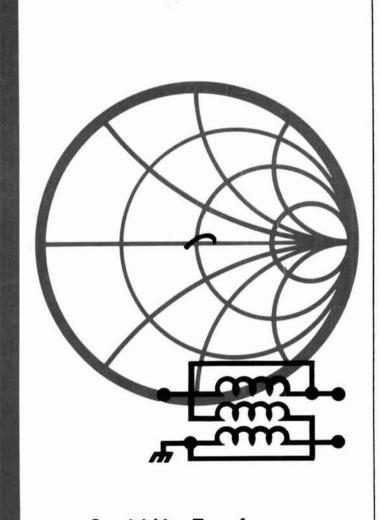


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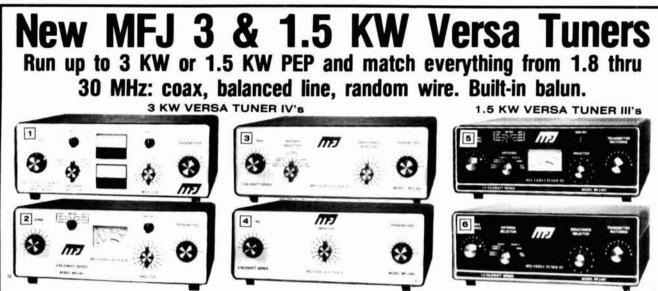
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 Antenna switch lets you select 1 coax thru tuner and 2 coax thru tuner or direct, or random wire and balanced line.
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or direct, or random wire and balanced line. Encapsulated 4:1 balun. If you already have a SWR/wattmeter, the MFJ-982 is for you. and balanced line. One year limited warranty. Made in U.S.A. **MFJ-980 3 KW VERSA TUNER IV**

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5 MFJ-962 1.5 KW VERSA TUNER III



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A versatile six position antenna switch lets you select 2 coax lines thru tuner or direct, or random wire and balanced line. Encapsulated 4:1 balun. Black front panel has reverse lettering.

6 MFJ-961 1.5 KW Versa Tuner III



6 position antenna switch lets you select 2 coax lines thru tuner or direct, or random wire and balanced line.

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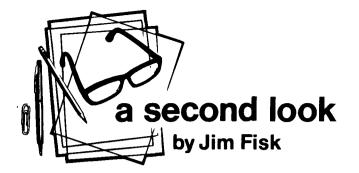
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The 1979 World Administrative Radio Conference (WARC) concluded just as the January issue was going to press, so I was unable to comment editorially on our good fortune in Geneva. It's no secret that I've been cautiously optimistic about WARC for some time, and as I predicted almost a year ago, Amateur Radio remains in possession of the same high-frequency spectrum as we've had for the past several decades; three new HF bands are our bonus for the 1980s. The international Amateur Radio community went to WARC very well prepared, thanks to long years of behind-the-scenes work by the International Amateur Radio Union (IARU), the various national societies, and dozens of Amateur Radio volunteers — these are the same people who remained optimistic about WARC throughout the long years of preparation. The Wizard of Woe, who expected WARC to be a complete disaster and forecast the end of Amateur Radio in the 1980s, was last seen skipping off to the Land of Oz with one foot in his mouth.

I first became aware of plans to pursue new ham bands at 10, 18, and 24 MHz more than five years ago, but quite frankly I considered the possibility of any new high-frequency spectrum as more pipe dream than reality. When W1RU announced the new bands at the Dayton Hamfest in 1974, I think most amateurs thought the pressure for new bands was a ploy to maintain the *status quo*; you know the old game: ask for more than you want so you can keep what you already got. Apparently the movers and doers of Amateur Radio who started preparing for WARC in the mid 1970s thought otherwise.

A. Prose Walker, W4BW, Chief of the FCC's Amateur and Citizens Division in those years was instrumental in arranging for the propagation studies that were conducted to show how long-distance Amateur communications links would be greatly enhanced and made more reliable with new allocations at 10, 18, and 24 MHz; Prose deserves a great deal of praise — and thanks — for getting the WARC preparations underway at such an early date. Without his foresight and experience I don't think Amateur Radio would have survived WARC in such good shape. Amateur Radio is also indebted to the dozens of dedicated volunteers who served on the FCC's Amateur Radio Advisory Committees, often at great personal expense, and to the ARRL and IARU staffers who convinced other national societies to get their WARC preparations underway early and then coordinated those activities.

What about those new bands? What are their propagation characteristics? How soon can we expect to have them available for Amateur Radio use? To answer the last question first, it will be quite awhile before you begin hearing ham signals on these bands; probably 1982 for 10 MHz, and not before 1985 for 18 and 24 MHz. The long delay for the top two bands is because Amateur Radio stations will not be permitted to take possession of these *exclusive* frequencies until all existing fixed services have moved to new assignments; the transition will occur between July, 1984, and July, 1989, so it will be nearly five years at the earliest, and significantly longer if any of the present users feel like dragging their feet.

It may not be obvious, but the frequencies of the three new ham bands were very carefully chosen — each band is very near the geometric mean of the two existing, adjacent bands. This is optimum band placement for maximum propagation enhancement, so it should be possible to maintain long-distance radio communications for many more hours each day than is possible with our present allocations.

Practically all of the new rigs which use phase-locked loops already cover the new bands (although they are not presently programmed to transmit there); some will require simple modifications, but you can be sure the manufacturers are already working on them — and will have mod kits available by the time the new bands are opened. And when the new frequencies do become available, most of the equipment makers will be marketing ham gear with even more features than they offer now! Thanks to WARC, the future for Amateur Radio is bright, and the decade of the 1980s promises to be exciting — both to Amateur Radio and to the technology which will become available to us.

Jim Fisk, W1HR editor-in-chief

The IC-251A is the newest addition to ICOM's all mode transceiver line. Like the matching IC-551, the IC-251A has dual digital VFO's, three memories, scanning (even SSB), and many other features you only get from ICOM. Both units include the no

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Both units include the no backlash, no delay light chopper, similar to the IC-701, as a standard feature at no cost. Coupled to the microporcessor, this provides split frequency operation as well as completely variable offsets.

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IC-551 specifications. IC-251A's specs are identical except where noted (in **bold**).

MIC 347

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- Frequency Coverage: 50~54MHz (143.8~148.19MHz)
- **RF Output Power:**

1441551

IC-251A

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NE YOX LOC

- SSB 10W PEP
- (1~10W adjustable) (10W) CW 10W
- (1~10W adjustable) (10W) AM 4W
- $(0 \sim 4W \text{ adjustable}) (-)$ FM* 10
- (1~10W adjustable) (1~10W)

Sensitivity: SSB/CW/AM Less than 0.5μV for 10dB S+N/N FM* More than 30dB S+N+D/N+D at 1μV Squelch Sensitivity: SSB/CW/AM

-

IC-551

0 0

000

N8 400

FM* 0.4 μV (0.4 μV)

Selectivity: SSB/CW/AM More than ±1.1KHz at -6dB(1.2) Less than ±2.2KHz at -6dB(2.4) (When Pass Band Tuning Unit is installed: less than 1KHz at -6dB)

515553

U

- FM* More than ±7.5KHz at -60B Less than ±15KHz at -60B
- Dimensions: 111mm (H) × 241mm (W) × 311mm (D)

Weight: 6.1kg (5kg)

Spurious Response Rejection Ratio: More than 60dB

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	Please send me: IC-551 specifications sheet; IC-251A specifications sheet; Ltst of Authorized ICOM Dealers.		
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All stated specifications are subject to change without notice. All ICOM radios significantly exceed FCC regulations limiting spurious emissions.



compact loop antenna Dear HR:

W6TC's article on the compact loop antenna in October ham radio essentially describes the antenna I have used for several years. I improvised the antenna out of desperation because my station was over pure sand, and grounded verticals simply did not work. Contrary to the title of his article, I was searching, not for a DX antenna, but a general-purpose antenna for all bands that would work reasonably well in an area with extremely poor ground conductivity. In addition, I wanted an antenna that would cover the entire 75/80-meter band. The antenna has satisfied all these requirements.

The antenna books give the resonant impedance of a full-wave loop as about 105 ohms; measurements confirmed this figure. To cover the 75/80-meter band, the loop was cut for about 3950 kHz (248 feet). A 17 μ H roller coil in series with the ungrounded side at the feed point resonated the antenna at 3550 kHz. The impedance had dropped to about 25 ohms. Fed with RG-8/U coax, the antenna has an SWR never more than 2:1.

The original antenna was in a pentagon configuration. The feed point was near ground, with the roller coil in a weather-proof metal box. The two ends of the loop sloped upwards in a sort of "inclined V" configuration to the first pair of masts, one on either side of the lot. With a maximum height of 20 feet, this antenna was definitely *not* a DX antenna. It was a high-angle radiator, as expected, but on 75 meters it outperformed a dipole at the same height up to distances of 400 to 500 miles. At increasing distance, even an inefficient vertical antenna did better.

In the CW portion of the 40-meter band a series inductance of 9 μ H is required for resonance. This results in rather high rf voltages on the antenna side of the coil, and an SWR of about 2.5:1. Numerous contacts with 1-watt QRP CW on 40 meters, in the middle of the day, at distances of several hundred miles and with consistent reports of S6-S9, indicate excellent performance.

At my present station, the antenna has been erected in a delta-loop configuration with about the same results. Feeders have been dispensed with (except for 10 feet of coax from transmitter to tuner). Romex wire carries the loop ends through the ceiling of shack up to the peak of the roof, and through the gable. (Since I'm in a remote area, I can get away with this. However, I can run a TV set on rabbit ears in the shack — except on 40 meters.)

> W.S. Skeen, W6WR Hornbrook, California

W3VT memory keyer Dear HR:

A builder of the deluxe memory keyer featured in the April issue has written, pointing out that under some conditions a slight click occurs in his unit during manual sending each time the keyer clock control line goes high. This click was also present in the original version, but was barely perceptible in the monitor and did not seem objectionable; perhaps in some other units it might be severe enough to be a problem. Following is an explanation and the solution.

The click occurs during manual sending when sending from memory has been stopped at a memory location with a high data bit: this can happen when the memory is stopped by the stop button or interrupted by manual sending. The memory output control U13B is held high by the low on the keyer control line at U13B pin 10, inhibiting the output even though the other three inputs (pins 9, 12, and 13) are high. When the control line goes high, U3B inverts the control line signal and applies the low to U13B pin 9, which continues to inhibit memory output. There is a momentary gap in the inhibition process, however, due to propagation delay in U3B.

The cure is simple and requires only slight surgery on the memory board.

First, remove the keyer clock control lead from the designated eyelet on the board. From this eyelet, connect a 0.047- μ F (or 0.1- μ F) capacitor to the nearby ground foil. Next, remove the jumper which connects the clock control line to pin 9 of U4, and replace the jumper with a 150-ohm resistor. Connect a new jumper from the bottom end of this resistor (near

(Continued on page 74)

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VHF, UHF, AND MICROWAVE ALLOCATIONS for the Amateur Service by the World Administrative Radio Conference (WARC) are just as impressive as those below 30 MHz. As outlined here last month, all present VHF and UHF bands remain essentially unchanged for U.S. Amateurs, though the replacement of Radio Location by the much more aggressive Fixed and Mobile Services as co-Primary sharers of 220-225 MHz signals a problem with that band. The addition of an Amateur Secondary allocation at 902-928 MHz is a positive move, and an Amateur slot in that band could become available when the WARC accord becomes effec-tive on January 1, 1982.

tive on January 1, 1982. <u>At Microwave</u>, Amateur Radio did lose 1215-1240 MHz but kept 1240-1300 with a satel-lite uplink subband at 1260-1270 MHz. The 2300-2450, 3300-3500, 5650-5925, and 10000-10500 MHz bands all remained Amateur Secondary as before, but with new subbands on all four for Amateur Satellite use. 24-24.25 GHz continues Amateur and Amateur Satellite, but we now have worldwide allocations at 47-47.2, 75.7-81, 119.98-120.02, 142-149, and 241-250 GHz with a good portion of most, Amateur and Amateur Satellite Primary. <u>Morse Code Ability</u> will still be required of Amateurs worldwide, but only below 30 MHz instead of 144 MHz as before. Another regulation change provides for the use of the 3.5, 7.0, 10.1, 14.0, 18.068, 21.0, 24.89, and 144 MHz Amateur bands for emergency communica-tions between both Amateur and other services during times of natural disaster. Resolu-tions CR prohibits broadcasters from using 7.0-7.1 MHz, with more success, it's hoped.

tion CR prohibits broadcasters from using 7.0-7.1 MHz, with more success, it's hoped, than similar previous attempts. Amateur Radio Came To Geneva as one of best — if not the very best — prepared of all

Amateur Radio Came To Geneva as one of best — If not the very best — prepared of all the services, and the results surely show it. Amateur Radio's tremendous success at the WARC reflects more than five years of effort by the IARU, national societies, and individ-ual Amateurs throughout the world. Special credit goes to the IARU's WARC delegation; G2BVN, G5CO, and SP5FM, representing Region 1, HK3DEU and YV5PBG, Region 2; JAINET, ZL2AZ and 9VIRH, Region 3; and VE3CJ, W1RU, K1ZZ, W4KFC, WA6IDN, and WØBWJ, IARU Headquarters. A heartfelt thanks from all Radio Amateurs.

AMATEUR RADIO'S STRONG WARC showing was due in no small part to the U.S. government's very positive pro-Amateur Radio position, which in turn was the result of a well organized and hard working FCC Advisory Committee on Amateur Radio (ACAR). About 70 U.S. Amazed and hard working FCC Advisory Committee on Amateur Radio (ACAR). About 70 U.S. Ama-teurs participated in the ACAR meetings, and their efforts are deserving of recognition. The ACAR rostor reads: K1CCL, W1HDQ, W1HGI, K1HZN, W1JR, W1NLB, W1NXY, W1RU and K1ZZ; W2ALS, W2DEO, W2DU, W2ECH, W2GHK, W2HD, WA2INB, K2MGA, and W2OD; W3ASK, W3BE, K3BRS, W3BLC, K3DC, W3DXA, W3FU, W3GGO, W3ICM, WA3IHW, W3JPT, W3KMV, WA3LND, W3OKN, W3PK, W3PS, W3SWD, W3VW, WA3WRS, and W3XO; W4AAW, WA4AJF, W4BRB, W4BW, K4BZ, K4BZF, W4FZ, W4GF, K4GTS, W4JOP, K4KDY, W4KFC, W4LVC, WB4LWX, W4MB, W4NSP, K4NSS, W4OO, K4VOW, W4ZC, and W42M; W5EYB; W6DW, W6EJK, W6GD, W6GNX, WA6GYD, K6HCP, K6HIJ, W6ISQ, WB6JPI, W6KAP, K6LFH, WB6NMT, W60HQ, W6SAI, W6UF, K6UQH, K6VGP, and W6ZH; KB9FO, W9JUV, and W9KOI; WØAR, WØBWJ, and WØIUF; and KH6IJ. The efforts of these Amateurs, often at considerable personal sacrifice, should be appreciated by Amateurs worldwide. Thanks K4NSS for invaluable sacrifice, should be appreciated by Amateurs worldwide. Thanks K4NSS for invaluable assistance in preparation of this list, for which W9JUV takes sole responsibility for errors and omissions. Any errors will be cheerfully corrected in next month's <u>Ham Radio</u>. Current callsigns of ACAR members, if known, were used in the list.

TYPE ACCEPTANCE TESTING of solid-state linear amplifiers for the Amateur Service has been reinstated, at least for a trial period through July 1st, after FCC's unexplained termination of the program in late 1978, which drew heavy fire from Ham Radio editor-inchief. WIHR. Ten Tec's new Model 444, the Hercules kilowatt linear, is being resubmitted and could be available within the next several months.

"VOLUNTEER AMATEUR EXAMS ARE ILLEGAL!", the FCC's General Council declared at a Commission meeting in late December. The bombshell announcement came as the Commissioners discussed terminating Docket 20679, the Notice of Proposed Rule Making that would have required volunteer examiners to submit photocopies of their licenses when requesting an The judgment appears to stem from recent Congressional concern that some governexam. ment agencies were improperly expanding their staffs without approval of Congress, by use of volunteers and contractors. When the legal staff reviewed the volunteer program in of volunteers and contractors. When the legal staff reviewed the volunteer program in connection with Docket 20679, they concluded it is illegal and thus must be terminated.

What Effect Termination will have on existing training programs and U.S. Amateur growth, not to mention the added burden on FCC's Field Offices, remains to be seen. For the first three years of the Novice program its exams were all Field Office administered; responsibility for Novice exams has rested with the Amateur Service since June, 1954.

A SUCCESSFUL LAUNCH of the Ariane LO1 launch vehicle was carried out December 24th, from Korou, French Guiana, after a 24-hour delay due to bad weather. Launch took place at 17152, successfully putting a ballast payload into a highly elliptical Earth orbit similar to the initial orbit expected for the AMSAT Phase III-A spacecraft. <u>The Test Proved</u> to be an excellent dry run. Both the launch vehicle and the communi-cations procedures to be used by the European Space Agency during and immediately after a launch attempt worked well. The test clears the way for the LO2 mission, scheduled for

late May, 1980, which will carry the Phase III-A satellite.

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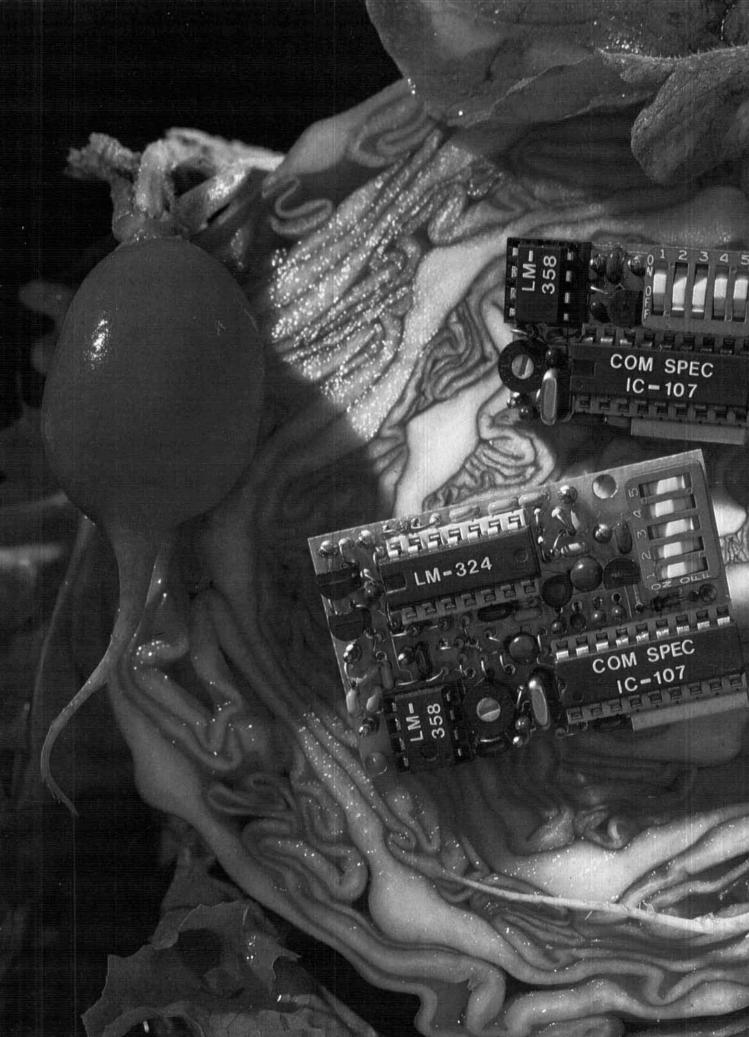
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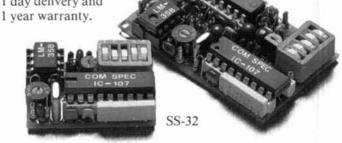




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77.0 XB	100.0 1Z	131.8 3B	173.8 6A	
79.7 SP	103.5 1A	136.5 4Z	179.9 6B	
82.5 YZ	107.2 1B	141.3 4A	186.2 7Z	
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600 1000 1500 2175 2805	697 770 852 941	1209 1336 1477 1633	1600 1650 1700 1750 1800	1850 1900 1950 2000 2100	2150 2200 2250 2300 2350	2400 2450 2500 2550

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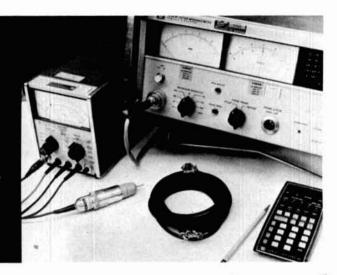
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a new class of coaxial-line transformers

A review of transmission-line transformers and balun theory, including problems with magnetic cores — Part 1 of a two-part series

Most coaxial-fed antennas require a balun for optimum performance; many also require a matching transformer. Typically, these baluns and transformers are made with magnetic core material such as ferrite. These devices are subject to arcing and linearity problems. Simple baluns and a new class of rf transformers which are not subject to these problems use only coaxial line in their construction — and they are easy to make.

