12 can be built into either unit easily. By connecting the output of the oscillator to the arm of the volume control the keyed tone will modulate the transmitter, and when the transceiver is switched to the receive position the oscillator can be used for code practice. Also, in the transmit position a slight amount of the audio tone is fed to the speaker, permitting the operator to monitor his own "fist."

The tone oscillator is mounted on a 2 × 2¼-inch piece of perforated unclad circuit board. This is 1/16 inch thick and is perforated with 1/16-inch diameter holes spaced approximately ¼ inch apart. Push-in clips are available for making connections, but in the units shown the connections were made by soldering the component leads together.

The emitter of Q_1, and one side each of C_{101}, R_{101}, and R_{102} (and C_{102} in the 2-meter unit) are connected together and a lead run from this connection to the chassis of the transceiver. This provides a common ground for the oscillator. The key jack, J_{101}, must be insulated from the panel, and either insulating washers or electricians' plastic tape can be used for this purpose. The jack is mounted on the panel between the microphone connector and the volume control. The oscillator assembly is supported by its own leads. When installing the board, be careful that none of the connections on the bottom touch any leads in the transceiver.

In order to monitor your own sending, a 330-ohm, ½-watt resistor should be connected between terminal 4 of the transmit-receive switch and the chassis, as shown in Fig. 13-12C. This feeds a very small amount of audio from the transmitter to the speaker. When transmitting A2 turn the volume control full on; otherwise, the audio oscillator output will be short-circuited to ground. For receiving or using the oscillator for code practice, the volume control should be set at a comfortable listening level. A switching circuit could be used so that the volume control setting wouldn't have to be changed, but this would have complicated the conversion and didn't seem worth the expense or crowding of components.

Metering Transmitter Output

One problem with the Twoer and Sixer is that external meters are required for tune-up, and there is no constant metering of the output. A low-cost milliammeter connected as a relative output indicator can be installed in each unit,

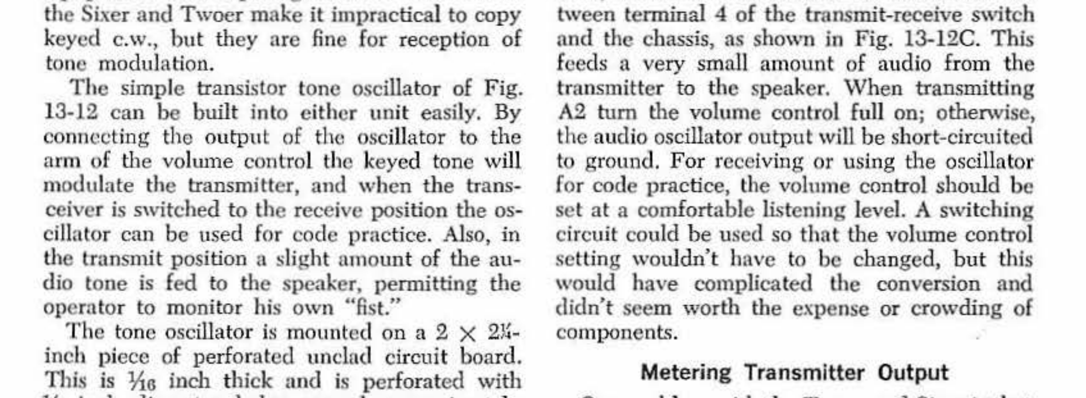


Fig. 13-12—Circuit diagram of the code oscillator. A is the 144-Mc. unit and B is for 50 Mc. Fixed resistors are ½-watt; resistances are in ohms (K = 1000) and capacitances are in μf. Capacitors are paper or Mylar, working voltage 25 or more. Component numbers under 100 refer to the original Heath circuit; those over 100 are the added components. C is the circuit for monitoring one's own sending.

C_{101}—0.05 μf., disk ceramic, paper, or Mylar.
C_{102}—0.001 μf., disk ceramic.
J_{101}—Single-circuit phone jack or phono jack.
Q_1—N-p-n, RCA type 40314 or similar.
R_{101}—1800 ohms, ½ watt.
RFC_{101}—2.7 μh. (Millen 34300-2.7 or similar).