To match the low impedance of two closely spaced dipoles, I needed a broadband 4:1 transformer for some experiments on a low-band phased array. My first inclination was to go to the handbooks to design a ferrite-core transmission-line transformer. However, I decided that there had to be a better way. I had just read Doug DeMaw's article, "The Whys and Hows of Bifilar Filament Chokes" in *QST*.¹ He expressed concern about saturation of the core material and corona from the windings to the core when operated at high power. My search began for a way to make broadband transformers without magnetic core material.

background

In conventional low-frequency transformers, closely coupled primary and secondary windings of the appropriate turns ratio are used. At radio frequencies, because of inevitable leakage reactance, narrowband tuned transformers are generally used for impedance transformation. Quarter-wave matching transformers may be used but are also narrowband; they are a quarter wavelength at only one frequency.

The availability of solid-state rf power devices with their capability for broadband performance created the need for broadband interstage matching, thus causing rapid development of transmission-line transformers.^{2,3,4} Development of ferrite materials also expanded rapidly during this period.

The low-frequency response of transmission-line transformers is limited by winding inductance, and the high-frequency response is limited by resonances from stray capacitance; therefore, ferrite material is used to extend the low-frequency limit of small transformers by increasing inductance. Thus broadband transmission-line transformers and baluns using ferrite cores have come into wide use in solid-state rf circuitry and Amateur antenna systems. While these cores are very useful, they have some disadvantages.

By George Badger, W6TC, 341 La Mesa Drive, Portola Valley, California 94025 This article shows how to build and design broadband rf transformers and baluns without magnetic cores.

problems with magnetic cores

Amateurs build or buy highly linear SSB equipment and effective lowpass filters to avoid TVI. We then subject our clean, harmonic-free signals to the uncertainties of ferrite-core transformers or baluns in our antenna systems. The cores in these devices are subject to saturation and, therefore, nonlinearity. High permeability ferrite cores are also susceptible to permanent damage at flux densities of a few hundred gauss.⁵ Tune up your linear into the wrong antenna just once and the damage is done.

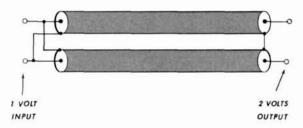


fig. 1. Broadband transmission-line transformers are made of two or more transmission lines connected in parallel at one end and in series at the other. One volt applied to two coax lines in parallel at the input results in 1 volt across *each* of the lines at the output. If these two lines are connected in series at the output as shown, the output will be 2 volts. In this way a 1:4 impedance stepup is achieved. Sufficient impedance must be provided over the length of the outside conductors to prevent the connections at one end from shorting the other end.

Magnetic materials such as ferrite, powdered iron, and specialty steel tapes have added greatly to the performance of components available to circuit designers. However, these materials should not be used in high-power circuits or antenna systems unless they are adequately characterized regarding power-handling capability and saturation effects. This is necessary so that interaction of the material with your system can be thoroughly understood. Put another way, sufficient core material must be used to keep the flux density well below the saturation level. Data on harmonic distortion measurements, taken at high power on a popular commercial ferrite core balun, are presented in part 2 of this article.

Ferrite baluns and transformers are usually wound with copper wire coated with thin enamel insulation. Pairs of wires are placed close together or twisted to make transmission lines, which are wound tightly onto the core. The conductors must be close, because the surge impedance of the wire pairs must be correctly related to the impedances to be matched. For this reason, thin insulation is often used. Inadequate insulation may result in arcing between wires or to the core when operated at high power.^{1,2} Those who have blown a balun in the heat of competition are all too familiar with these problems.

transmission-line transformers

Basically a transmission-line transformer consists of two or more parallel lengths of transmission line connected in parallel at one terminal and connected in series at the other (**fig. 1**). For example, if two lengths of coaxial line are connected in parallel at the input and 1 volt is applied, 1 volt appears at the output end of each of the two lines. If the output ends are connected in series so that the two voltages add, the output is 2 volts, thus creating a 1:2 voltage increase (1:4 impedance transformation). **Fig. 1** also can be used to describe a 4:1 impedance reduction; for example, from 50 ohms to 12.5 ohms.

Sufficient rf impedance must be provided between the input and output ends of the transformer of **fig. 1** to prevent the connections at one end of the lines from shorting the other end of the lines. The impedance is usually provided by wrapping the transmission lines around magnetic cores.

the Collins balun

By far the best balun I've ever used is the Collins balun which, to my knowledge, was first described in a book published by the Collins Radio Company entitled *Fundamentals of Single Sideband*.⁶ The Collins balun derives its name from this reference. I believe the earliest reference to an Amateur application was in an article by K2HLT in *G.E. Ham Notes* in 1960.⁷ The Collins balun is rarely mentioned in Amateur literature, which is surprising in view of its superb performance. However, Bill Orr, W6SAI, describes one in his *Radio Handbook*.⁸

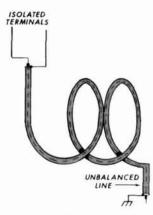


fig. 2. Simple coiled length of coaxial line isolates output terminals from ground.

Perhaps the reason the Collins balun hasn't gained popularity with Amateurs is that it's quite bulky when made with RG-8/U. The balun is extremely simple. No exotic materials are used in its construction; only coaxial cable and insulated wire. I've used these baluns for years with various antennas and never had a failure. One has been on my three-element 10-15 meter quad for eight years with no sign of deterioration. There are only two disadvantages to the Collins design: 1) when made with RG-8/U, the balun is bulky — too large for installation on a clean-design antenna system; and 2) the balun is useful only at 50 ohms. This article shows how to eliminate these disadvantages.

balun theory

Baluns convert energy from unbalanced coaxial line to balanced two-wire line by isolating the two balanced terminals from ground. As in the transmission-line transformer, this is often accomplished by coiling transmission lines around magnetic material so the impedance to ground from both output terminals is high compared with the characteristic impedance of the input coaxial line. By using this technique, shown in fig. 2, the two balanced terminals are "floated" with respect to ground by the isolation provided by the coiled-line impedance. However, a simple coiled length of transmission line is often not adequate because it doesn't contribute to the balance of the system.⁹ For a balun to make this contribution, the impedance ground from both terminals must be nearly matched.

Accordingly, in the Collins balun, a dummy length of coax is wound as a continuation of the isolating winding, so that the coil consists of the original length of coiled coax of **fig. 2** plus an equivalent length of dummy line, as shown in **fig. 3**.

The dummy-line center conductor is unused and is left floating, or both ends may be shorted to the outer conductor if desired. The dummy length of line causes the impedance to ground, from each of the two output terminals, to be nearly equal. The isolation impedance (common-mode impedance) is held higher than the coax-line characteristic impedance over a wide frequency range by the distributed capacitance and inductance of the combined coil. The coil must have sufficient inductance so the impedance, at the lowest operating frequency, is higher than the line surge impedance. As the frequency is increased, the impedance increases through parallel self-resonance, then decreases as the frequency is further increased.

Because the self-resonant circuit consisting of the distributed capacitance and inductance of the combined coil is loaded by the low characteristic impe-

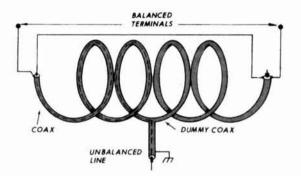


fig. 3. Broadband coil balun is evolved from the coiled coaxial line of fig. 1.

dance of the line, the impedance versus frequency curve is broad. Balun performance therefore is not critical with respect to frequency. Data taken on measurements of the common-mode impedance on a typical Collins balun are presented in Part 2.

The symmetry provided by the dummy line makes balun performance less dependent on common-mode impedance and is therefore often essential in baluns and balanced systems.⁹ The isolation, balance, and impedance match of this class of balun are superb over the hf Amateur bands. Specific designs, performance data, and a systematic design procedure are presented in Part 2.

new class of transformers

Faced with the need to match a very low-impedance antenna system, I decided to try to develop a 4:1 transmission-line transformer based on the principles of the well-proven Collins balun. The transformer was successfully developed; in fact, a new class of wideband transformers evolved from this work.

One of the nice things about an avocation — as compared with a vocation — is that you're not on a time schedule. I found that the performance of the 4:1 transformer was so good that the idea of other transformer designs based on the same principles looked interesting. I shelved the phased-array project long enough to enjoy the freedom to explore the possibilities of these transformers. The result was a series of broadband balanced and unbalanced transformer designs that are extremely simple, made entirely of coax, and, most important, don't depend on ferrite or powdered-iron materials.

design concept

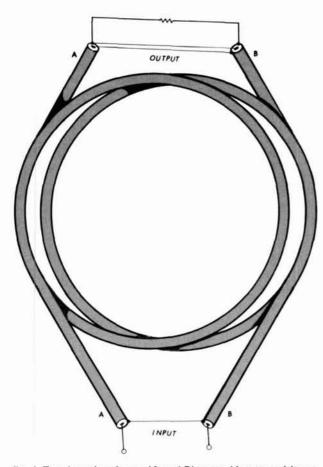
Because the Collins balun so successfully isolates the balanced output terminals from the unbalanced coaxial line input, it seemed reasonable that a similar broadly resonant configuration would provide the isolation necessary to the series and parallel lines of **fig. 1**. From previous experience I'd found that it's unnecessary to wind the Collins balun on a cylindrical form as shown in **fig. 3**. It's sufficient to randomwind the turns without a coil form, as shown in the photo. I decided to try winding the two lines of **fig. 1A** into a continuous winding similar to the Collins balun of **fig. 3**.

My first try was with a nine-turn coil random wound on a nominal 25-cm (10-inch) diameter shown in **fig. 4** (Only three turns are shown in the drawing for clarity.) Two 254-cm (100-inch) lengths of 50-ohm line were used. The transformer was tested by inserting a 12.5-ohm low-inductance load at the output and measuring the input impedance with a Hewlett Packard Vector Impedance Bridge. While the transformer made the 12.5-ohm resistor appear to be a 50-ohm resistor over a wide range, the useful range was centered in the broadcast band. As the frequency was increased through 3.5 MHz, the input

impedance magnitude increased rapidly, and the impedance phase angle was greater than 30 degrees. However, the useful frequency range was 4:1. Encouraged, I removed half the turns, expecting equivalent results more nearly centered over the ham bands. Results were disappointing. A good match was achieved only over a narrow frequency range.

line impedance

Note that the load at the transformer output of **fig**. 4 terminates two coax lines connected in parallel. If the lines have 50 ohms characteristic impedance the load must be 25 ohms for the lines to be properly terminated. At the input the two lines are connected in series, so the input impedance will be 100 ohms. Thus the transformer of **fig**. 4 will match a 25-ohm balanced load to a 100-ohm balanced line, but per-



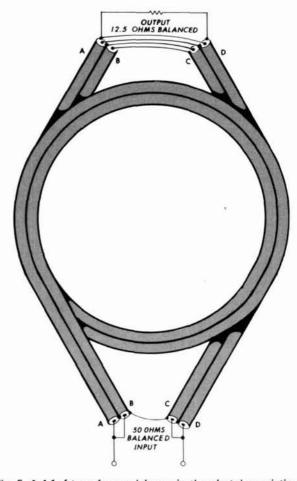


fig. 4. Two lengths of coax (A and B) wound into a multiturn coil similar to the Collins balun of fig. 3. Only three turns are shown for clarity. Note that the input ends of the lines are connected in series and the output ends are connected in parallel. Made with 50-ohm coax, this transformer will match a 100-ohm balanced line to a 25-ohm balanced load.

fig. 5. A 4:1 rf transformer (shown in the photo) consisting of two parallel 127-cm (50-inch) lengths of RG-58A/U random wound into a 7 turn, 11.5 cm (4.5 inch) nominal diameter coil. Each of the two paralleled 50-ohm lines (AB-CD) is connected in series at the input and in parallel at the output. The excellent performance is shown in table 1.

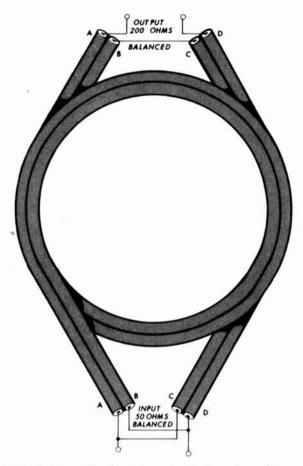


fig. 6. A 50- to 200-ohm balanced-to-balanced transformer made from two 127-cm (50-inch) 100-ohm lines each consisting of two 50-ohm lines in series. The 100-ohm lines are connected in parallel at the input and series at the output. Performance data are shown in table 2.

formance will be poor when trying to match a 12.5ohm load to a 50-ohm line.

50/12.5-ohm transformer

To match a 50-ohm line to a 12.5-ohm load, 25ohm transmission line must be used so that the 50ohm source will be correctly terminated by two 25ohm lines in series. Similarly, the two 25-ohm lines in parallel at the output will be correctly terminated by a 12.5-ohm load. I decided to make 25-ohm lines by connecting pairs of 50-ohm line in parallel. A transformer similar to that in **fig. 3** was wound with two lengths of transmission line, each made with a pair of 50-ohm RG-58A/U lines in parallel (25-ohm line).

After much cutting, winding, soldering, and data taking, the optimum design of **fig. 5** evolved. The design looks somewhat more complex than the simple configuration of **fig. 4** because each length of transmission line consists of two 50-ohm lines in parallel; however, in other respects the configuration is exactly the same as that of **fig. 4**. Note that the two 125-cm (50-inch) 25-ohm lines are connected in parallel at the

output and are properly terminated by the 12.5-ohm load. This transformer design was useful over a wide frequency range. The VSWR is low over the high-frequency Amateur bands. The excellent data are shown in **table 1**.

construction

While it's not necessary to cut the lengths of coax to exactly 127 cm (50 inches), this length wound into a seven-turn, 11.5-cm (4.5-inch) nominal diameter coil is optimum to cover the high-frequency bands. The detail of coil winding is unimportant. You can bind the lines tightly with tape or leave them loose. You can hold the coil in your hands and separate the turns several cm (2 inches) before the magnitude or phase of the match are greatly affected.

However, the length of the short between the coax-line outer conductors at the input end of the balun is critical. Performance is degraded at the high-frequency end of the useful range unless the coaxial outer conductors are soldered together into a common joint as shown in the photo and **fig. 5**. Performance deteriorates if the shorting lead is as long as 2.5 cm (1 inch). The cross-connected leads at the output should be short. In the version shown, the length of the center conductors exposed beyond the outer conductors is about 2 cm (0.75 inch). Construction is shown schematically in **fig. 5** and in more detail in the photo.

measurements

Impedance measurements were made with the Hewlett-Packard rf vector impedance meter model 4815A shown in the photo. The vector impedance meter reads directly in impedance magnitude and phase angle. Data in this article are presented using the abbreviation Z for magnitude of the impedance and θ for the corresponding phase angle. Balance measurements, described in Part 2, were made with the HP rf voltmeter model 4-10C also shown. The terminating resistor is critical to the evaluation of these balun transformers. The VSWR looking into the transformer is, of course, no better than the quality

table 1. Frequency, impedance magnitude, phase angle and calculated VSWR for the balanced-to-balanced 4:1 transformer of fig. 5.

Fo	Z	θ	
(MHz)	(ohms)	(degrees)	VSWR
1.8	50	13	1.26
3.5	51	8	1.15
4	51	6	1.11
7	49	3	1.06
14	54	5	1.12
21	55	1	1.10
28	53	- 1	1.06
30	53	- 1	1.06

of the load. Shown in the photo is a bundle of eight parallel-connected, 100-ohm, 1/4-watt resistors soldered directly to the 50/12.5-ohm broadband transformer for minimum inductance.

performance

Referring to **table 1**, note that the VSWR between 3.5-30 MHz is less than 1.15. Even on 160 meters, the VSWR is only 1.26. Data were recorded only in the ham bands; however, with each configuration, I swept the impedance bridge over the full range looking for spurious resonances. In the designs presented here, none were found within the frequency range of the data shown.

[•] Data are presented in tabular form. Displaying experimental results in this convenient form required calculation of VSWR from the impedance-magnitude and phase-angle data. These calculations are long and tedious, so a Hewlett Packard HP65 programmable calculator was used. N6AIG suggested the idea and wrote the program. The program is very useful and is included in the appendix. The program calculates VSWR based on an impedance of 50 ohms and can be modified easily for any impedance.

50/200 ohm transformers

After the 50/12.5 ohm transformer was optimized, it seemed reasonable to expect that a 50/200 ohm (1:4 impedance stepup) transformer could be made using the same principles. In this case, 100-ohm lines were used so they would match 50 ohms when connected in parallel at the input and would be properly terminated by 200 ohms at the output. The 100-ohm lines were made with two pairs of RG-59A/U cable; each pair was connected in series. The outer conductors of each pair were connected together at both ends. Each pair is 127 cm (50 inches) long. The lengths of coax were wound into a compact package of seven turns at a nominal diameter of 11.5 cm (4.5 inches). The package looks very similar to the 50/12.5 ohm transformer shown in the photo. The transformer is shown schematically in fig. 6. Performance data recorded for this transformer are listed in table 2.

To show the effect of increasing the length of the lines and increasing the number of turns, another similar 50/200 ohm transformer was built. Data taken on this transformer are also shown in **table 2**. The modified transformer was optimized for the low bands. It is similar to the configuration of **fig. 6**. However, the coax lines are twice as long, 254 cm (100 inches), and have twice as many turns on the same nominal diameter. Note that the increased length makes the transformer useful on 160 meters and substantially improves VSWR on 80 and 40

table 2. Data taken on the 50-ohm balanced to 200-ohm balanced transformer. (A) lists data for the transformer made with 127-cm (50-inch) lines as shown in fig. 6. (B) lists data for another version optimized for the low bands. This transformer has the same configuration as that of fig. 6; however, the two 100-ohm lines are made with pairs of 254-cm (100-inch) lengths of RG-58A/U.

Fo	Ζ	θ				
(MHz)	(ohms)	(degrees)	VSWR			
A. Data for transformer of fig. 6						
3.5	56	20	1.45			
4	57	18	1.41			
7	52	7	1.14			
14	51	1	1.03			
21	47	5	1.11			
28	46	15	1.32			
30	46	16	1.34			
	B. Data for low	-band version				
1.8	55	10	1.22			
3.5	55	2	1.11			
4	55	0	1.10			
7	51	- 4	1.08			
10	46	- 2	1.09			
14	46	- 17	1.37			

meters. The low-band transformer performance is good through 10 MHz; however, the VSWR climbs to 1.37 at 14 MHz.

The common-mode impedance of the transformers described in this article is sufficiently high so that the transformers can be driven from either a balanced or an unbalanced coax transmission line. Twostage balun transformers with improved isolation are described in Part 2.

I determined the efficiency of the 4:1 transformer of fig. 5, by carefully measuring the input complex impedance and the complex impedance of the load, driving it with a signal generator and measuring the input and output rf voltages. I used the HP4815A rf Vector Impedance Meter and the HP410C rf Voltmeter shown in the photo. Input power was determined by calculating the power in the real part of the input impedance; the output power was calculated similarly using the complex load impedance data. Efficiency was determined by dividing the output power by the input power. Measurements and calculations were made for each of the bands, from 160 through 10 meters. As you might expect, efficiency was lowest on 10 meters; however, efficiency was greater than 95 per cent on all bands.

power-handling capability

The 4:1 transformers may be made with pairs of RG-58A/U lines connected in series for 100-ohm surge impedance or connected in parallel for 25-ohm surge impedance. Each length of RG-58A/U must therefore handle only 50 per cent of the power delivered by the transformer. Because RG-58A/U is rated for more than 500 watts at 30 MHz, these RG-58A/U transformers will handle 1000 watts of rf. I verified this by connecting two 12.5-ohm transformers back-to-back into a dummy load. At 1 kW, heating was not discernible at 7 MHz. At 30 MHz, the transformers became warm to the touch after about one minute, key down. Part 2 describes how to build baluns capable of several kilowatts overload into severely mismatched loads.