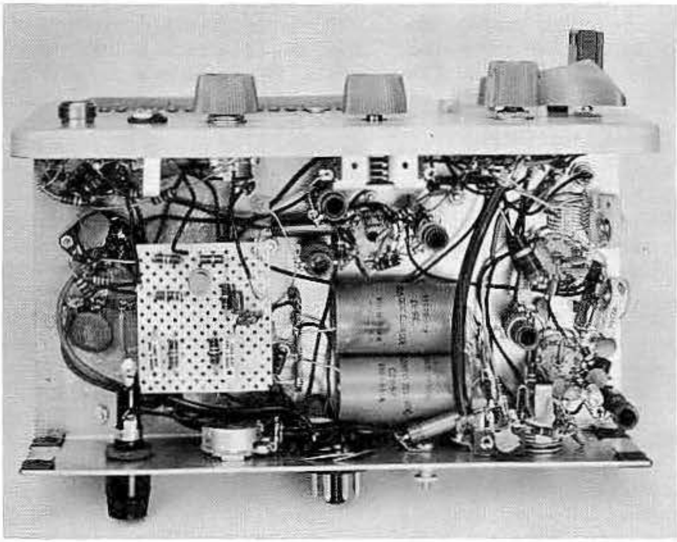


Fig. 13-13—Interior of the "Lunchbox" with the code oscillator installed.

for constant monitoring of the power going to the antenna.

The meters in Fig. 13-14 are edgewise miniature S meters. There is adequate space on the panel for both the meter and control R_{103} just below the nameplate. The ungrounded end of R_{103} is connected to the meter-jack side of R_{13} (a 3300-ohm, ½-watt resistor) by an insulated wire fed under the chassis through a grommet below the meter.

Amplifier Tank Circuit Modification

Another worthwhile improvement can be made by changing the output tank circuit from capacitive to inductive coupling to the antenna. This reduces the possibility that undesired frequencies generated in the multiplier stages will reach the antenna.

Remove the coupling capacitor that goes from the tank coil (L_3 in the Sixer and L_4 in the Twoer) to terminal 11 of the transmit-receive switch. In the 2-meter unit, insert one side of a 3-30 compression trimmer capacitor under the nut that holds the tube socket for V_4, at the chassis-edge side. The new coupling loop, L_{101} in Fig. 13-16, is made of insulated No. 14 or 16 solid wire. The loop for the Twoer is one turn the same diameter as the tank coil, inserted between the first and second turns, at the feed-through capacitor end. One end of the loop is connected to terminal 11 of the transmit-receive switch and the other end to the ungrounded side of the 3-30 compression trimmer. Keep these lead lengths as short as possible.

Using the lamp dummy load that comes with the kit, tune the tank capacitor and the compression trimmer for maximum lamp brilliance. The output meter will read maximum when the lamp is the brightest. It may be necessary to reduce the sensitivity by means of R_{103} to keep the meter from going off scale.

Try moving the loop in relation to the tank coil, for maximum brilliance of the lamp load. Be sure to turn off the power to the transceiver when making this adjustment because the B-plus voltage is present on the tank coil and you

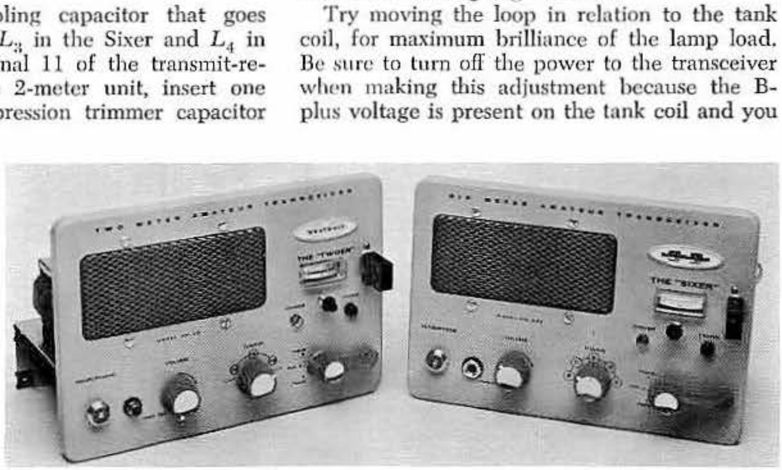


Fig. 13-14—The Twoer and Sixer, complete with code oscillator, relative power meter and front-panel crystal socket.

could get a dangerous shock.