Part 2 of this article, which will appear in March, 1980, ham radio, will describe how to build more useful balun transformers and three specific 1:1 baluns, including one for vhf. These designs are capable of conservatively handling high power into widely varying loads. The impedance match (VSWR) and balance of these new transformers and the baluns will be compared with popular commercially available rod and toroid core balun transformers. In addition, harmonic distortion data taken at 2 kW PEP on a typical commercial ferrite core balun will be included.

Balun performance is usually described when working into a "flat" load. In the real world, baluns must work into widely varying loads as frequency is changed across the band. Measured performance of baluns with varying loads will be included and comparative data presented in Part 2. How to design balun transformers to your needs and how to modify a currently popular balun will also be described.

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appendix

Program for calculating VSWR for a given load impedance, Z_L using the HP-67 calculator.

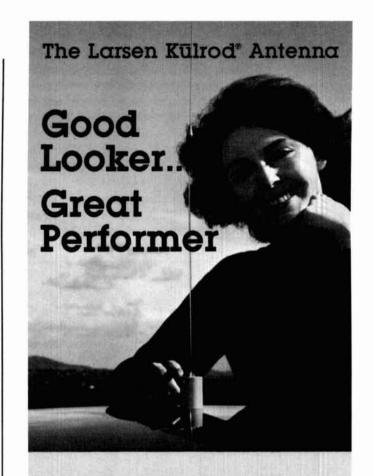
reflection coefficient,
$$\Gamma = \frac{Z_L/Z_0 - Z_L}{Z_L/Z_0 + Z_L}$$

 $VSWR = \frac{I + |\Gamma|}{I - |\Gamma|}$

running instructions

step	instruction	input	keys	output
1	(clear memory - enter program)			
2	initialize		hRTNR/S	_
3	input magnitude	ZL	R/S	-
4	input phase	angleo	R/S	
5	copy VSWR	-	—	VSWR
6	return to step 3		R/S	

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Yagi antenna design experiments confirm computer analysis

Comparison of calculated Yagi performance with NBS measurements confirms validity of theoretical approach

A mathematical model of the Yagi antenna, as described last month,¹ is required for computer analysis of antenna gain, front-to-back ratio, and operational bandwidth. Once the model has been constructed, a computer program can be written. To avoid endless consultations, I prefer not to supply actual computer programs; for those readers who would like to create their own programs, I will explain the information flow I use in programs for computing element currents and Yagi performance.

writing programs

I first create a labeled input file containing all of the necessary information about the Yagi. The label itself is usually an abbreviated reminder of the particular Yagi. This input file contains a statement of the number of parasites and the number of drivers. Next is given information on each parasite, *i.e.*, X and Y coordinates, length (*LE*) and radius (*RO*), all specified in terms of wavelength at a given central design frequency. Next is given information on each driver, *i.e.*, X and Y coordinates, length (*LE*) and radius (*RO*), driving point voltage and phase.

The element-current program first calls for the input data file label and asks at what frequencies (in terms of the center design frequency) the computations are to be made. With this starting information, the input file is called and read. For each frequency specified, a computation is made to determine the (complex) values of all terms in the Z matrix; selfimpedances are calculated (from eqs. 4, 5, and 6, reference 1) and mutuals are given by an interpolation routine or power series approximation (using eqs. 7 and 8).¹

Once the Z matrix is complete and the voltage vector noted from the input file, matrix inversion is accomplished; this generates the individual element complex current solutions. The entire result is then written into an output file for later use. I have found it convenient to include in the output file a statement of frequency for each computation and the number of parasites and number of drivers. For each element, I include its X and Y coordinates, its resonant frequency and Q, and magnitude and phase of its current. The element-current program continues to compute until all initially specified frequencies are satisifed.

This output file now becomes a useful input file to other computer programs which are designed to produce displays of interesting Yagi properties. One such program calculates and plots the pattern both for the H-plane and the E-plane. This is done by reading the new input file, and for a specified sequence of elevation angles (say every 1 or 2 degrees), computing the radiated energy flux density. This is easily done from the known element positions and the known element complex currents; the elevation angle and positions geometrically determine the effective phase delay of each element from the reference origin. The vector sum of all contributions is then proportional to radiated field strength, the square of which is proportional to radiated energy flux density. This calculation provides values which are referenced to a single element carrying a unit current; this can be referred to an isotropic source by correcting for the gain of a single element (2.15 dBi).

This (energy flux density) pattern is calculated for each frequency (read from the new input file) and

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can be easily plotted by one of the plot routines. I believe it is most useful if the pattern is specified in dB relative to an isotropic source,² which I label dBi; such values will be shown throughout this series. In this pattern program I have also found it convenient to compute and display the driver's driving point impedance; the largest energy flux value found in the entire elevation angle search (usually in 1 degree intervals), which I label as MAX H-GAIN (dBi), and the angle at which it occurs, labeled H-ANGLE; and the front-to-back ratio (ratio of energy flux at H-ANGLE to energy flux at 180 degrees minus H-ANGLE) labeled FRONT/BACK GAIN (dB).

Another program of great utility is one which reads

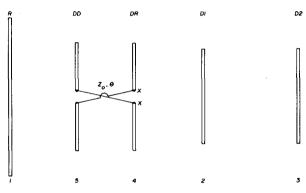


fig. 1. Layout of a Yagi with a broadband feed system of the type manufactured by KLM. The main driver (DR) and dependent driver (DD) are connected through a transposed transmission line with characteristic impedance Z_o and phase angle θ . Performance of this type of Yagi can also be analyzed by computer, as discussed in the text.

the new input file (element current file) and computes and displays a number of useful Yagi properties. These are maximum gain (dBi) and angle at which it occurs, reverse gain (dBi) at the same reverse elevation angle, front-to-back ratio, and the driving point impedance of the first driver. Since this program does not spend time displaying the complete patterns, it is fast and therefore capable of rapidly running through a large number of different situations.

other antenna types

Programs can be written that will perform equivalent calculations for different antenna systems. First of all, a somewhat different element current program is needed for a conventional Yagi if the driver *current* is specified rather than its voltage (using **eqs. 10** and **11** from reference 1). You can also write a different program to handle a broadband drive system as used in the KLM antennas, but to do this requires a little manipulation. In this type of antenna the drive system is not just a single element but actually consists of a voltage-fed main driver whose input terminals also feed a dependent driver in parallel through a crossed transmission line with characteristic impedance Z_0 and phase delay angle Θ . With conventional transmission line equations you can relate voltages and currents at both ends of this transmission line.

As an illustration, a 5-element KLM-style beam can be simulated as shown in **fig. 1**. The parasites are elements 1, 2, and 3; the main or master driver is 4, and the dependent or slave driver is 5 (connected through the transposed transmission line). Start with the five linear equations in matrix form:

$$\begin{vmatrix} Z_{11} & Z_{12} & Z_{13} & Z_{14} & Z_{15} \\ Z_{21} & Z_{22} & Z_{23} & Z_{24} & Z_{25} \\ Z_{31} & Z_{32} & Z_{33} & Z_{34} & Z_{35} \\ Z_{41} & Z_{42} & Z_{43} & Z_{44} & Z_{45} \\ Z_{51} & Z_{52} & Z_{53} & Z_{54} & Z_{55} \end{vmatrix} \times \begin{vmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \\ I_5 \end{vmatrix} = \begin{vmatrix} 0 \\ V_4 \\ V_5 \end{vmatrix}$$
(1)

 V_4 (the main driver voltage) is given, but V_5 (dependent driver voltage) must be determined from the transmission line equations which relate V_5 and I_5 to V_4 and transmission line current at terminals XX. For a transposed lossless transmission line:

$$V_5 = -V_4 \sec \Theta - j Z_0 I_5 \tan \Theta$$
 (2)

so that the final five linear equations become in matrix form:

	Z_{11}	Z_{12}	Z ₁₃	Z ₁₄	Z15	1	$ I_1 $		0	
	Z_{21}	Z_{22}	Z ₂₃	Z_{24}	Z25		I_2		0	
	Z ₃₁	Z 32	Z33	Z34	Z35	+	$ \begin{array}{c} I_1\\ I_2\\ I_3\\ I_4 \end{array} $	=	0	(3)
Ì	Z ₄₁	Z ₄₂	Z43	Z44	Z45					
					$Z_{55}+jZ_0 \tan\Theta$		15		− V4 sec θ	

Note that the input file for this antenna must contain information about the number of parasites, the number of independent or master drivers, and the number of dependent or slave drivers; the dependent or slave driver must contain a statement as to which is its master driver, and must provide Z_0 and Θ for the transposed transmission line. The *Z* matrix and *V* vector must be modified as shown, then the program can proceed as before.

validation

With the tools for computing a wide spectrum of antenna characteristics at hand, it is crucial to ask for some experimental validation. I have already commented on the inaccuracies inherent in physical modeling (although such inaccuracies are believed to be of little consequence), and have discussed approximations which have been used in various computations. These approximations are expected to be most serious for those antenna properties

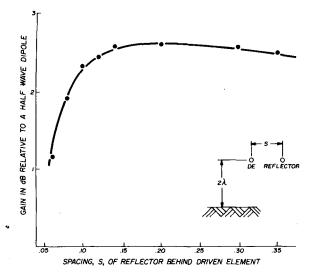


fig. 2. Gain in dB of a driven element and reflector for different spacings between elements (*from NBS Report 688*).

which depend critically on precise element currents, such as F/B ratio. On the other hand, it should be possible to calculate other properties such as gain or directivity with reasonable accuracy (comparable to the accuracy of the currents themselves). The calculations are expected to produce superior results for short antennas (few elements) and for frequency regions relatively close to design center frequency.

Experimental information suitable for validation is frustratingly difficult to obtain! While a great amount of experimental results have been published, it is extremely difficult to find examples of *accurate* measurements made under conditions where external factors have been properly considered or eliminated.

NBS Yagi experiments

For example, let's examine the experimental results reported by Peter Viezbicke in NBS Technical Note 688.3 This publication provides a rich range of gain and pattern measurements for a wide variety of Yagi configurations. In most cases, all relevant dimensions of the Yagi design are given, so this publication is a fine vehicle to test the validity of theoretical computational methodology; shortly, I will attempt to make such comparisons. At the same time. Technical Note 688 is a frustrating report; it contains several cases of inconsistent information and many times there is lack of supporting technical information. Careful documentation and editing at the time of publication would have made this report of a really excellent series of measurements much more valuable.

I must first comment on the NBS experimental approach. Viezbicke states that *all* tests on the

(receiving) test antennas were carried out using a nonconducting Plexiglas boom mounted three wavelengths (3λ) above ground (readers of the NBS paper will note that the very first figure shows 2λ). The (transmitting) generator antenna was mounted 320 meters (1050 feet) away and at a height above ground to illuminate the receiving antenna at "grazing angles." I interpret this to mean that the transmitter is also at a height over ground of 3λ . The nature of the intervening ground, which is highly relevant, is never mentioned! A reference half-wavelength dipole is mounted about 5λ to the side of the antenna under test (also at 3λ above ground). All tests were made using a frequency of "400 MHz" but the frequency precision is never mentioned, and this may be guite important, as shall be seen.

If my interpretation of the experimental setup is correct, received field strength at the receiving site will be the vectorial sum of the field from the direct ray and the field from the ray reflected from the ground at a point midway between transmitter and receiver. At these grazing angles, the reflectivity of the ground is probably near unity and for horizontally polarized radiation ground reflection will give a phase change of 180 degrees. Thus, the two rays interfere and nearly cancel each other, making the received energy nearly zero! It can be seen instantly that the actual received energy is *extremely* sensitive to the nature of ground midway between transmitter and receiver, *i.e.*, its true reflection coefficient, its height, and its degree of flatness (which influences its focus-

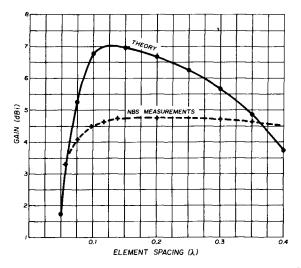


fig. 3. Comparison of computed gain with published NBS measurements (see fig. 2) for the same 2-element Yagi shows very poor agreement. As discussed in the text, however, the disagreement was apparently caused by measurement errors; when the NBS E and H plane patterns for the same antenna (s = 0.2) are summed and properly averaged, the isotropic gain is calculated to be 6.71 dBi (compared with 6.70 dBi in the THEORY curve in the graph).

ing properties). Moreover, since the reference antenna "sees" a different ground patch midway to the transmitter, its inherent sensitivity can be different from that of the antenna under test. For example, if the ground patch seen by the reference antenna is only 1 foot (30 cm) higher than the ground patch seen by the test antenna, a systematic error of about 2 dB will occur! It is clear, therefore, that the NBS experimental setup invites systematic errors.

Yagi gain

Let me turn now to the experimental results reported by Viezbicke. I will start with the measured gain of

table 1. Optimized lengths of parasitic elements for Yagi antennas of six different lengths (reflector spaced 0.2) behind the driven element. element diameter 0.0085λ).

		Length	of Yagi i	n Wavel	engths	
Length of	0.4	0.8	1.2	2.2	3.2	4.2
Reflector, λ Director Length, λ	0.482	0.482	0.482	0.482	0.482	0.475
1st	0.442	0.428	0.428	0.432	0.428	0.424
2nd		0.424	0.420	0.415	0.420	0.424
3rd		0.428	0.420	0.407	0.407	0.420
4th			0.428	0.398	0.398	0.407
5th				0.390	0.394	0.403
6th				0.390	0.390	0.398
7th				0.390	0.386	0.394
8th				0.390	0.386	0.390
9 th				0.398	0.386	0.390
10th				0.407	0.386	0.390
11th					0.386	0.390
12th					0.386	0.390
1 3 th					0.386	0.390
14th					0.386	
15th					0.386	
Spacing between						
directors, in λ	0.200	0.200	0.250	0.200	0.200	0.308
Gain relative to half-wave						
dipole, dB	7.1	9.2	10.2	12.25	13.4	14.2
Isotropic gain, dBi	9.25	11.35	12.35	14.40	15.55	16.35

a 2-element (dipole and reflector) Yagi as a function of spacing S. This is shown in the NBS paper as Fig. 1 and for convenience is reproduced here as fig. 2. Unfortunately, Viezbicke failed to state an essential dimension - the actual length of the reflector itself; I assume that it is 0.482λ (which was usually used for other Yagis). On this assumption I have calculated a theoretical result shown in fig. 3. The theoretical gain is expressed in dBi (referenced to an isotrope); a direct comparison can be made by adding 2.15 dB to the experimental NBS findings.

The comparison in fig. 3 is totally unsatisfactory;

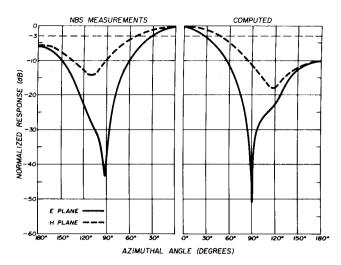


fig. 4. Comparison of the E and H plane radiation patterns for a 2-element Yagi with 0.2λ spacing as measured by the National Bureau of Standards, left, and predicted by theory, right, shows good agreement.

not only are the shapes of the plots different, the peak gains (at say S = 0.2) are very different! Viezbicke gets 4.77 dBi, whereas I calculate 6.70 dBi. In an attempt to understand this conflict, look at Fig. 12 in the NBS publication which illustrates the measured pattern of the same 2-element Yagi (S = 0.2) in both E and H planes. I have numerically summed and properly averaged the (power) response over the total 4π solid angle and have found the directivity, or isotropic gain, to be 6.71 dBi! Thus, the measured pattern exhibits gain which agrees with my theoretical result (see fig. 4). I can only conclude that some unexpected system error was present for the NBS series of measurements shown in their Fig. 1.

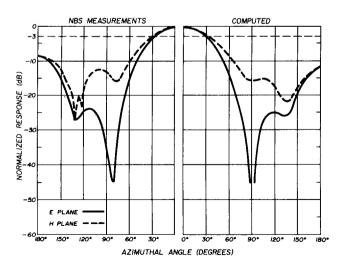


fig. 5. Comparison of the E and H plane radiation patterns for a 3-element, 0.4λ long Yagi, measured by NBS, left, and predicted by theory, right.

In the NBS paper a set of six specific Yagi designs is shown ranging in overall length from 0.4 to 4.2 wavelengths. If you refer directly to the NBS publication you will note that the first director for the 0.4 λ 3element design is shown in NBS Fig. 9 to be 0.442 λ long rather than 0.424 λ , as shown in NBS Table 1. With this correction, the specifications for these six NBS Yagi designs are listed here in **table 1**.

I have calculated the theoretical gains for these six Yagi designs and also their theoretical patterns for comparison with those shown in the Viezbicke paper (NBS Figs. 14 through 19). For convenience in making comparisons, the NBS patterns and my newly calculated theoretical patterns are shown side by side

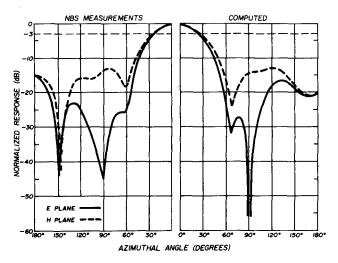


fig. 6. Comparison of the *E* and *H* plane radiation patterns for a 5-element, 0.8 λ long Yagi, measured by NBS, left, and predicted by theory, right.

in **figs. 5** through **10**. It is apparent that for all cases there is a striking similarity, not only in the qualitative details of lobe structure, but in most quantitative aspects! Careful scrutiny of the experimental results show some variances between the two halves; theoretically, of course, the two halves are totally symmetrical. Agreement of this kind is gratifying and demonstrates that the experimental patterns were made with great care and also that the theoretical calculations seem to give valid answers. This is especially comforting since the 4.2 λ Yagi design is very long and contains many parasites. This is precisely the situation where theoretical approximations (cylindrical element resonant lengths, mutual, and self-impedances) are most sensitive.

The gain of each of these Yagi designs can be obtained in several ways, and it is illuminating to compare all methods. **Table 2** shows a comparison of three experimental methods, all derived from the

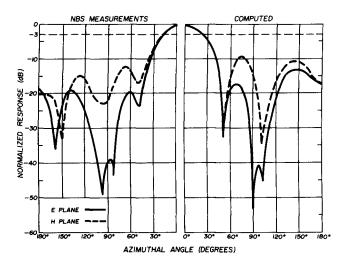


fig. 7. The *E* and *H* plane radiation patterns of a 6-element Yagi with a length of 1.2 λ — measured (left) vs theory (right).

NBS data, and my theoretical calculations. Column 1 shows the NBS measured gain referenced to a half-wavelength dipole, but corrected to give isotropic gain (dBi). Column 2 shows the gain calculated from the measured half-power main beam angles by the usual formula, $dBi = 10 \log (41253/\Theta_H \Theta_E)$, where the *H* and *E* half-power angles are measured in degrees.⁴ The third column is derived entirely from the experimental NBS patterns; in each case I have calculated the directivity by appropriately summing all 10 degree intervals. This pattern averaging, if carefully done, should yield a reasonably reliable result, free from any systematic error (due to ground reflections, for example). The last column is the result of my theoretical calculations made on each design.

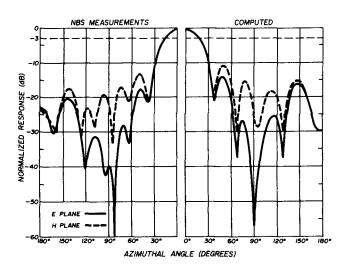


fig. 8. The *E* and *H* plane radiation patterns for a 12-element Yagi with a length of 2.2λ — measured (left) vs theory (right).

In table 2, note that column 1 (NBS measurements) and column 4 (theory) are in agreement within Viezbicke's stated estimated accuracy of about 0.5 dB with the exception of the value for the 2-element Yagi (discussed earlier). Column 2 is derived from measured half-power angles; it is quite difficult to determine such angles with any great precision and, moreover, the method is nothing more than a crude approximation at best. Column 3 (derived from NBS experimental patterns) is in remarkable agreement with theory in all cases, especially considering that the summation in 10 degree intervals is really too coarse and that values in these 10-degree intervals were only ''eyeballed'' from the published NBS patterns.

table 2. Gain of six different NBS-designed Yagi antennas, in dBi, as determined by four different methods.

NBS Yagi type	NBS measure- ments	calculated from half-power beamwidth	pattern integration	computer derived
2 element (0.2λ)	4.77	7.50	6.71	6.70
3 element (0.4λ)	9.25	10.02	9.62	9.16
5 element (0.8λ)	11.35	11.86	11.41	10.73
6 element (1.2λ)	12.35	13.90	12.64	11.80
12 element (2.2λ)	14.40	15.28	14.28	14.04
17 element (3.2λ)	15.55	16.63	15.47	15.20
15 element (4.2λ)	16.35	17.38	16.22	15.71

table 3. Measured and calculated gain in dBi of Yagi antennas with average director lengths.

		NBS	
NBS Yagi type	director length	measured . gain	computed gain
5 element (0.8λ)	0.4260λ	11.27	10.68
6 element (1.2λ)	0.4240λ	12.24	11.71
12 element (2.2λ)	0.4017λ	13.92	13.62
17 element (3.2λ)	0. 394 6λ	14.83	14.68
15 element (4.2λ)	0.4008λ	15.55	15.15

effect of director length

Fig. 7 in the NBS report shows an interesting experimental result: The selected designs are superior in gain to simplified Yagis (all directors of equal length) for booms longer than about one wavelength. Although Viezbicke did not give the lengths of the directors used for the simplified Yagis, I have made calculations using directors with lengths that are the *average* of those used in each of the five longer Yagis. His measurements of gain and my theoretical calculations are shown in **table 3**. Again, these results are in satisfactory agreement.

Fig. 11 shows a graph of my theoretical results for

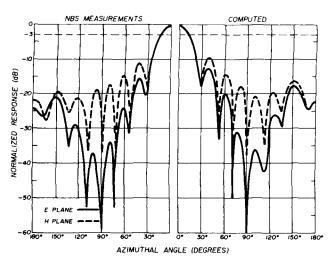


fig. 9. *E* and *H* plane radiation patterns of a 17-element, 3.2λ long Yagi — measured (left) vs theory (right).

both **tables 2** and **3**, together with the published NBS experimental points. Note that the theoretical results support the idea that while the simplified (equal director length) Yagi is just as good as a more sophisticated design for booms shorter than one wavelength, a slight gain improvement in the gain of long Yagis is apparently possible by using directors of different lengths.

Viezbicke's Fig. 9, reproduced here as fig. 12, shows the results of an interesting experiment in which gain for a given Yagi design was measured using a series of different director lengths. Gain curves were produced for each of a wide range of element diameters. These measurements allow an interesting test of the theory of the reactance of cylindrical elements (see eqs. 4, 5, and 6 in my previous article).¹ In principle, the peak of each gain curve

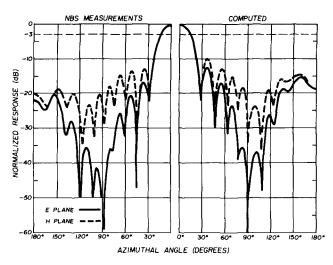


fig. 10. *E* and *H* plane radiation patterns of a 15-element, 4.2λ long Yagi — measured (left) vs theory (right).

should correspond to a single element reactance; this reactance value should, of course, be the same for all peaks. The reactance of the directors can be calculated from eqs. 4, 5, and 6 using the lengths of an element which gives the peak gain for each element diameter. From fig. 12, I have estimated, as carefully as possible, the lengths of elements which produce peak gain; table 4 shows the results.

The standard deviation actually corresponds to an element length variation of only 0.3 per cent; it is likely that such an error could easily occur in estimating the position of the peak gain or even in physically constructing the elements used in the experiment. Thus, this experiment seems to give strong support for the theory of cylinder length resonances.

table 4. Element lengths (estimated from fig. 12) which produce maximum gain, their resonant length *LER*, and the reactance X in ohms. Average reactance is -77.22 ohms; standard deviation is 2.24 ohms.

element	length			X
diameter	inches	(cm)	LER	(ohms)
0.08 cm	13.163	(33.4)	0. 4816 λ	- 81.07
0.16 cm	13.033	(33.1)	0.4791λ	- 75.63
0.32 cm	12.767	(32.4)	0. 4759 λ	- 75.88
0.64 cm	12.358	(31.4)	0.4714λ	- 78.57
0.95 cm	12.093	(30.7)	0.4680λ	- 78.28
1.27 cm	11.860	(30.1)	0. 4650 λ	- 77.93
1.59 cm	11.674	(29.7)	0.4622λ	- 76.68
2.54 cm	11.19 1	(28.4)	0.4548λ	- 73.70

gain variations

Perhaps one of the most interesting experimental results of the NBS work is the strange "oscillating" gain characteristic shown in Viezbicke's Figs. 4, 5, and 6, reproduced here for convenience in figs. 13, 14, and 15. I have attempted to calculate a number of these cases and although the oscillating gain phenomenon does show up, the detailed agreement between experiment and theory is not impressive. However, I have found that the exact behavior of these long and heavily (director) loaded arrays is critically dependent on frequency and/or exact director lengths. Accordingly, I have run calculations for several frequencies around the central frequency (normalized to unity). The results are shown in fig. 16, where the theoretical calculated points for each frequency are connected by line segments for clarity.

Superimposed on the plots of **fig. 16** are Viezbicke's experimental points, which should correspond to the theoretical calculations. The theoretical results generally show that the accurately computed points do not really lie on a smooth curve; this is the reason connecting lines are used. Theory *does* give the oscillating gain phenomenon, but it also shows

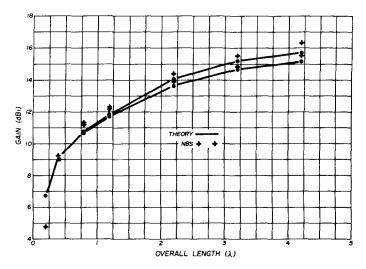


fig. 11. Graphical comparison of the data from tables 2 and 3 showing theoretical (computed) gain and measured gain figures published by NBS. Computed curves support *NBS Report 688* which shows slight gain increase is possible with different-length directors.

that the details depend sharply on the *exact* transmitter frequency. Pretty good agreement can be obtained between experiment and theory if it is assumed that slightly different frequencies (± 1 to 3 per cent) were used from experiment to experiment. It is perhaps significant that the most likely frequency does not appear to be systematically high or systematically low. It is unfortunate that Viezbicke neither specified frequency accuracy, nor published the exact frequency for each experiment.

The theoretical results are interesting from three points of view. First, for a given frequency the calculated points do not appear to lie on a smooth curve; it is likely that the gain varies somewhat with side- and back-lobe structure. The calculated F/B ratio varies greatly from point to point and it would seem reasonable that there would be some reaction on forward

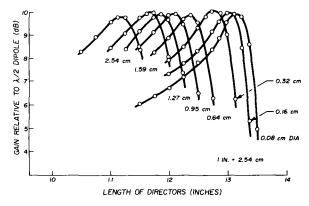


fig. 12. Measured gain vs director length at 400 MHz for a Yagi 1.25 λ long, using three directors of different length and diameter, spaced 0.35 λ .

gain. Secondly, for long Yagis the results vary significantly with the exact frequency used! Third, detailed calculations show that beyond the end of the first "oscillating" gain cycle, Yagi gain is not maximum in the plane of the Yagi; instead, the front beam appears to develop a central dimple causing maximum gain to occur at an appreciable elevation angle. (The plots in **fig. 16**, however, show only the gain at zero elevation angle because this is what was presumably done in the NBS experiments).

Note that all three of these behavioral aspects of long Yagis derive from theoretical computations; they were not even suspected by the NBS experimenters! This illustrates the point that an extensive variety of Yagi configurations can be explored *much*

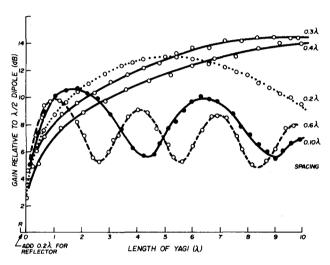


fig. 13. Yagi gain as a function of length (number of directors) for different constant spacings between directors of length equal to 0.382λ (Fig. 4, NBS Report 688).

faster theoretically, and with much more relative accuracy, than can be accomplished experimentally. As a result, new concepts and understanding are emerging.

summary

Let me summarize the overall comparison of the experimental results of the NBS group with the theoretical results. The comparisons show total agreement on Yagi *gain* for an astonishing range of models with the single exception of the direct measurements on a 2-element beam. A gain figure derived from the measurement of the *pattern* of the 2-element beam, however, *does* agree with theory; it is my belief that some unanticipated error was made by the NBS group in their measurements.

The comparisons also show excellent agreement for Yagi *patterns*, again over a wide range of models. This should be a very sensitive test of the accuracy with which mutual and self-impedances are repre-

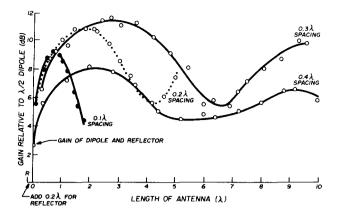


fig. 14. Yagi gain as a function of length (number of directors) for different constant spacings between directors of length equal to 0.411λ (Fig. 5, *NBS Report 688*).

sented, as well as the resonant frequency calculations of cylindrical elements. This latter point is also strongly supported by the consistent values of parasitic element reactance for elements whose diameters vary over a range of greater than 30 to one.

Finally, the strange "oscillating" gain phenomenon observed by Viezbicke can be reproduced theoretically. Agreement is only qualitative if the NBS group used an *accurate* 400.0 MHz frequency in their tests; however, the comparisons suggest that they may have actually used a nominal "400 MHz," with frequencies within a range of ± 3 per cent. In this event, the comparisons indicate a potentially remarkable agreement in quantitative aspects as well.

Since these comparisons include very long Yagi models (up to 6λ long) and correspondingly large numbers of parasitic elements (up to 40), it seems certain that the computational methodology I have outlined should be generally trustworthy. This should be especially true for shorter antennas (with fewer elements) with which I will be primarily concerned throughout this series of articles. Especially noteworthy is the fact that all computed results contain

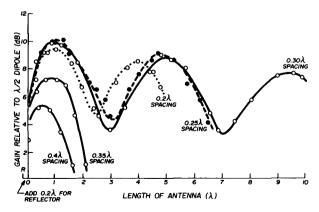
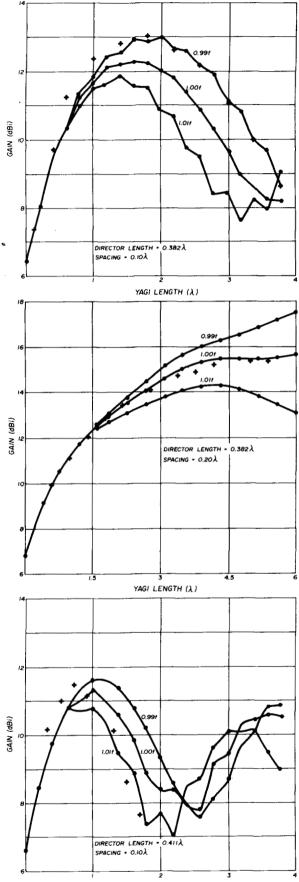


fig. 15. Yagi gain as a function of length (number of directors) for different constant spacings between directors of length equal to 0.424λ (Fig. 6, *NBS Report 688*).



YAGI LENGTH (λ)

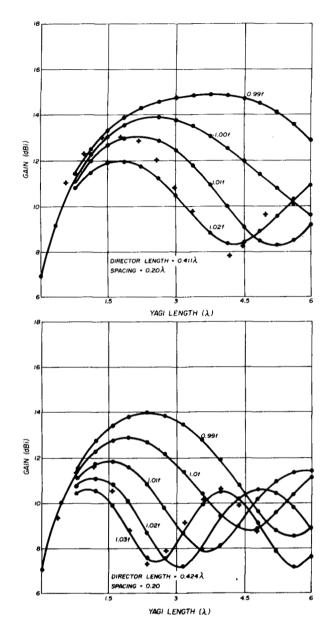


fig. 16. Graphs of Yagi gain vs overall antenna length for various director spacing and director lengths at small frequency differences; frequency designations on each plot have been normalized to the center design frequency. Points which are not interconnected are experimental data published in *NBS Report 688*.

no adjustable constants or parameters; they are all derived from basic physical principles using adequately accurate mathematical approximations.

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ham radio

high-performance broadband balun

Theory and design of broadband 1:1 baluns, with construction details for an improved model of the well-known W1JR balun

Sometimes equipment that looks the simplest to build is really the trickiest to make work, and vice versa. Take, for instance, a balun. What sounds simpler than wrapping a few turns of transmission line around a toroid and getting a balun? Or do you? My experience is that you may, but then again you may not.

Over the last few years many articles have appeared on the 1:1 balun, but absolutely no information — to the best of my knowledge — on why they work, how to design one, or, equally important, how to test one.

Almost 100 years ago, the eminent British physicist, Lord Kelvin, said, in effect, "If you can express what you are speaking of in numbers, you understand it. If you cannot express it in numbers, your knowledge is very meager and unsatisfactory." From this standpoint, our knowledge of 1:1 baluns is "very meager and unsatisfactory" indeed!

In this article I give a brief description of the historical development of the balun, explain how a 1:1 balun works, include a procedure for designing a balun, show some pitfalls that are all too easy to fall into, and finally present some test results on a balun I have built.

brief history

The concept of a broadband balun made by winding a transmission line into an inductance originated in 1944 by G. Guanella¹ of the Brown-Boveri Company in Switzerland. In 1959, Ruthroff² of Bell Laboratories wound the transmission line onto a ferrite toroid and obtained bandwidth ratios as high as 20,000:1. His paper appears to be the takeoff point for the present balun designs. In the Amateur literature, Dick Turin,³ Jerry Sevick,⁴ Bill Orr,^{5,6} and Joe Reisert⁷ have all presented articles on broadband baluns.

theory of operation

Consider a balanced transmission line of characteristic impedance $Z_{ch} = R_L$, wound into an inductance, and feeding a balanced load RL, as shown in fig. 1 (A). The normal transmission line currents, i_1 and i2, are out of phase; their magnetic fields cancel, and the inductive reactance is essentially zero or, at least, very low compared with the transmission-line characteristic impedance. As far as these currents are concerned, the balun appears only as an added length of transmission line, of characteristic impedance Z_{ch} , tied to the end of the usual unbalanced transmission line whose characteristic impedance is also Z_{ch} . If the load is balanced with respect to ground, the voltage at the junction point G will be zero with respect to ground; the current i_g will also be zero.

That this is true can be seen from the following argument: Assume that the voltage at G is not zero; the unbalance can be represented by a voltage generator, which causes a current i_g to flow, as shown in fig. 1 (B). This current divides into two in-phase currents, $i_g = i_3 + i_4$, which flow through the two windings of the balun. Since these currents are in phase, their magnetic fields will add instead of cancelling, and the balun winding will appear as an inductance. If the inductive reactance of the balun winding is sufficiently large compared with $R_L/2$, the in-phase currents that cause the unbalance will be essentially zero compared with the normal transmission line currents, i_1 and i_2 .

Thus, if the balun is working properly, it will provide a high impedance to in-phase components and a low impedance to out-of-phase components, effectively isolating the balanced and unbalanced sides of the balun.

In theory, then, the problem of designing a balun boils down to simply winding an inductance to provide a high impedance over the required frequency

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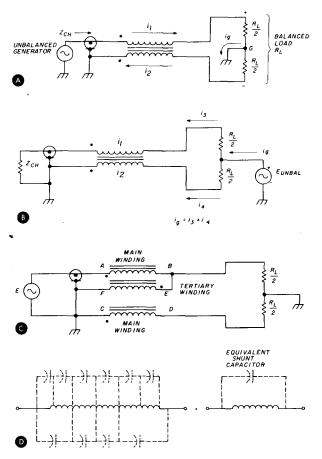


fig. 1 (A): Basic 1:1 balun schematic. Circuit transforms an unbalanced transmission line of impedance Z_{ch} to a balanced load R_L , where $R_L = Z_{ch}$. (B): Schematic of a 1:1 balun showing an unbalanced condition represented by an unbalanced generator, E_{UNBAL} . (C): A 1:1 balun with a tertiary winding, E-F. (D): The left-hand drawing shows an inductor with a few of the possible infinitesimal stray capacitors. These can be equated to one shunt capacitor, as shown in the right-hand drawing.

range. Unfortunately, life is not always simple; neither are baluns, for although their theory is simple, their actual design is guite involved. First, designing an inductance to provide a high impedance over a wide frequency range can be a substantial problem. Second, for the winding to be a high impedance over any frequency range, the current required to magnetize the core must flow without disrupting the desired signal current and the voltage relationship existing in the balun or load. In a 1:1 balun, this is accomplished by adding a tertiary winding, as shown in fig. 1 (C). This tertiary winding, in turn, must be designed and constructed so that it does not create any additional problems of its own. But, as we'll see later, although the tertiary winding will considerably complicate matters, it is absolutely essential if acceptable balance and good low-frequency response are to be obtained. I'll now discuss these problems in detail as they are related to balun design.

inductance

In designing an inductor to be used over a wide bandwidth, the basic problem is capacitance - stray capacitance. Schematically, an inductor is drawn as a coil of wire; note the solid line in fig. 1 (D). Physically, however, stray capacitance is present as suggested by the dotted lines. Actually, the capacitance is distributed; i.e., it exists between every infinitesimally small length of coil and every other infinitesimally small length of coil; it is not discrete, as implied in fig. 1 (D). The effectiveness of these incremental capacitors depends on how much energy each one stores. The energy stored by a capacitor is equal to energy = $\frac{1}{2}$ CE², where C is the capacitance in farads and E is the voltage across the capacitor terminals; therefore, it is important to minimize the capacitance across those parts of an inductance where the voltage is the largest; *i.e.*, at the ends of

Since it's very difficult to consider all the incremental capacitors in analyzing the behavior of an inductance, it has become standard practice to hypothesize a single capacitor connected across the entire coil storing the same amount of energy as all of the incremental capacitors and to call this the stray capacitance of the coil.

the coil. Unhappily, this is not always possible!

As the frequency across the inductance is increased, a frequency will be reached where the inductance and stray capacitance will become parallel resonant; at this frequency, the inductor will appear as a very high resistance. As the frequency is further increased, the reactance decreases to zero. At the series-resonant frequency, the coil impedance becomes very low. For a balun, the series-resonant region must be avoided, as all isolation between the balanced and unbalanced sides of the balun is lost. My experience has been that it's necessary to limit operation to about one octave below (one-half) the series-resonant frequency, or else the balun phase and amplitude balance balun will be upset.

Probably the most readable and useful discussion of wideband inductor design in either the professional or Amateur literature was written by Vernon Chambers⁸ over 25 years ago. Chambers' article refers to the design of rf chokes, but the basic problems are similar to baluns. More recently, Doug DeMaw⁹ has written a very helpful article on toroid inductor design.

As implied previously, one of the most effective ways of reducing the distributed capacitance is to separate the ends of the winding as much as possible. This brings us to the concept of the super toroid.

super toroid

The super toroid was developed in the 1970s by

T.A.O. Gross,¹⁰ primarily as a means of reducing stray pickup from external fields, but was introduced to the Amateur community only recently by Reisert.⁷ With the super toroid, one-half the circumference of the toroid is wound in the usual manner. The winding is then taken across the diameter of the core and the last half wound in the opposite direction, as shown in **fig. 2**. The advantage of this type of winding for balun application is that the ends of the winding, where the voltage is the highest, are at opposite sides of the core where the capacitance is minimum.

Every inductor I've wound using the super toroid technique has had bandwidth characteristics superior to the same core wound with an equal number of turns in the usual manner! Why not use the super toroid concept, then, in the balun? The problem is the tertiary winding.

tertiary winding

As we know, a balun functions by providing a high impedance, usually inductive, between the balanced

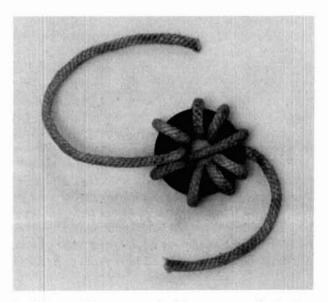


fig. 2. A toroidal core wound with a super toroid winding. This core is wound with clothes line to make it show up better in the photograph.

and unbalanced sides. To obtain this inductance, it's necessary that a path be provided for magnetizing current. Ruthroff² states that with the balanced load disconnected, a dc path must exist between the unbalanced input side and ground. The way of providing this is to use a tertiary winding as shown in fig. 1 (C).

This circuit is easier to visualize if redrawn in the form of an autotransformer as shown in **fig. 3**. Voltage levels with respect to ground are also shown, and it is seen that the effect is that of a 1:1 autotransformer with the voltages on the balanced side low-

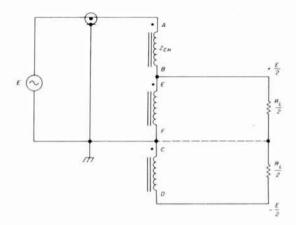


fig. 3. A balun with a tertiary winding drawn in the form of an autotransformer. The voltage across each winding is $\frac{E}{2}$ volts.

ered by $\frac{E}{2}$ volts, compared with the unbalanced input voltage. The total voltage across either the input of the balun or the balanced output, however, is still the same, *E* volts.