The 6-meter installation is slightly different. The trimmer capacitor is mounted on a 3-lug terminal strip with the center terminal grounded. The strip is mounted between the crystal socket and the socket for V_4, using the unused coil mounting hole as the mounting point. A 2-turn link, with the turns just slightly smaller in diameter than the tank coil, is made from insulated No. 16 or 18 solid wire. This is positioned just inside the tank coil at the feed-through capacitor end. The adjustment procedure is the same as with the 2-meter unit.

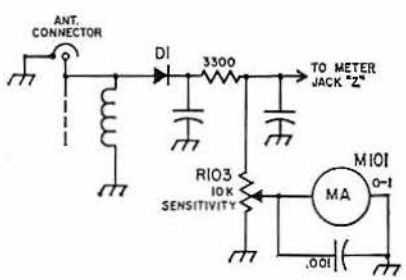


Fig. 13-15—Addition of the metering circuit.

M_{101}—0—1 milliammeter (Radio Shack 22-004, World Radio Labs 99M194).

R_{103}—10,000-ohms, ¼-watt control.

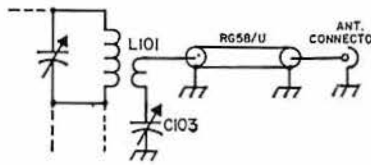


Fig. 13-16—Tuned output circuit for the Twoer and Sixer.

C_{103}—For 144 Mc., 3-30-pf. compression trimmer; for 50 Mc., 8-50-pf, trimmer (Centralab type 822-AN or similar).

L_{101}—See text.

Remove the wire from terminal 12 of the transmit-receive switch and the antenna output terminal. A length of RG-58/U is substituted for this lead, grounding the outer conductor at both ends.

External Crystal Socket

A crystal socket on the front panel makes frequency changing much easier. This mounts on the front panel alongside the meter, and a short length of Twin-Lead, fitted with a crystal socket plug (Millen type 37412), is used to connect it to the chassis-mounted crystal socket. If you have a defunct crystal, it can be removed from its holder and the Twin-Lead soldered to the holder pins, to make a plug.

L-MATCH FOR COAX-FED V.H.F. ARRAYS

Whenever coaxial line is used to feed a driven element directly, there must be provision for converting the unbalanced line (coax) to the balanced load represented by the dipole. Some method of raising the driven element impendance is also required. The simple L-Match of Fig. 13-17 was devised by Ralph Campbell, W4KAE, to do both jobs. Its derivation is shown in Fig. 13-18.

If the question of balance is ignored, an inductive stub of U shape, solid line in the sketch, can be used in conjunction with a shortened driven element, to effect an impedance match between the transmission line and a driven element of lower impedance. Such a stub tried by W4KAE in a 2-meter Yagi having a two-piece dipole fed with coax gave something approaching an impedance match, but left the problem of balance unsolved. Checks with an r.f. probe showed that the portion connected to the inner conductor was hot with r.f., but the other side was practically cold. In effect, the driven element was acting like the fed portion of a ground-plane antenna, with the other half and the metal boom acting like the radials.

When the loop was bent toward the side of the dipole that was connected to the inner conductor of the coax, (broken line in Fig. 13-18) the balance of power in the halves of the dipole improved, and the s.w.r. indication on the line went lower than could be obtained with a perpendicular stub. This led to experiments with a boot-shaped loop, Fig. 13-18B, varying the position of the "toe" with respect to the hot half of the driven element. Soon nearly perfect balance was achieved, and the s.w.r. indication was brought down to 1.05:1. Presumably very careful adjustment of the length of the driven

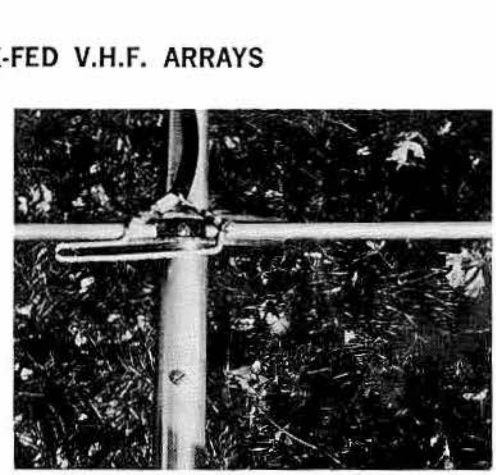


Fig. 13-17—The L-Match as installed by W4KAE.