The path for the magnetizing current is also evident: from the unbalanced generator through windings **A-B** and then **E-F** to ground. The important point to note is that the magnetizing current does not pass through either half of the load impedance. If the balun is to couple a balanced generator to an unbalanced load, and the unbalanced side is open circuited, the magnetizing current will pass through the tertiary winding, **E-F**, and the lower half of the transmission line, **C-D**, back to the balanced generator without passing through the load impedance.

practical considerations

To show the practical need for tertiary winding, I'll discuss the effects that the lack of a tertiary winding will have on balun performance. Consider the 1:1 balun with no tertiary winding, as shown in fig. 4. Because of the instrumentation considerations, it's easier to use a balanced load, consisting of two equal-value resistors, each of value $\frac{R_L}{2}$, and to make measurements at the unbalanced side rather than vice versa, which would require a balanced impedance bridge. The center of these two resistors may be grounded, as in fig. 4 (B), or ungrounded, as in fig. 4 (A). If the balun is working properly, there will be no difference whether the resistors' center is grounded or not. Any differences can give an important clue as too the type of problem being encountered. In all cases, I'll assume the total load impedance, R_L , is equal to the characteristic impedance, Z_{ch} , of the transmission line.

First, let's assume the center tap of the load is not grounded, as in fig. 4 (A). The resistive component

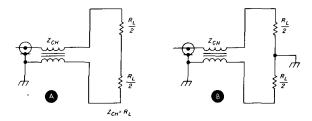


fig. 4. Test circuit for a 1:1 balun *without* a tertiary winding, with and without the load impedance centertap grounded. $Z_{ch} = R_L$ in both cases.

of impedance looking into the unbalanced end will be R_L ohms regardless of the frequency. In fact, it will be R_L ohms if measured with a dc ohmmeter.

Thus, you can be lulled into a false sense of security if you depend on impedance measurements alone. This happened to me. I ran down to the low frequency limit of the signal generator, and the impedance remained on 50 ohms! But let's look at what happens to the balance in the output voltage in both magnitude and phase. This latter point does not appear to have been considered by most writers on baluns.

phase shift

Ideally the voltages between point **B** and ground and **D** and ground (**fig. 3**) should be equal in magnitude and opposite in phase. Each of these voltages should also be equal to one-half the voltage between **A** and ground (the input voltage). As the frequency is lowered below that at which balun action is effective, the voltage at **B** will approach that at **A**, and the voltage at **D** will approach zero. The total voltage across R_L is still equal to the input voltage but not balanced with respect to ground.

The need for magnetizing current can be verified by making voltage and phase measurements across the two halves of the balanced load. The phase and

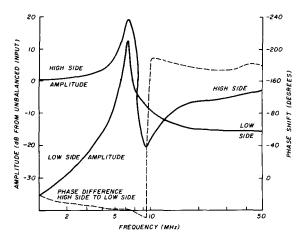


fig. 5. No tertiary winding. Phase and amplitude of balanced output voltages — load centertap not grounded, as in *fig.* 4(A).

magnitude of the output voltage as a function of frequency are shown in **fig. 5** for the case where the load resistance centertap isn't grounded. The amplitude measurements are shown in dB with respect to the input voltage; the voltage across each half of the load resistor should be one-half, or 6 dB down, from the input voltage. The phase measurements are the phase difference between the balanced output voltages; these, of course, should be 180 degrees apart. These measurements are plotted in **fig. 5**.

After looking at **fig. 5**, you're probably thinking, "No well-behaved balun would act like that. There must be something wrong with his test setup." My answer is that the balun is *not* well behaved and there's *nothing* wrong with the test setup.

Let's look at the same balun, measured at the same time, with the same equipment, but with the

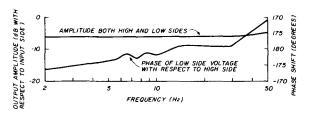


fig. 6. Same as in *fig. 5* but with the load centertap grounded to provide a path for the magnetizing current. This balun is usable between 3 and 36 MHz.

load-resistor centertap grounded. The magnetizing current now has somewhere to flow. The amplitude and phase responses shown in **fig. 6** are very close to what one would expect. The balun appears to be usable over 3-35 MHz, just by grounding the load impedance centertap! This, I believe, demonstrates the necessity of providing a path for the magnetizing current.

Before leaving the subject, look at the input impedance of the balun with and without the load centertap grounded, as shown in **fig. 7**. There is very little difference between the two impedance curves, so we can conclude that impedance measurements by themselves are not necessarily a good measure of a balun's actual performance.

So, a path for the magnetizing current is absolutely necessary, but grounding the centertap may not be practical in all cases and it causes an unbalance even when it is possible. This result leads us to look at another way of providing a path for the magnetizing current, namely, a tertiary winding.

coupling between windings

As previously noted, the tertiary winding is connected between the high side of the balanced output and ground as shown by **E-F** in **figs. 1 (C)** or **3**. You'll

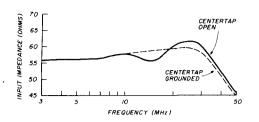


fig. 7. Input impedance of 50-ohm balun without tertiary winding. Load impedance = 55.7 ohms balanced.

observe that it must support a voltage of $\frac{E}{2}$ volts, the

same as each of the main windings. In fact, if accurate amplitude and phase balance are to be maintained at the balanced terminals, the voltage across all three windings must be the same at all frequencies. This implies that the coupling between all three windings must be *very* tight; indeed, the tightness of coupling is one factor that determines the balun bandwidth, a need that doesn't seem to have been pointed out before.

The degree of coupling between the main windings is not a problem as these windings are part of a transmission line. The coupling between these windings and the tertiary winding, however, is crucial. I'll discuss how to accomplish tight coupling when I describe construction details.

The requirements imposed on the tertiary winding are basically the same as those on the main winding: first, the tertiary must support a voltage of $\frac{E}{2}$; second, the series-resonant frequency should be at least one octave above the highest operating frequency; and third, the tertiary winding must have the same number of turns as the main winding and be very tightly coupled to it.

impedance levels

There seems to be a difference of opinion about the inductive reactance required on a balun winding. The usual number given suggests that the reactance at the lowest operating frequency should be at least ten times the characteristic impedance of the balun. This might be a good number to use if the entire input voltage, *E*, were across the balun winding. Actually, only one-half the input voltage is across any one winding so that the reactance-to-characteristicimpedance ratio can be reduced to 5. In his article, DeMaw⁹ suggests a value of 4; this ratio will give a maximum VSWR of 1.64:1, which is better than most Amateurs achieve anyway.

design procedure

With the preceding background, a design proce-

dure can now be worked out as follows:

1. Number of turns and core size. The first problem is to select a core size and determine the number of turns. These two choices are interrelated. If good high-frequency performance — above 30 MHz — is required, small physical size is important. Where good low-frequency performance is important and/or high power will be used, a large size is indicated. If all three characteristics are desired, obviously compromises must be made.

Based on the experience of Reisert,⁷ I started with an Indiana General F-568-1 core of Q-1 material. Next, I wound a test coil on the core; it's not necessary to use actual transmission line for this test winding provided the winding has the same form as that of the final winding. I used no. 16 (1.3-mm) enamel wire with the turns spread out to occupy the same space as the final transmission line (RG-141/U coax in this case).

Next, I measured the coil impedance over the frequency range of interest, 3.5-30 MHz. The reactance should be at least four to five times the characteristic impedance over the entire frequency range desired.

It's a good idea to extend the impedance measurements to at least one octave above and below the wanted frequencies to ensure that the impedance is well behaved. You should find that the reactance peaks up at a given frequency, then falls off, and eventually goes negative. The peak is the parallel-resonant frequency of the winding. The frequency at which the reactance is zero is the series-resonant frequency; this frequency must be at least twice the highest frequency at which the balun will be used.

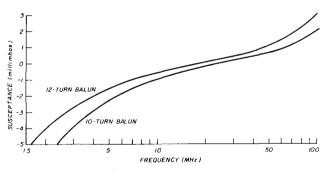


fig. 8. Susceptance of a ten- and twelve-turn coil on a F-568-1 core of Q1 material.

Generally speaking, it will be found that, for a given core and winding form, the ratio of highest-tolowest usable frequency will be approximately constant. Changing the number of turns will just slide this ratio up or down on the frequency scale.

Measurements made on a 10- and 12-turn balun are shown in fig. 8. I've plotted susceptance

 $(B = \frac{1}{X})$ since a high reactance or low susceptance is desired; the curve would be off the paper over most of the frequency range for any reasonable reactance scale. If it's required that the reactance be 250 ohms (five times the characteristic impedance), the required susceptance range is between -4 milli-

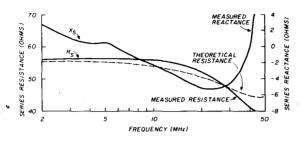


fig. 9. The measured resistive and reactive components of input impedance of a 12-turn balun with a tertiary winding. The load impedance was 55.7 ohms. The theoretical resistive component is shown by the dotted line.

siemens and 4 millisiemens.* This places the low-frequency limit at about 2 MHz for the twelve-turn winding and 3 MHz for the ten-turn.

The real (resistive) component of impedance, if measured on a series equivalent basis, should be no more than one-tenth, and preferably one-twentieth, the load impedance. If this component is measured on a parallel basis, it should be at least ten to twenty times the load impedance.

If you can't get enough turns on the core, you'll have to use a larger core or a higher permeability material. I found that twelve turns would cover 2 MHz through 30 MHz. Comparing my frequency range with Reisert's,⁷ I found that Reisert required the reactance of the winding to be at least ten times the characteristic impedance of the line, while I allowed the reactance to drop to five times the load impedance.

We can now turn our attention to the transmission line.

2. Transmission line. It doesn't seem to be generally appreciated, but the characteristic impedance of a balun transmission line *must* match the load impedance. This is not because of VSWR considerations but to obtain a flat frequency response at the high-frequency end of the balun. This implies two requirements which are seldom true in the Amateur case; First, the antenna impedance must be accurately known, whereas it is usually estimated. Second, even if you do know your antenna impedance, how do you design a transmission line suitable for balun

*Millisiemens are the SI units of conductance and susceptance replacing the millimho; the abbreviation is mS.

use with that impedance? Since the design of balun transmission lines would take an article itself, I'll not consider it further.

To see the effect of a mismatch, let's jump ahead and look at the measured input impedance of the finished balun, which is shown in **fig. 9**. Here, the 50ohm coax was terminated in a 55.7-ohm balanced load. By Amateur standards, this is a very close match, yet the input impedance sags by more than 10 per cent at 30 MHz. Why?

To find out, let's go back to the basics. The length of coax used to wind the balun measured a quarterwavelength long at 45.33 MHz. At this frequency, then, the input impedance for a 55.7-ohm termination should be 44.88 ohms caused by the quarterwave impedance inversion effect. For lower frequencies, the theoretical input impedance, as calculated from the transmission-line equation, is shown by the dotted line; in **fig. 9** note that the measured and theoretical are within 1 or 2 ohms to beyond 30 MHz. I could therefore have dramatically flattened the input impedance curve by securing a better impedance match at the output end.

If a balun designer were to look at the drooping input impedance curve, without realizing its true cause, he would think he had a bad design. On the contrary, there's nothing wrong with his balun that a better match wouldn't cure.

A mismatch has another detrimental effect namely, it adds reactance. Assuming the mismatched load is a pure resistance and the mismatch is small, which I have, the maximum reactance will be generated when the balun is about one-eighth wavelength long, or at about 22.6 MHz in this case. Notice the dip in reactance around that frequency.

This also points out the fact that, where a mismatch may occur, the balun winding should be as electrically short as possible. If the balun described were to be used on a triband beam for, say, 10, 15, and 20 meters, at least one turn, and possibly two, should be removed from the balun.

The rise in reactance at the low-frequency end is caused by the shunting effect of the tertiary winding. If it's assumed that the series equivalent reactance must be less than one-tenth the load impedance, the limiting frequency for this effect is less than 1.5 MHz.

Since this balun will be used in a 50-ohm system, a 50-ohm line will be used: RG-141/U, which is coaxial cable. I'm not particularly happy about using coaxial cable for both electrical and mechanical reasons, but I haven't been able to construct a 50-ohm balanced line that I consider satisfactory; so coax it is.

construction

Before winding the toroid, let's consider the terti-

ary winding. I said earlier that the tertiary winding should have the same number of turns as the main winding and must be coupled very tightly to it. This latter requirement means that the tertiary must be wound as close as physically possible to the main winding.

To ensure this, I removed the fabric covering from the RG-141/U but left the outer Teflon layer of insulation. I placed a length of no. 18 (1-mm) enamel wire against the Teflon and covered it with heat-shrink tubing. When the tubing had shrunk to size, I had a compact assembly that could be wound as a single unit on the Indiana General F-568-1 toroid core.*

The proof of a balun is in how well the output voltages are balanced with respect to ground in both phase and amplitude. Fig. 10 shows these measurements. The amplitude deviation is from equal amplitude (zero dB), and the phase is the actual phase of the low side of the balun with respect to the high side. This value, of course, should be 180 degrees.

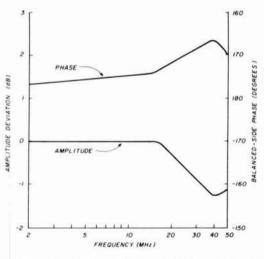


fig. 10. Amplitude and phase measurements at the balanced terminals.

These measurements were made with a Hewlett-Packard model 8405A Vector Voltmeter. Notice that both deviations peak at about 40 MHz. This could be a manifestation of the impedance mismatch and the quarter-wavelength frequency. If so, it points out again the importance of an accurate impedance match between balun and load impedances. Even considering the mismatch effect, the balance isn't too bad. It's within 1 dB and 10 degrees to 30 MHz.

conclusion

I've attempted to explain how a balun works and

*A complete kit of parts for the high-performance balun is available from Radiokit, Box 429, Hollis, New Hampshire 03049.

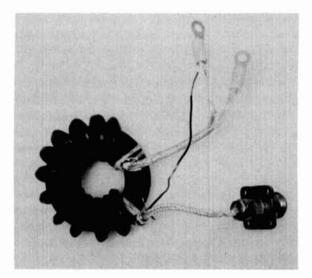


fig. 11. The finished balun.

to clarify some of the problems you must be careful about when designing and building a balun. I've emphasized the importance of 1) tight coupling between the main and tertiary windings of a balun and 2) good impedance matching between the balun and load. I feel strongly about this to the extent that I believe balun manufacturers should include impedance information in their specifications.

I've also given one method of building a balun to minimize the adverse effects of these problems. I don't pretend that the balun design I've presented is the only way around the physical limitations of a balun; it's simply my best so far.

I hope my discussion of balun problems inspires others to experiment and to produce a better balun. With ingenuity, testing and a bit of luck, the balun of the future can indeed be better than ever.

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third-generation Touch-Tone decoder

Another approach to the *Touch-Tone* decoder for your repeater or remote-base station

If you own a repeater or a remote-base station, you've probably been faced with the problem of coming up with a good *Touch-Tone** decoder. Now you can throw away your toroids and NE567s. Also, you can throw away the falsing problems that go with these devices and still not throw away half a month's salary in the process.

I'm talking about the Mostek MK5102N-5 decoder chip.¹ This device accepts the audible *Touch-Tones* and converts them to either row-column information or BCD four-bit binary information. The chip also provides a valid strobe for logic interfaces. The only external parts required are a 3.579545-MHz TV colorburst crystal and a band separation filter-limiter circuit.

To prevent falsing, the decoder requires 40 milliseconds of valid tone before the strobe and outputs generate output information. When used in the circuit described here, an input level variation above 20 dB presents no falsing problems. Differences in level

**Touch-Tone* is the registered trademark of the American Telephone and Telegraph Company.

between the high and low tones (twist) of over 6 dB don't affect performance.

features

The circuit provides the following features for use in a repeater or remote-base application:

1) Sixteen outputs for use in control circuits

2) Requires five tones of proper frequency and order within five seconds of the first digit before a control output is generated

3) Provides *, #, 1, and 0 for use in autopatch applications (optional)

4) All ten digits are available for use in a regeneration autopatch system (optional)

5) Strobe and BCD information are available for use in external circuits

6) Control outputs (command outputs) and phonepatch outputs can be either high-true or low-true

7) Requires a single + 12 Vdc supply

8) Cost is approximately \$100 for all features

Touch-Tone decoding has been covered in numerous articles. Therefore, I won't rehash all the basic information. Reference 1 gives a complete description on how the MK-5102-N5 works. For our purposes let's just say that we put high group and low group in and get BCD information out.

By James Wyma, WA7DPX, 12952 Osborne St., Arleta, California 91331

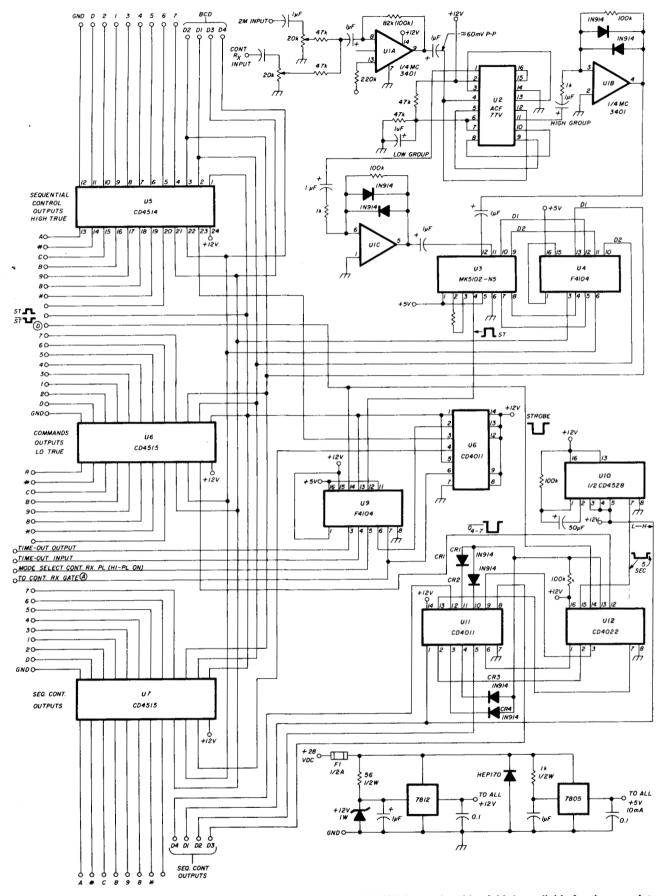


fig. 1. Improved *Touch-Tone* decoder schematic using the Mostek MK5102N-5 decoder chip. A kit is available for the complete decoder or optional circuits (see text).

circuit description

The *Touch-Tone* decoder schematic is shown in **fig. 1.** Let's start at the audio input. IC-1A is one-fourth of a MC3401 op-amp used as a buffer amplifier. The MC3401 is a special op-amp constructed to run on single-ended power supplies. You'll note that provisions are made for two inputs on the decoder (RX audio and control RX audio). These inputs should be low level (approximately 20-30 mV p-p). If your receiver audio level is much higher than this, you may have to change the input-pot value or the value of the feedback resistor in the op-amp (between pins 8 and 9).

The gain and input pot should be adjusted for a level of approximately 100 mV p-p (22 mV rms) when a tone corresponding to the digit **5** is being received (measured at IC-2 pin 4). Take care in setting up this level. If excessive levels, such as 1 volt p-p, are shown at this point, the decoder won't function properly. If both inputs are used, provisions should be made for gating the two signals so that only one signal is present at a time. This can be done by using a COR or PL logic level from the control receiver to gate a CD4066 analog switch.

When a signal (or PL) is received on the control receiver, the repeater input is gated off and the control input is turned on. When in the repeater-input mode (no control signal), the opposite occurs. If only one input is needed, delete the additional pot, resistor, and coupling capacitor.

separation filter-limiter

I mentioned earlier that the MK5102 needs a bandseparation filter and limiter circuit. The ACF7711 (IC-2) performs band separation. This could be done by cascaded bandpass amplifiers using op-amps as in reference 1. However, the ACF7711 has the advantages of size and temperature stability.

To obtain the stability of the ACF7711 in a discrete op-amp circuit, high-precision capacitors and resistors are required, which aren't cheap. The two 47k resistors and the 1- μ F capacitor on IC-2 pins 6 and 7 allow the filter to function on a single-ended supply. Don't omit the 1- μ F capacitor at this point.

Speaking of capacitors, let's make a point at this time. All the 1- μ F capacitors should be of the low-leakage type. The two outputs (high and low group) of the separation filter feed into limiters (IC-1B and IC-1C). These amplifiers feed square-wave signals to the decoder. (For a more detailed description of the band separation filter and limiter, refer to the data sheets for the ACF-7711 and the Mostek MK-5102-N5, which are available from references 2 and 3 respectively.)

Incidentally, a pin-for-pin equivalent of the ACF-7711 is made by Data Signal Corporation of Watertown, Massachusetts. However, I was informed by their sales manager, Mr. Clarence L. Walker, Jr. (after two letters and three phone calls to reach him) that: "We have plenty of business without dealing with a bunch of cheap hams who always want something for nothing." So be it. I hope you bear this in mind if you plan to do business with this company.

decoder-translator

The limited signals are fed into the high- and lowgroup inputs of Mostek decoder IC-3. The output format pin (pin 6) is grounded so that the outputs are in the form of BCD-coded information. Note that the MK-5102-N5 is a 5-Vdc CMOS device. However, the control logic used in our repeater is high level CMOS (12 Vdc). This was done to obtain a higher noise immunity level on the circuits. This function is performed by IC-4 and IC-9.

IC-4 translates BCD information. IC-9 translates the strobe to high level CMOS. The F4104 IC provides both noninverted and inverted outputs. In the strobe case, both outputs are used. The BCD information from IC-4 is fed to IC-5, IC-6, and IC-7. The BCD information is also fed into the board edge connector for use in other auxiliary equipment (such as LED number display, auto-dialer, *Touch-Tone* regenerator, or additional control circuits).

The CD4514 and CD4515 (IC-5, 6 and 7) are BCD-to-16 output decoders. The CD4514 and CD4515 are identical except for the output states. The CD4514 has true outputs high, while the CD4515 outputs are low. Other than this, the two-are pin-forpin identical. What this means is that IC-6 could be replaced with a CD4514 if you needed command outputs that were high instead of low. The same is true for IC-7. The device used in IC-5 must be a CD4514 (unless you don't want things to work).

mode-select logic

After running through the translator, the strobe is fed to IC-8 pins 1 and 5. Pins 1 and 5 are inputs to two sections of the NAND gates. The other inputs (pins 2 and 6) of these two NAND gates go to IC-9 pins 6 and 7. These two outputs of IC-9 are the non-inverted and inverted outputs of the input on pin 5. IC-9 pin 5 is what I call the "mode-select input" for the decoders.

The state of the mode select (high or low) determines whether the sequential control outputs or the phone patch outputs are enabled. To explain this, let's go through the logic when the mode select input is high. This condition corresponds to the control mode when the sequential control decoder is enabled. A high on IC-9 pin 5 produces a high on pin 6. This action causes a high on IC-8 pin 2. When a valid *Touch-Tone* digit is decoded, the strobe is high. This action places a high on IC-8 pin 1.

When both inputs (pins 1 and 2) of IC-8 are high, the NAND output is low (pin 3). This output is connected to IC-5 pin 23 (the enable pin for the CD4514). When this pin is low, the inputs are enabled. When high, the outputs all go to a low state. It's necessary to return all the outputs to a low state for the sequential decoder to function correctly. During this time, the phone-patch outputs are disabled because IC-7 pin 23 is held high. A high on IC-9 pin 5 places a low (inverted output) on IC-8 pin 6. Consequently, even though pin 5 is strobed high, the output remains high. This high state keeps IC-7 disabled.

When the mode-select input is low a similar process occurs, except that IC-7 is enabled and IC-5 is disabled. IC-6 outputs are disabled by the sequential decoder. (This will be explained later.) The mode select input can be controlled by a COR or PL output on the control receiver. If your application doesn't require the phone patch output, the mode select pin should be wired to +12 volts. The phone-patch decoder (IC-7) can be deleted from the board. If you want the phone patch outputs but not the command outputs, ground the mode-select input. If this is done ICs 5, 6, 10, 11, and 12 can be omitted.

sequential decoder

The heart of the sequential decoder is a CD4022 (IC-12) Johnson Counter, the electronic equivalent of a stepping relay. For the counter to advance, the two clock inputs must be in the correct states and the master reset must be low. Each time the inputs are in the correct conditions (a code match) the counter advances one position. When the counter advances for the fourth time, a carry output is generated. This output is called Q4-7. Q4-7 will be low during the fourth through seventh advances of the counter.

To explain, let's run through a typical code sequence. The decoder is set up so that the first four digits are a common address for command functions. The fifth digit determines the actual command output from IC-6. From the time that the first digit is sent, the next four digits must be sent within approximately five seconds. The reason for this is that when someone starts sending random digits on a pad, he will have a hard time hitting the correct digits within a five-second period.

Say that our address is 4-1-3-7. The outputs from the IC-5 are strapped to the D1 through D4 inputs (see **fig. 1**). One way to do this is to use a wire wrap edge connector on the PC board. The correct code is wire wrapped from outputs to inputs. In our case, the output pins for 4, 1, 3, and 7 would be strapped to D1, D2, D3, and D4 respectively. The four digit inputs go to one of the inputs of each section of the quad NAND gate, IC-11. The remaining input of the NAND gates is connected to IC-12, pins 2, 1, 3, and 7, the Johnson counter. These four pins correspond to 0_0 , 0_1 , 0_2 , and 0_3 of the counters. These are the first four positions in the counter outputs.

The digit inputs, IC-11, diode CR1 through CR4, and the counter outputs form a sequential coincidence circuit. Each time a correct digit is in the proper position of the sequence, the counter will advance one position on the outputs.

In our example the first digit is 4. A "4" places a high on IC-11 pin 1. The counter is in the reset position, so output 0_0 (pin 2) of IC-12 is high. These two highs cause a low on the NAND-gate output (IC-11 pin 3). This signal is coupled through CR1 to the counter clock input. The other counter clock input is connected to the inverted strobe output of translator chip IC-9.

These two simultaneous inputs cause the counter to advance if the master reset pin (IC-12 pin 15) is low. The master reset is connected to the IC-10 \overline{Q} , a CD4528 monostable multivibrator. IC-10 is triggered by the high input from D1 (IC-10 pin 4). The resistor and capacitor values on IC-10 pins 1 and 2 multivibrator determine the time constant (the length of time that the MR pin [pin 15] on the Johnson Counter stays low).

The values shown in **fig. 1** give you about five seconds to enter the remaining three address digits (D2-4) and the command digit. When the coincidence circuit has received digits 4-1-3-7-9, a low is placed on IC-12 pin 12 (Q4-7 output). This low turns on the enable input of IC-6 (pin 23). With the chip enabled, the BCD information on IC-6 input is converted to one of the sixteen corresponding output digits. These outputs are the command outputs, which drive the logic functions in your remote base or repeater.

The CD4515 produces low true outputs. As previously stated, a CD4515 can be used as IC-6 if you need high outputs to drive your logic. The command outputs will remain latched until the timer resets the master reset on the counter. These outputs will then return to all low for the CD4514 or all high for the CD4515. If you need latched command functions, the outputs could be used to drive a CD4043 or CD4044 quad R-S latch. If you need to drive relays, a CD4049 or CD4050 could be used as a buffer for the relays. If you need to drive outside information (amp or other equipment), the BCD-coded information and strobe are brought out to the edge connector pins.



power supply

If a 12-volt battery is used, the resistor and zener diode should be used. (Omit the 7812 IC. It must have at least 14 Vdc to work). If a higher voltage is used, the 7812 regulator can be used (omit the zener). Select the value of the 560-ohm resistor for your particular input voltage. The value of the 1k resistor on the 7805 regulator should be selected so that approximately 7.5 volts is present at the regulator input. A 5-volt diode could be used. However, I prefer the regulator IC because it filters out a lot of the garbage present on the supply line. The HEP170 diode is for reverse-polarity protection. The fuse can be either on the board or mounted externally. If externally mounted, install a jumper on the PC board where F1 is shown.

a kit is available

To assist those who have trouble finding parts, I've prepared a kit for the decoder. Please send checks or money order to Reliable 2-Way Radio, 513 W. 10th St., Casa Grande, Arizona 85222. Here's a list of the various options and prices:

complete kit $-$ all parts and PC board	\$140
assembled and tested	add \$165
kit — phone only (no sequential decoding)	\$130
complete kit less PC board	\$115
PC board only	\$ 25
Prices include sockets for all ICs (MACE7711) The price doesn't include edge	장애 동안 위험에 참 안 이 것 같아요?

ACF7711). The price doesn't include edge connector for board. The edge connector is a Masterite, part no. 000206-1159. The connector is available for \$6 from the above address.

If you have questions or comments about the circuit, send me a letter with an SASE for a reply. I'll be glad to answer your questions if I can. Good luck on your remote base or repeater.

references

1. Larry Nickel, W3QG, "Single IC Touch-Tone Decoder," ham radio, June, 1978, pages 26-32.

2. ACF7711 data sheets, General Instruments, 600 West John Street, Hicksville, New York 11802.

3. MK-5102-N5 data sheets, Mostek Corporation, 1215 West Crosby Road, Carrollton, Texas 75006.

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1. Hoskins, William J., W7JSW, "Simple New TT Decoder," 73, April, 1976, page 52.

2. Everhart, J.H., WA3VXH, "Toward A More Perfect Touch-Tone Decoder," 73, November, 1976, page 178.

 Buffington, E.E., W4VGZ, "Digital Autopatch," 73, April, 1977, page 166.

4. 1978 Data Catalog, page 5-70, Micro Electronics, Danbury, Connecticut.

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how to modify **Surplus cavity filters** for operation on 144 MHz

Simple conversion of surplus 417/GRC filters for top performance on 2 meters

While browsing through an electronics surplus catalog recently I came across a rather obscure item that immediately caught my attention: a small photograph of what appeared to be a dual-section resonant cavity assembly described as a "Bandpass Filter for 417/GRC Receivers." The published operating frequencies are listed in **table 1**. Since I had been playing around with homebuilt cavities for some time and was putting one to good use at my 2-meter base station, I thought it would be worthwhile to look into the surplus filters.

I ordered the F-194/U bandpass filter that covers the 2-meter Amateur band* and was very pleased to find that it was indeed a dual-section tunable cavity resonator, beautifully built both electrically and mechanically, probably at considerable government

*\$12.95 from Fair Radio Sales, Post Office Box 1105, Lima, Ohio 45802.

expense. The unit was neatly calibrated, rugged, and gold plated with low-loss Teflon insulation.

I ordered several more of the same model for experimental use but was told that that particular unit was sold out. It then occurred to me that perhaps the lower frequency units could be converted for use on the 144 MHz 2-meter band. I placed an order for the only low-frequency models that were available: F-239/U (58.5-67 MHz), F-192/U (100-121 MHz) and F-193/U (121-142 MHz). I was soon pleased to find that I was able to convert all three models for operation in the 2-meter Amateur band. The purpose of this article is to make more Amateurs aware of this unique surplus item and to outline the modification procedure for 2-meter operation.

bandpass filters

Cavity type bandpass filters have been around a long time and have been referred to by many names, such as resonant re-entrant cavity, coaxial bandpass filter, coaxial tank filter, coaxial TVI filter, tuned cavity filter, stripline filter, trough line filter, etc. Such cavities have been built in many different sizes and shapes; filter construction projects in the Amateur magazines have been based on common household items from beer cans to coffee tins, metal chassis, rectangular project boxes, and paint cans. The theory of operation of a cavity resonator can be described briefly as such: A cavity is an enclosure or partial enclosure of any size and shape having conducting walls or surfaces that can support oscillating electromagnetic fields within it and possesses certain

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resonant frequencies when excited by electrical oscillations.

Most Amateurs know that a quarter-wavelength of coaxial cable, shorted at one end, is equivalent to a parallel resonant circuit; the resonant cavity is similar and can be seen as a wide-diameter quarter-wavelength coaxial line shorted at one end, using air as the dielectric. These filters have been used for many years as the very high *Q* tank circuits in vhf and uhf transmitters and as the tuned rf circuits of receivers.

The radio-frequency current is maximum at the shorted end of the coaxial line or cavity, so a great deal of care must be exercised to ensure a good lowresistance rf contact. Because of skin-effect, copper, silver, or even gold is used to provide high conductivity.

The open end of the coaxial cavity exhibits a very high impedance and, therefore, high voltage. If construction requires insulation at the open end, care

table 1. Resonant frequencies of the surplus tunable	
bandpass filters for 417/GRC receivers.	

	MHz		MHz
F-238/U	50.0-58.5	F-196/U	184-205
F-239/U	58.0-67.0	F-197/U	205-226
F-240/U	67.0-76.0	F-199/U	224-254
F-241/U	75.0-84.0	F-200/U	254-284
F-242/U	84.0-92.5	F-201/U	284-314
F-192/U	100 - 121	F-202/U	314-344
F-193/U	121 - 142	F-203/U	344-374
F-194/U	142 - 163	F-204/U	374-404
F-195/U	163 - 184	F-236/U	550-600

must be used to select very low-loss materials so the very high Q will not be degraded.

Optimum performance can be expected from a coaxial resonant cavity when the electrical length of the cavity and its inner conductor is a full guarterwavelength long; the frequency of the cavity can be changed by varying the length of the inner conductor. In those cases where a full quarter-wavelength cavity is impractical, shorter lengths can be used with capacitance loading. A coaxial line or cavity shorter than a quarter-wavelength is electrically equivalent to an inductive reactance and requires the addition of capacitive reactance to equalize and achieve resonance; this is accomplished by adding a low-loss capacitor across the open end of the cavity. A variable capacitor provides a convenient method of tuning the cavity to the desired frequency. Generally speaking, the least amount of added capacitance results in highest Q and maximum efficiency (fig. 1).

surplus filters

The physical size of bandpass filter assemblies

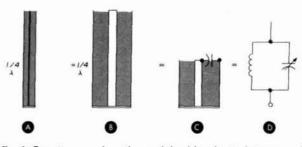
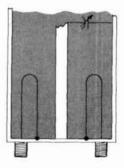


fig. 1. Quarter-wavelength coaxial cable, shorted at one end (A), is equivalent to shortened, capacitance-loaded cavity (C). Performance is similar to lumped L-C tuned circuit (D).

used in the 417/GRC receivers is the same, although the length of the cavities vary according to frequency coverage. The maximum length of the actual cavities in these assemblies is 9 1/8 inches (23.2 cm). That length is fine where it approaches a guarter-wavelength at the higher frequency ranges, but is far short of optimum at lower frequencies. Though the full 9 1/8 inches (23.2 cm) of cavity space is available, the designers did not take full advantage of the space. Shorter lengths were used on all cavities starting with the F-193/U (121-142 MHz). The F-193/U cavity, for example, is 7 3/4 inches (19.7 cm) long, and the F-194/U (142-163 MHz) is only 6 3/4 inches (17.2 cm) in length. Both the F-239/U and the F-192/U use the full 9 1/8 inches (23.2 cm). For conversion to 144-148 MHz, it is best to select one of the lower frequency units to take full advantage of the additional length. (Note that a full quarter-wavelength on two meters is approximately 20 inches [51 cm]). The reduction in length below the optimum quarterwavelength lowers the Q of the cavity, but it is still higher than the much lower Q afforded by conventional lumped L-C circuits at these frequencies.

fig. 2. Placement of loops for input/output coupling to the cavity. For light loading (highest Q), the loops must be small and spaced well away from the center conductor.



coupling

Rf energy is usually coupled into and out of this type of resonant cavity with pickup loops placed diametrically opposite each other in the electromagnetic field that exists in the shorted high-current end of the cavity (fig. 2). The loops are similar and allow the cavity to be used bilaterally. The size of the pickup loop, its position, and its proximity to the center conductor are the factors that determine the degree of coupling to the cavity. Large loops and close proximity to the center conductor provide close coupling; small loops spaced away from the center conductor result in loose coupling. Variable coupling in large commercial cavities is provided by rotatable pickup loops.

Close coupling reduces the selectivity and lowers insertion loss; conversely, loose coupling increases both selectivity and insertion loss. Two or more cavities can be cascaded for wider bandpass with steeper selectivity skirts, or to provide increased selectivity,

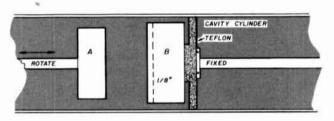


fig. 3. Original configuration of the F-193/U filter which was designed to cover 121-142 MHz; carefully remove 1/8 inch (3 mm) of material from part B to convert to two meters.

depending upon the degree of input/output coupling used in each cavity.

filter applications

Bandpass filters of the type discussed here are most often used to minimize or eliminate intermod and desense interference which originates from sources outside the desired band. On 2 meters such interference may originate in the 150-170 MHz commercial band; considerable interference also originates in the local fm and television broadcast bands and in the 120-136 MHz aircraft band. In most cases the cavity bandpass filter is placed in series with the transmission line to the antenna and attenuates all signals which fall outside its sharp passband.

As an example, I am located not far from two powerful broadcast stations, an fm station on 91.3 MHz and a Channel 2 television transmitter (video on 55.25 MHz). The two signals get into the front end of my receiver through the antenna, mix, and produce a broad, garbled fm signal centered on 146.55 MHz. Insertion of a bandpass cavity in the transmission line completely removes the offending signals and allows simplex communications on 146.55 MHz with the weakest signals.

Most Amateurs use transceivers on 2 meters so the cavity is in the transmission line during transmission as well as reception. This is a bonus because the cavity provides many benefits when used in the



fig. 4. Setup for adjusting the converted dual-section cavity of 144-MHz; tune for minimum VSWR at the desired operating frequency.

transmitted output. Spurious out-of-band emissions, including harmonics, are greatly attenuated, thus minimizing possible illegal interference with other services; this also reduces the possibility of TVI.

surplus filter conversion

Conversion of the 417/GRC bandpass filters does not require the addition of any parts. The F-193/U unit, which was originally manufactured to cover 121-142 MHz, requires only a simple modification to tune it up on the Amateur 2-meter band. Remove the six hex nuts from the rear of the assembly and carefully slide out the stationary portion of the cavities and its housing. Referring to fig. 3, part B is the fixed portion of the variable concentric capacitor which provides a fixed lumped capacitance because of its proximity to the cavity wall; variable capacitance is introduced by the movable capacitor section, part A, fig. 3, which is controlled by the front panel knob. The conversion is made by simply removing no more than 1/8 inch (3 mm) from part B (shown by the dotted line) with a hacksaw, grinder, or hand file; remove all burrs, reassemble the cavity, and the conversion is complete except for testing.

A dipmeter may be used effectively to check the

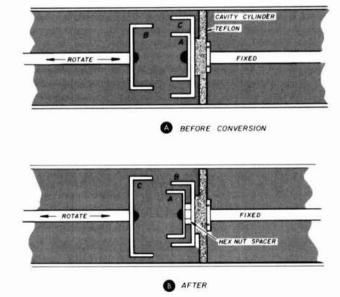


fig. 5. Conversion of the F-192/U filter (100-121 MHz) to two meters. Original arrangement before conversion is shown at (A); (B) shows the modification with a hex nut which moves the resonant frequency to the 144-MHz band.

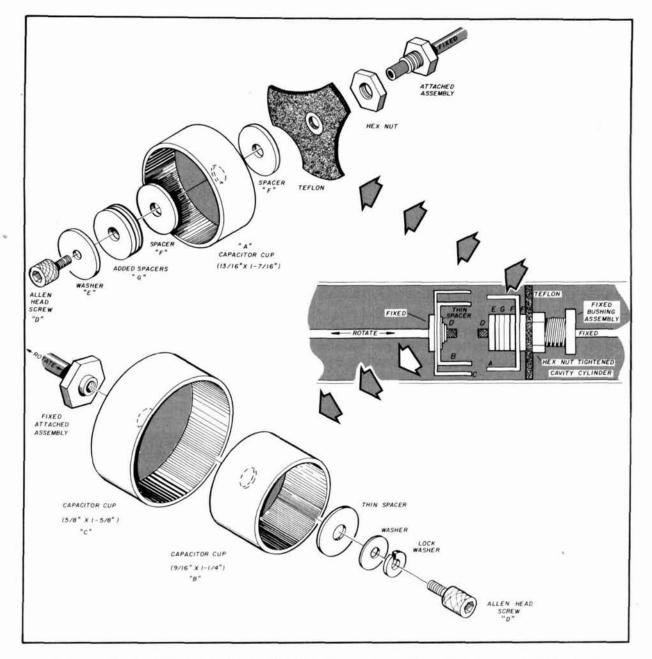


fig. 6. Modification of the F-239/U filter (58.5-67.0 MHz) for operation on the 144-MHz Amateur band (see text).

frequency range. Use a very small pickup loop connected to either port of the cavity (about 1/2 inch [12 mm] or less) and couple it lightly to the dipmeter. With the front dial turned fully clockwise, a sharp dip should occur around 130 MHz; fully counterclockwise, the dip should occur at about 155 MHz. Note that the dips are very sharp in this lightly loaded condition and can be easily missed. Check both input and output in the same way. The dial can be recalibrated in any manner you wish.