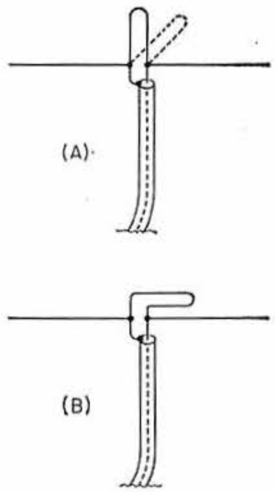


Fig. 13-18—Evolution of the L-Match. Perpendicular loop, solid line, raises feed impedance, but leaves problem of unbalance unsolved. Moving loop to the right partially corrects balance. Boot-shaped loop, B, combines impedance-matching and balun effects.

element, and the length and position of the stub, could bring the s.w.r. down even lower, though checks on many supposedly well-matched 2-meter arrays might show his match to better than most.

The effect of the stub amounts to inductive loading at the center of the dipole, so the end-to-end length must be physically less than that of a driven element fed by other means. The dipole length and the size and position of the balun loop vary with frequency, and with the feed impedance of the array in question, so no one size can be right for all 2-meter Yagis. The dimensions shown in Fig. 13-19 were optimum for a "store-bought" 15-element Yagi, operated at 145 Mc. The mounting screws of the dipole, which serve as connection points for the coax and balun loop, are 2¼ inches apart. The hoop is made of aluminum ground wire, about 9 inches overall, including the "eyes" at each end for slipping over the mounting screws.

Some variation in loop inductance, and thus in impedance matching, can be made by varying the spacing between the upper and lower portions. Balancing effect is related to the position of the loop "toe" with respect to the driven element. Obviously these effects interlock, so a cut-and-try approach is indicated. The array should be set up at least one wavelength above flat ground, with no reflecting objects in the field for many wavelengths out in front. An alternative recommended in our antenna chapters is to point the array straight up while making matching adjustments.

The L-Match should be of workable dimensions for 50- or 220-Mc. arrays. Suggested lengths of wire for making the loops are 25 and 6 inches, respectively, for 50-ohm feed in Yagis of otherwise conventional design.

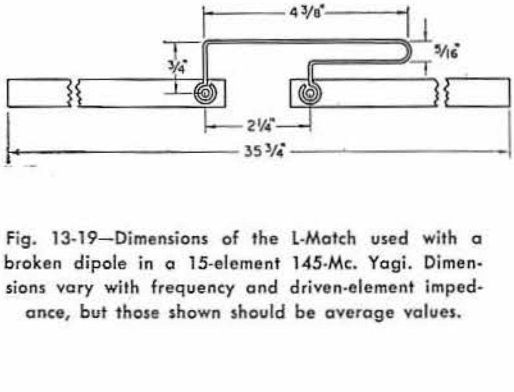


Fig. 13-19—Dimensions of the L-Match used with a broken dipole in a 15-element 145-Mc. Yagi. Dimensions vary with frequency and driven-element impedance, but those shown should be average values.

PILOT LAMP NOISE GENERATOR

Here's a noise generator hint that came from W9EHX, via W9KLR. It uses a 2-volt 60-ma. pilot lamp instead of a crystal diode, but is otherwise similar to the diode job described in chapter 11. The r.f. choke didn't appear necessary but is advisable to prevent loss of noise through battery circuit. Output is constant, in comparison to the rather variable results obtained with some crystal diodes. The original was built in an old soup can from junk parts, which kept the cost close to nothing. The load resistance should be equal to the impedance of line used to feed antenna. It shows about 4.5 db. of noise with a good 144-Mc. converter. More noise could be obtained with increased filament voltage, at expense of shorter lamp life. Other types of lamps may draw more current, but they deliver about the same noise output, so a 60-ma. type is most economical. The coax fitting should match that on converters to be checked. Make all leads as short as possible.