If a dipmeter is not available, a receiver can be used to find out if you are in the 2-meter range. Tune

in a rather weak signal and as you tune the cavity to resonance at that frequency, the signal should pick up nicely.

After you have determined that the conversion is successful, you can install the unit permanently in the transmission line as shown in **fig. 4**. Adjust the cavities for lowest VSWR. When transmit frequency is moved appreciably, about 500 kHz, you may have to retune the cavity for lowest VSWR. I suggest you keep the VSWR meter and the cavities as close as possible to the transceiver.

The F-192/U filter was originally designed to cover

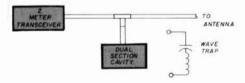
100-121 MHz. Follow the same procedure as previously when taking the assembly apart. The conversion simply involves rearranging the configuration as shown in **fig. 5**, before **A**, and after **B**. The hex nut used as a spacer is borrowed from the rear of the assembly. Make sure the Teflon insulator is reassembled with the extrusion part as shown because this provides a lowest leakage path at the high impedance point. When testing this unit, note that the capacitor will short out at the full clockwise position. The frequency range should be about 135-175 MHz.

The F-239/U filter was designed to cover 58.5-67 MHz and requires more extensive modification. Before taking the assembly apart, remove the three screws visible on the outside surface of the cavity. Carefully slide out the fixed portion of the assembly and remove all the fixed and variable capacitor sections from the center conductor and the rotor section. Select the parts needed as shown in **fig. 6** and reassemble. You will have quite a few pieces of hardware left over for your junk box. The only additional parts you will need are two or three thin washers that can be used as spacers (part **G**, **fig. 6**); as shown, the modified cavities cover 120-170 MHz. If more selectivity is desired, the pickup loops can be shortened by any convenient method.

In addition to using the dual cavity assemblies in the conventional manner as bandpass filters, they can also be used as series-resonant traps by shunting them across the line as shown in **fig. 7**; in this application the filter is used as a wave trap. If you wish, you may separate the two cavity sections by disconnecting the connecting link between them; you will then have two separate series-resonant traps as shown in **fig. 8**. You can also separate the two cavities for use as two individual bandpass filters by removing the connecting link and installing two connector sockets. Because of the limited space, I suggest you use BNC connectors.

summary

When you become familiar with the action of these fine cavity assemblies and have digested the many good articles that have appeared in the various handbooks and magazines, you will find many other uses for them around your hamshack. You could even build a duplexer for a repeater using these units as the foundation.





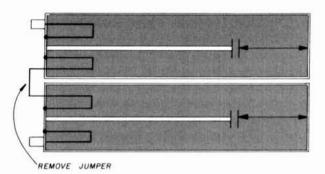


fig. 8. Separation of the two cavities for use as two tuned, single-section wave traps.

I have included a list of magazine articles from the past ten years for the convenience of those readers who wish to delve further into the subject of cavity filters.

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setter in amateur

The TS-180S with DFC (Digital Frequency Control) is Kenwood's top-of-the-line all solid-state HF SSB/CW/FSK transceiver covering 160 through 10 meters, with outstanding performance and many advanced functions, including four tunable memories to provide more operating flexibility than any other rig!

TS-180S FEATURES:

- Digital Frequency Control (DFC), including four memories and digital up/down paddle-switch tuning. Memories are usable in transceiver or split modes, and can be tuned in 20-Hz steps up or down, slow or fast, with recall of the original stored frequency. (Also available without DFC.)
- All solid-state; 200 W PEP/160 W DC input on 160-15 meters, and 160 W PEP/140 W DC on 10 meters.
- Improved dynamic range, with improved circuit design and RF AGC ("RGC"), which activates as an automatic RF attenuator to prevent receiver overload.
- Adaptable to three new bands, and VFO covers more than 50 kHz and DFC 100 kHz above and below each band. • Built-in microprocessor-controlled digital display. Shows actual
- frequency and switches to show the difference between the VFO and "MT" memory frequencies. Blinking decimal points indicate "out of band." (An analog monoscale dial is also included.)
- . IF shift (passband dialing to eliminate QRM).
- · Dual SSB filter system (second filter is optional) to provide very sharp receiver selectivity, improved S/N, and 30 dB compression with RF speech processor on transmit.

- Tunable noise blanker, to eliminate cross modulation from strong signals when noise blanker is on.
- Selectable wide and narrow CW bandwidth on receive (500-Hz CW filter is optional).
- SSB normal/reverse switch (proper sideband is automatically selected with band switch)
- Dual RIT (VFO and memory/fix)
- Available without DFC. Digital frequency display still included, with differential function showing difference between VFO and "digital hold" frequencies.

OPTIONAL ACCESSORIES:

- DF-180 digital frequency control (for TS-180S without DFC).
 YK-88CW 500-Hz CW filter.
- YK-88SSB second filter for dual-filter system.



KENWOOD ... for the discerning Amateur who demands quality.

R-1000



SP-100

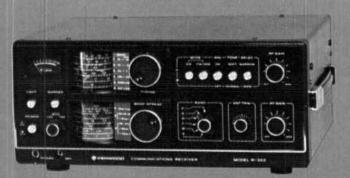
R-1000

The R-1000 is a highly advanced communications receiver. Up-conversion, PLL circuitry and other new technology provide optimum sensitivity, selectivity, and stability from 200 kHz to 30 MHz. Featuring easy-to-operate single-knob tuning and digital frequency display, it's perfect for listening to shortwave, medium-wave, and long-wave bands. Even SSB signals are received perfectly. Included is a quartz digital clock and timer.

R-1000 FEATURES:

- · Continuous frequency coverage from 200 kHz to 30 MHz.
- · 30 bands, each 1 MHz wide.
- · Five-digit frequency display and illuminated analog dial.
- · Quartz digital clock and ON/OFF timer.
- Multi-modes... AM (wide and narrow), SSB (USB and LSB), and CW.
- Three IF filters ... 2.7 kHz for SSB and CW, 6.0 kHz for AM narrow, and 12 kHz for AM wide.
- Effective noise blanker.
- · Built-in speaker.

- Three antenna terminals.
- RF step attenuator.
- Tone control.
- Recording terminal
- Remote terminal, for access to timer relay ON/OFF circuit and muting circuit.
- SSB sensitivity of 0.5 µV from 2 to 30 MHz.
- . More than 60 dB IF image ratio.
- . More than 70 dB IF rejection.



The R-300 all-band communications receiver covers 170 kHz to 30 MHz in six bands. It's ideal for listening to foreign broadcasts and other exciting transmissions throughout a wide range of the radio spectrum.

R-300

R-300 FEATURES:

- Continuous frequency coverage from 170 kHz to 30 MHz, in six bands.
- Multi-modes ... AM, SSB, and CW.
- · High sensitivity, selectivity, and image ratio
- 500-kHz marker.
- Three-way power supply (AC/batteries/external DC), with automatic switching from AC to DC in the event of AC power failure.

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TRIO KENWOOD COMMUNICATIONS INC. 1111 WEST WALNUT/COMPTON, CA 90220



Bench Power Supply

A good quality regulated bench supply is probably one of the most useful devices you can have in the workshop or ham shack. Whether you're just getting started in electronics and looking for a first project or are an experienced builder or experimenter in need of an additional bench supply, here's a supply that will fill your needs. This versatile power supply can be built in just a few hours, uses readily available components, and the finished product is a neat, professional looking package. You'll be proud to say, "I built it."

description

Quick and easy construction is a feature of this power supply, which is accomplished with a single PC board. Such construction keeps mechanical and sheet-metal work to a minimum. Most components are on the PC board. Full voltage and current metering are provided. The output is isolated from the case to allow its use as a positive or negative supply.

Output current is 3 amperes over the range of 1.5-15 volts, with short circuit and overload protection provided by the IC voltage regulator U1. Output ripple is at a low level. The power supply is physically small. The basic design isn't limited to bench-type power supply applications, because the PC board lends itself to installation in fixed equipment. No changes in the PC board pattern are required for use as a fixed supply.

the circuit

The bench supply schematic is shown in **fig. 1**. Component count is very low. The basic design con-

By Ken Powell, WB6AFT, 6949 Lenwood Way, San Jose, California 95120

sists of a full-wave power supply followed by a threeterminal IC regulator. Power transformer T1 is an 18volt 4-ampere unit. It isolates the power supply from the ac line while furnishing 18 volts to the bridge rectifier. T1 primary is fused, and the transformer frame is held at ground potential through the three-wire ac line cord. The on-off switch is connected to the output-level control for convenience and safety. The bridge rectifier is a single-unit device rated at 4 amperes 100 PIV. The pulsating dc output from the rectifier is filtered by a pair of 1500- μ F, 35-Vdc radiallead electrolytic capacitors.

The filtered dc is applied to the input of the LM-350K IC voltage regulator. The output level is controlled by a divider network formed by R1 and R2. Load current is displayed by M1, a 3-ampere dc meter. M2, a 15-volt dc meter, displays output level. The power-supply output is obtained from five-way binding posts. The output jacks, meters, and level control and switch are the only components not mounted on the PC board.

construction

PC board construction was chosen for the power supply for simplicity, ease of construction, and repeatability. Heavy glass-reinforced board should be used because the power transformer is fairly heavy and we want to avoid cracked and broken land patterns as the board ages and is subjected to stress. The etching pattern is shown in **fig. 2**.

After etching and drilling, the on-board components can be mounted and soldered as in **fig. 3**. Pay particular attention to the polarity of the capacitors and the bridge rectifier. Coat regulator U1 with heatsinking compound on the bottom surface where it mates with the heat sink. Fasten these items to the PC board as a unit with M3.5 (6-32) hardware. The regulator is not insulated from the heat sink.

This completes component mounting and soldering. Check the PC board against **fig. 3** to ascertain that all polarized components are properly oriented.

The ac cord and front-panel interconnect wiring, also shown in **fig. 3**, can be installed next. The interconnecting wiring should be no. 16 (1.6 mm) stranded to minimize voltage drop and provide flexibility. Solder these wires to the PC board and leave them about 25 cm (10 inches) long for connection to the

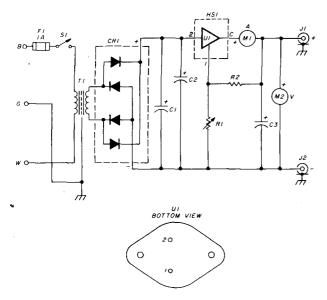


fig. 1. Schematic of the bench power supply. Basic design consists of a full-wave power supply followed by a three-terminal IC regulator.

Parts list for fig. 1:

C1, C2	1500-µF, 35-Vdc	BA 18A1506-5
C3	2.2-μF, 35-Vdc	RS 272-1407
CR1	4 amp, 100 PIV	RS 276-1171
F1	1-amp fuse	RS 270-1283
FH1	fuse holder	RS 270-739
HS1	heatsink	BA 12A2229-9
IC1	LM-350K	CSC LM-350K
J1, J2	binding posts	RS 274-662
M1	3-amp meter	CSC D1-918
M2	15-volt meter	CSC D1-920
R1	5K	RS 271-1714
R2	220, 1/2-W	RS 271-015
T1	18-volt, 4-amp	RS 273-1514
S 1	switch, spst	RS 271-1740
CASE	14.7×22.5×14.4 cm (5-7/8×9×4-7/8	RS 270-281

- inches)
- BA: Burstein-Applebee, 3199 Mercier St., Kansas City, Missouri 64111
- CSC: Circuit Specialist Co., Box 3047, Scottsdale, Arizona 85257
- RS: Radio Shack, local stores

Note: A kit is available which includes the following parts: 1 etched and drilled PC board 1 regulator, LM-350K (IC₁) 1 heatsink (HS₁) 2 capacitors, 1500 μ f/35 Vdc (C₁ and C₂) 1 resistor, 220 ohms (R₂) 1 capacitor, 2.2 μ f/35 Vdc (C₃)

The cost of this kit is \$21.00 plus \$1.50 postage and handling. Ask for kit #PS-25-3 from J. Oswald, 1436 Gerhardt Avenue, San Jose, California 95125.

front panel components. They can be trimmed to length as the front panel is installed.

Drill and punch the front panel using caution to avoid damaging the smooth finish on the panel. If a chassis punch isn't available for making the meter holes, a coping saw will work fine in the soft aluminum. When you're satisfied that all the front panel components fit well, label the on-off level control with dry transfer labels. Apply a coat of clear acrylic to the panel to protect the lettering. The front panel components can be mounted after the panel is dry.

At this time the PC board should be mounted to the cabinet base. Use standoffs to elevate the board above the base. Cut a small notch into the lower edge of the cabinet back and install a strain-relief or grommet for the ac line cord. Lay the front panel down in front of the PC board and complete the interconnecting wiring as in **fig. 3**. This wiring can be laced or spot-tied for a neat appearance (photo). The cabinet can now be assembled. Your project should look pretty much like the photo of the completed prototype. If a cabinet is used other than that listed in the parts list, you may have to modify the construction procedure a bit, but the end result should be the same.

test and applications

After assembly, the front-panel meters should be adjusted for mechanical zero. There are no other adjustments, so the little supply is now ready for the smoke test. Apply ac power and advance the output level control clockwise. The voltmeter should track nicely from about 1.5 to slightly over 15 volts. Output ripple can be checked on a scope if you have one, but, if not, don't be overly concerned about it. The two prototype units I built were smooth and ripple free.

The little supply is now ready to go to work. The voltage and current range will let you power up quite a variety of equipment and circuits. I've used the

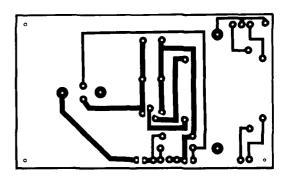


fig. 2. Full-size printed-circuit layout; component placement is shown in fig. 3.



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ANTENNA COMPONENTS

Antenna wire, stranded #14 copperweld\$.06 ft.
Antenna Wire, stranded #15 copperweld
Antenna wire, stranded #16 copperweld
Van Gorden HI-Q Baluns, 1:1 or 4:1 9.95 ea.
Unadilla/Reyco, W2AU Baluns, 1:1 or 4:1 14.95 ea.
Van Gorden HI-Q center insulators 4.95 ea.
Unadilla/Reyco, W2AU center insulators 9.75 ea.
Ceramic "Dogbone" end insulators, pair
Unadilla/Reyco plastic end insulators, pair 3.50
Nylon guy rope, 450 lb. test, 100' roll 3.49
Poly guy rope, 275 lb. test, 100' roll 3.00
Unadilla/Reyco W2VS Traps, KW-10 thru KW-4021.95 pr.
Belden 8214 RG-8U type foam coax
Belden 8219 RG-58 A/U foam coax 11 ft.
Berk-Tex 6211 RG-8X foam coax., Ultraflexible
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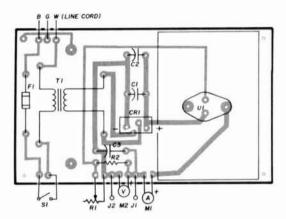


fig. 3. Component side of the circuit board; photographs show placement of the power transformer, heatsink, and other larger components.

supply on TTL projects, op amps, vhf rigs, and a number of auto radio and tape player combinations.

The properties designed into the IC regulator take most of the worry out of the unit. If you apply a shorted piece of gear to the output, or connect a rig that draws more current than the supply can furnish, the regulator will shut down and no damage will occur. If the unit gets too hot, it will shut down. When normal conditions are restored, the supply will bounce back and go to work again. It's a tough little performer and just the ticket for a guy like me, who's apt to make an error now and then.

conclusion

There are many variations that could be made in the supply to suit individual requirements. Components aren't critical, although I don't think I'd go to lower values in any case. Changes such as using different meters with shunts and multipliers would be no problem, and a single meter with shunts, multipliers, and switching could be used. For fixed applications the level control could be replaced with a fixed resistor and the meters eliminated. Layout and mechanical details aren't critical, but the regulator IC should be kept physically close to the filter capacitors.

The cabinet should be well ventilated. If the supply is powered up outside the case, keep in mind that 117 Vac is on the PC board and the interconnecting wiring! The low voltages usually associated with solid-state equipment tend to encourage carelessness.

The supply was a very enjoyable project in that the assembly time was minimal (about three hours) and no sophisticated equipment was required for building, de-bugging, or adjustment. I don't think anyone should have any apprehension about trying this one for a weekend project.

ham radio



Requires 12"x12"x36" of concrete.)

GENERAL FEATURES

The Hinged Base Plate allows tower to be tilted over for access to antenna and rotor from the ground.

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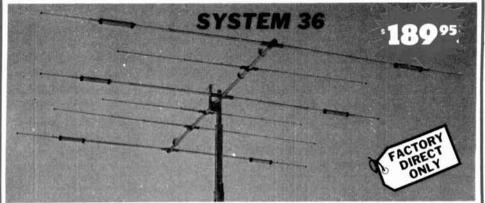
All towers use high strength heavy galvanized steel tubing that conforms to ASTM specifications for years of maintenance-free service. The large diameters provide unexcelled strength. All welding is performed with state-of-the-art equipment. Top sections are 2" O.D. for proper antenna/rotor mounting. A 10' push-up mast is included in the top section of each tower. Hinge-over base plates are standard with each tower. The high loads of today's antennas make Wilson crank-ups a logical choice. Prices and specifications subject to change without no

(Requires 18"x18"x48" concrete.)

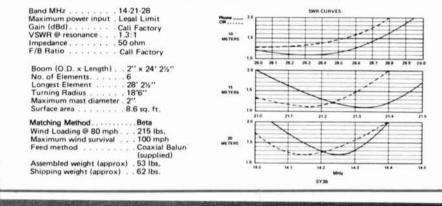
TILT-OVER BASES FOR TOWERS ROTATING BASE FIXED BASE The RB Series was designed for the Amateur who wants the add-The FB Series was designed to provide an economical method of ed convenience of being able to work on the rotor from the moving the tower away from the ground position. This series of house. It will support the tower in bases will give that ease plus roa completely free-standing vertical tate the complete tower and anposition, while also having the tenna system by the use of a capabilities of tilting the tower heavy duty thrust bearing at the over to provide an easy access to the antenna. The rotor mounts at base of the tower mounting position, while still being able to tilt the top of the tower in the conthe tower over when desiring to ventional manner, and will not romake changes on the antenna Tilting the tower over is a tate the complete tower. system one-man task with the Wilson bases. (Shown above is FB-45B ... \$114.95 RB-45B ... \$179.95 the RB-61B. Rotor is not FB-61B ... 169.95 RB-61B... 249.95 included.) ***NEW*** ORDER WILSON ST64 - 64 ft. Stacking Tower ST77 - 77 ft. Stacking Tower FACTORY DIRECT SYSTEMS, INC. CALL FOR INFORMATION

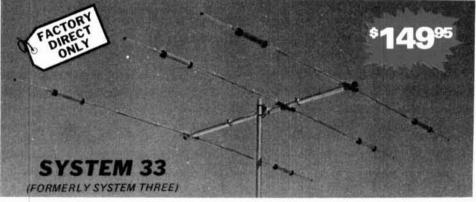
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WILSON SYSTEMS INC. MULTI-BAND ANTENNAS



A trap loaded antenna that performs like a monobander! That's the characteristic of this six element three band beam. Through the use of wide spacing and interlacing of elements, the following is possible: three active elements on 20, three active elements on 15 and four active elements on 10 meters. No need to run separate coax feed lines for each band, as the bandswitching is automatically made via the High-Q Wilson traps. Designed to handle the maximum legal power, the traps are capped at each end to provide a weather-proof seal against rain and dust. The special High-Q traps are the strongest available in the industry today.





Capable of handling the Legal Limit, the "SYSTEM 33" is the finest compact tri-bander available to the amateur. Designed and produced by one of the world's largest antenna manufacturers, the traditional quality of workmanship and materials excells with the "SYSTEM 33". New boomto-element mount consists of two 1/8" thick formed aluminum plates that will provide more clamping and holding strength to prevent element misalignment. Superior clamping power is obtained with the use of a rugged 1/4" thick aluminum plate for boom to mast mounting. The use of large diameter High-Q traps in the "SYSTEM 33" makes it a high performing tri-bander and at a very economical price. A complete step-by-step illustrated instruction manual guides you to easy assembly and the lightweight antenna makes installation of the "SYSTEM 33" quick and simple.