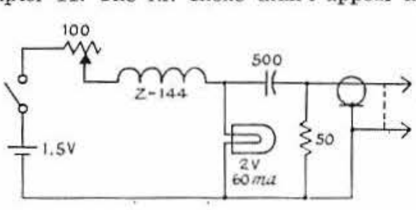


Fig. 13-20—Noise generator using 2-volt 60-ma. pilot lamp, in place of the usual crystal diode.

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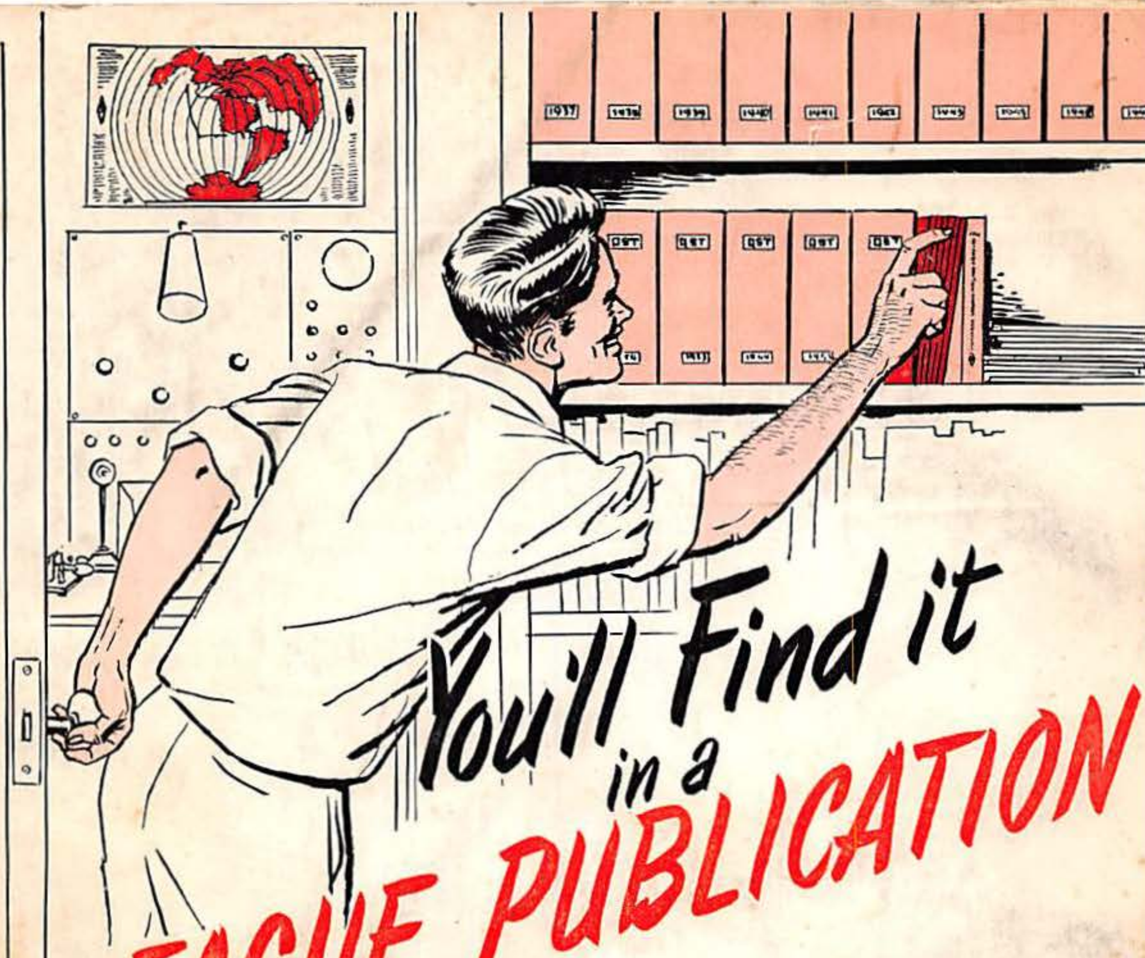

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