SPECIFICATIONS .

2" x 14'4'

" O.D

5.7 sq. ft

27'4"

15'9'

Band MHz Maximum power input Gain (dbd) VSWR at resonance Impedance F/B Ratio 14-21-28

1.3:1

50 ohm

Legal Limit

Call Factory

Call Factory

4286 S. Polaris Ave., Las Vegas, Nevada 89103

Prices and specifications subject to change without notice

Boom (O.D. x length) No. of elements Longest element. Turning radius Maximum mast diameter Surface area

WILSON

SYSTEMS, INC





WV-1A 4 BAND TRAP VERTICAL (10 - 40 METERS)

No bandswitching necessary with this vertical. An excellent low cost DX antenna with an electrical quarter wavelength on each band and low angle radiation. Advanced design provides low SWR and exceptionally flat response across the full width of each band.

Featured is the Wilson large diameter High-Q traps which will maintain resonant points with varying temperatures and humidity.

Easily assembled, the WV-1A is supplied with a hot dipped galvanized base mount bracket to attach to vent pipe or to a mast driven in the ground.

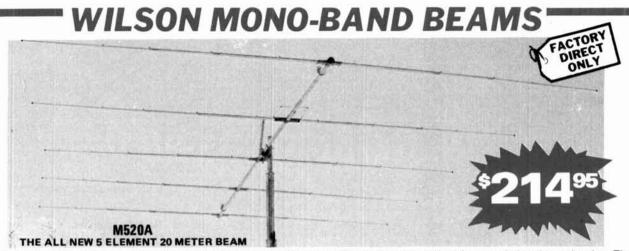
Note: Radials are required for peak operation. (See GR-1 below)

SPECIFICATIONS

- 19' total height
 Self supporting no guys required
- Weight 14 lbs.
- Input impedance: 50 Ω
- Powerhandling capability: Legal Limit
- Two High-Q traps with large diameter coils
- Low angle radiation
 Omnidirectional
- Omnidirection
 performance
- Taper swaged aluminum tubing
- Automatic bandswitching
- Mast bracket furnished
 SWB: 1 1:1 or less on all
- bands



The GR-1 is the complete ground radial kit for the WV-1A. It consists of: 150' of 7/14 stranded copper wire and heavy duty egg insulators, instructions. The GR-1 will increase the efficiency of the GR-1 by providing the correct counterpoise.



At last, the antennas that you have been waiting for are here! The top quality, optimum spaced, and newest designed monobanders. The Wilson System's new Monoband beams are the latest in modern design and incorporate the latest in design principles utilizing some of the strongest materials available. Through the select use of the current production of aluminum and the new boom-to-element plates, the Wilson Systems' antennas will stay up when others are falling down due to heavy ice loading or strong winds. Note the following features:

1. Taper Swaged Elements – The taper swaged elements provide strength where it counts and lowers the wind loading more efficiently than the conventional method of telescoping elements of different sizes.

2. <u>Mounting Plates – Element to Boom</u> – The new formed aluminum plates provide the strongest method of mounting the elements to the boom that is available in the entire market today. No longer will the elements tilt out of line if a bird should land on one end of the element.

3. Mounting Plates - Boom to Mast - Rugged 1/4" thick aluminum plates are used in combination with sturdy U-bolts and saddles for superior clamping power.

4. <u>Holes</u> – There are no holes drilled in the elements of the Wilson HF Monobanders. The careful attention given to the design has made it possible to eliminate this requirement as the use of holes adds an unnecessary weak point to the antenna boom.

With the Wilson Beta-match method, it is a "set it and forget it" process. You can now assemble the antenna on the ground, and using the guide-lines from the detailed instruction manual, adjust the tuning of the Beta-match so that it will remain set when raised to the top of the tower.

The Wilson Beta-match offers the ability to adjust the terminating impedance that is far superior to the other matching methods including the Gamma match and other Beta matches. As this method of matching requires a balanced line it will be necessary to use a 1:1 balun, or RF choke, for the most efficient use of the HF Monobanders.

The Wilson Monobanders are the perfect answer to the Ham who wants to stack antennas for maximum utilization of space and gain. They offer the most economical method to have more antenna for less money with better gain and maximum strength. Order yours today and see why the serious DXers are running up that impressive score in contests and number of countries worked.

Band Mun		F/B Ratio	-		Impedance	Matching	Elementa	Longest	Boom O.D.	Boom Length	Turning Radius	Aras Aras (Sq.Ft.)	. SO mph	Maximum Mast	(1.84.)
20		25 dB	500 KHz	1.1:1	50 Ω	Beta	5	36'6"	2"	34'2%"	25'1"	8.9	227	2"	68
				1.1:1	50 Ω	Beta	4	36'6"	2"	26'0"	22'6"	7.6	189	2"	50
					50 Ω	Beta	5	25'3"	2"	26'0"	17'6"	4.2	107	2"	41
	-				50 Ω	Beta	4	24'2%"	2"	17'0"	14'11"	3.1	54	2"	25
	-			-	50 Ω	Beta	5	18'6"	2"	26'0"	16'0''	2.8	72	2"	36
				-	50 Ω	Beta	4	18'3"	2"	12'11"	11'3"	1.4	36	2"	20
	20 20 15 15 10	20 11.5 20 10.0 15 12.0 15 10.0 10 12.0	Wei dBd Ratio 20 11.5 25 dB 20 10.0 25 dB 15 12.0 25 dB 15 10.0 25 dB 10 12.0 25 dB	20 11.5 25 dB 500 KHz 20 10.0 25 dB 500 KHz 15 12.0 25 dB 400 KHz 15 10.0 25 dB 400 KHz 10 12.0 25 dB 400 KHz 10 12.0 25 dB 15 MHz	20 11.5 25 dB 500 KHz 1.1:1 20 10.0 25 dB 500 KHz 1.1:1 15 12.0 25 dB 400 KHz 1.1:1 15 10.0 25 dB 400 KHz 1.1:1 15 10.0 25 dB 400 KHz 1.1:1 10 12.0 25 dB 1.5 MHz 1.1:1	20 11.5 25 dB 500 KHz 1.1:1 50 Ω 20 10.0 25 dB 500 KHz 1.1:1 50 Ω 15 12.0 25 dB 600 KHz 1.1:1 50 Ω 15 12.0 25 dB 400 KHz 1.1:1 50 Ω 15 10.0 25 dB 400 KHz 1.1:1 50 Ω 10 12.0 25 dB 105 KHz 1.1:1 50 Ω	Band Merry Gain dBd F/B Ratio Automatic Vision Vision of Maching Imputation Matching 20 11.5 25 dB 500 KHz 1.1:1 50 Ω Beta 20 10.0 25 dB 500 KHz 1.1:1 50 Ω Beta 15 12.0 25 dB 400 KHz 1.1:1 50 Ω Beta 10 12.0 25 dB 400 KHz 1.1:1 50 Ω Beta 10 12.0 25 dB 1.5 MHz 1.1:1 50 Ω Beta	Band Mark Gain Bdd F/B Ratio Automatic Vision Magnetic Magnetic Implements Matching Emerget Emerget 20 11.5 25 dB 500 KHz 1.1:1 50 Ω Beta 5 20 10.0 25 dB 500 KHz 1.1:1 50 Ω Beta 4 15 12.0 25 dB 400 KHz 1.1:1 50 Ω Beta 5 10 12.0 25 dB 15 MHz 1.1:1 50 Ω Beta 4 10 12.0 25 dB 1.5 MHz 1.1:1 50 Ω Beta 5	Base Gain F/B Plantage Vorme Impedance Matching Element Longest 20 11.5 25 dB 500 KHz 1.1:1 50 Ω Beta 5 36'6'' 20 10.0 25 dB 500 KHz 1.1:1 50 Ω Beta 4 36'6'' 15 12.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 26'3'' 15 10.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 24'2''' 10 12.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 24'2''' 10 12.0 25 dB 1.5 MHz 1.1:1 50 Ω Beta 5 18'6''	Base Gain F/B Production Views Matching Matching Element Longest Boom 20 11.5 25 dB 500 KHz 1.1:1 50 Ω Beta 5 36'6'' 2'' 20 10.0 25 dB 500 KHz 1.1:1 50 Ω Beta 4 36'6'' 2'' 15 12.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 36'6'' 2'' 15 10.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 24'2½'' 2'' 10 12.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 24'2½'' 2'' 10 12.0 25 dB 1.5 MHz 1.1:1 50 Ω Beta 5 18'6'' 2''	Basic F/B Retio ****** ******************************	Basic F/B Participation Marching Example Longest Boom Boom Furning 20 11.5 25 dB 500 KHz 1.1:1 50 Ω Beta 5 36'6'' 2" 34'28'' 25'1'' 20 10.0 25 dB 500 KHz 1.1:1 50 Ω Beta 4 36'6'' 2" 34'28'' 25'1'' 15 12.0 25 dB 600 KHz 1.1:1 50 Ω Beta 4 36'6'' 2" 34'28'' 25''' 15 10.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 24'23'' 2''' 17''' 14''11'' 10 12.0 25 dB 1.5 MHz 1.1:1 50 Ω Beta 5 18'6'' 2''' 26'0'' 16'''' 10 12.0 25 dB 1.5 MHz 1.1:1 50 Ω Beta 5 18'6''' 2''' 26'''' 14''1''''	Band F/B Participation Matching Element Longest Boon Boon Ratios Fundamente 20 11.5 25 dB 500 KHz 1.1:1 50 Ω Beta 5 36'6'' 2" 34'2%' 25'1'' 8.9 20 10.0 25 dB 500 KHz 1.1:1 50 Ω Beta 4 36'6'' 2" 34'2%' 25'1'' 8.9 15 12.0 25 dB 600 KHz 1.1:1 50 Ω Beta 4 36'6'' 2" 26'0'' 22''' 7.6'' 15 12.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 24''' 2''' 17'6'' 4.2 15 10.0 25 dB 400 KHz 1.1:1 50 Ω Beta 4 24'2%'' 2''' 17'0'' 14'11'' 3.1 10 12.0 25 dB 1.5 MHz 1.1:1 50 Ω Beta 5 18'6''' 2''' 26'0''	Number Gain F/B Number Number Importance Matching Emmeth Longest Boor Longest Boor Lungth Turning Suffer Weishaud 20 11.5 25 dB 500 KHz 1.1.1 50 Ω Beta 5 366" 2" 34'2%' 25'1" 8.9 227 20 10.0 25 dB 500 KHz 1.1.1 50 Ω Beta 4 36'6" 2" 34'2%' 25'1" 8.9 227 15 12.0 25 dB 600 KHz 1.1.1 50 Ω Beta 4 36'6" 2" 26'0" 7.6" 4.2 107 15 10.0 25 dB 400 KHz 1.1.1 50 Ω Beta 4 24'2%'' 2" 17'0" 14'11" 3.1 54 10 12.0 25 dB 1.5 MHz 1.1.1 50 Ω Beta 5 18'6" 2" 26'0" 14'11" 3.1 54 <t< td=""><td>Base Gein F/B F/B Town business Matching Example Longest Boom Boom</td></t<>	Base Gein F/B F/B Town business Matching Example Longest Boom Boom

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voltage-controlled resistance for Wien Bridge oscillators

A simple improvement that uses a photocell as a voltage-controlled resistor

Wien bridge oscillators^{1,2} require some sort of voltage-controlled resistance in the feedback network to maintain unity loop gain as the frequency is varied over a wide range. In practice, part of the oscillator output voltage is used to control the variable resistance element so that the output voltage remains constant, or nearly so. The voltage-controlled resistance must not respond to the output-waveform instantaneous value, but only to its average amplitude. Thus, the voltage-controlled resistance device must have a long time constant compared with the lowest-frequency period.

I've experimented with these circuits over the years, using incandescent lamps, diodes, and thermistors.³ These methods all have shortcomings of one kind or another, and I think the idea described here works better than any of those others.

homebrew device

Shown in **fig. 1** is a simple voltage-controlled resistor that can be homebrewed with readily available parts. Such devices may be available already assembled but perhaps not as available as the two parts needed to build your own. The controlled resistor is a cadmium sulfide (CdS) photocell, obtainable from Radio Shack for 99 cents (catalog number 276-116). Its resistance changes from several megohms in darkness to about 100 ohms in bright light. It's most sensitive to yellow or green light. A green light-emitting diode (LED) causes the CdS resistance to vary inversely to the dc control voltage applied to the LED circuit.

To prevent ambient light from interfering with operation, the LED-CdS assembly should be sealed as shown in **fig. 2**. The cover may be almost any type of scrap material, such as a square of copper-clad circuit board. Drill a hole in the cover for the LED to stick through into the photocell, then epoxy the parts together and spray thoroughly with black paint. Be sure to mark the anode lead of the LED before epoxy is applied.

circuit details

Fig. 3 shows the complete oscillator. U1 is a dual jfet op amp, TL072CP, made by Texas Instruments. It gives excellent performance in this circuit. The oscillator signal at pin 7 of U1A is buffered and detected to provide a dc control voltage for the LED. The

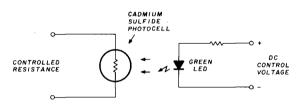


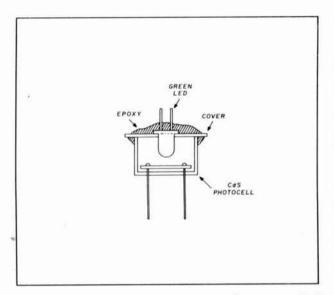
fig. 1. Homebrew voltage-controlled resistor.

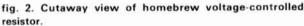
1N914 is required because the emitter-base breakdown rating of the 2N2222 is only 5 volts. The 10ohm resistor in series with the filter capacitor is required for stability. The CdS photocell is connected into the oscillator circuit in the same way a thermistor would be used. Other information on Wien bridge oscillators may be found in the references cited.

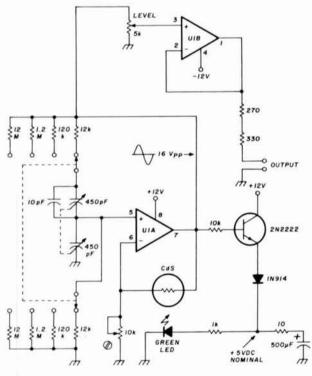
operation

The 10-k trimpot is adjusted for maximum undistorted output in the normal manner. If the trimmer capacitors on the tuning capacitor are properly adjusted, output amplitude should be flat within about 0.1 dB between 30 Hz and 200 kHz. If frequencies lower than 30 Hz are required, the $500-\mu$ F filter capacitor must be increased in value to prevent distortion.

By Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75248







U1 jfet op amp, TLO72CP. For availability contact T.I. Supply Co., 6000 Denton Drive, Dallas, Texas 75235.

fig. 3. Wien bridge oscillator with improved gain control circuit.

references

1. H. Olson, W6GXN, "Resistance-Capacitance Oscillators," ham radio, July, 1972, pages 18-25.

2. H. Olson, W6GXN, "A Transistor Wien Bridge Oscillator," 73 April, 1967, pages 36-39.

3. C. Hall, WA5SNZ, "New Op Amp Challenges the 741," ham radio, January, 1978, pages 76-78.

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a carrier-operated relay for your Heath HW-2036

Simple mods you can make to enjoy your a-m/fm radio and your 2-meter rig at the same time

If you have problems doing more than two things at once, then this modification is for you. Try it if you own a HW-2036 2-meter rig and have trouble trying to decide between listening to an interesting conversation on the rig and soothing background music from your a-m/fm radio while traveling down the road.

I've installed a COR (carrier operated relay) in the 2036 to switch the a-m/fm radio on and off when the 2036 is receiving a signal. The a-m/fm set goes off

when a signal is received on the 2036, which results in only one radio being on at a time - a real blessing for eliminating the hectic atmosphere in an automobile when both radios are running simultaneously and you are in rush-hour traffic.

operation

The circuit is simple and I'm surprised that Heath hasn't made it an available option on the 2036. The basic idea is to turn off the a-m/fm radio during the transmit and receive modes of the 2036.

The benefit of turning off the a-m/fm radio during the transmit is twofold. First, there's no possibility of transmitting your soothing background music over the 2-meter rig, which is not looked upon favorably by the FCC, to say the least. The second benefit is that listening to the a-m/fm radio is automated while going down the road.

circuit

The diodes (fig. 1) are used solely for isolation to keep this circuit from affecting anything going on at the connection points in the 2036. The RC network on the relay gives a delay of 1-2 seconds to keep the a-m/fm radio off during short repeater dropouts or when working simplex.

By Gary L. Long, WD5HYQ, Route 5, Box 267, Muskogee, Oklahoma 74401

construction

I located all the parts beside the lever switches on the 0/5-kHz switch side of the HW-2036. A little Super Glue will hold everything in place for wiring. The transistor isn't critical; just be sure it can handle enough source current for the relay.

0/5-kHz – COR bypass. If you live in a signal area with very few repeaters that require the 0/5-kHz switch, you can make it a dual function switch (0/5kHz – COR bypass). This beats installing another switch in the rig and messing up the looks of the 2036. This addition is useful when the a-m/fm radio is broadcasting something you don't want interrupted while listening to it. All that's necessary is to remove the wire going to SW6 from pin X on the synthesizer board and reconnect it through an isolation diode to the switch as shown in fig. 1.

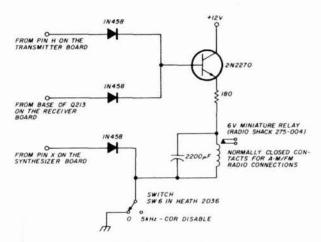


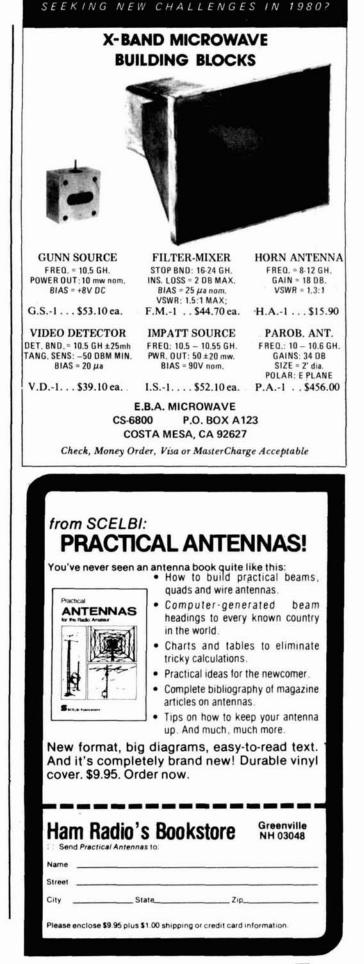
fig. 1. Circuit for turning off your a-m/fm radio during transmit-receive modes of the Heath HW-2036 2-meter transceiver. Circuit features a 0/5-kHz — COR bypass arrangement.

A-m/fm connections. All that's left to do is to connect a pair of wires to the relay contacts and feed them out through the same slot in the back of the radio where the power cable comes out. The easiest way to connect these wires is to pull the a-m/fm radio fuse and connect these wires right at the fuse block. Don't forget to install an in-line fuse on one of the wires before you turn on the power.

one last item

I had been plagued with a birdie problem, and while I had the rig apart I decided to go hunting for birdies. This critter appeared on 146.85 MHz, 146.52 MHz, and several other places. My solution was to put a mica capacitor across the base-collector junction of Q409 on the synthesizer board. The value appears to be critical. The best response was obtained in my rig by using about 820 pF.

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plasma-diode experiments

Plasma diode detector using a neon glow lamp today's answer to the slotted-line standing-wave detector

The use of the plasma in glow lamps and specially designed discharge tubes for microwave detection is not new.¹⁻⁴ In most cases, means are provided for focusing the microwave energy on a selected point of the discharge, with the detector output appearing as a variable component of the lamp current.

The plasmas operate at quite high power levels as far as biasing is concerned (20 mA and 250 volts).

This type power supply is out of line for the experimenter working with transistors and ICs.

The drawback of the conventional kind of plasma detector is its high noise. Relying on Maxwell's displacement current, research workers have, however, constructed glow-lamp microwave detectors that are free from the annoying lamp-current noise. In certain cases they are even more sensitive than the old standby, the crystal diode. These new detectors operate in the gigahertz region up to 100 GHz (3-mm wavelength).

starvation-current mode

A recent research and development project at Sercolab revealed that worthwhile detector sensitivity can be achieved in a "starvation current" glow-lamp mode, with the demand on supply power reduced to about 0.1 per cent. This kind of operation brings the glow-lamp microwave detector within every experimenter's reach, since the entire power supply consists of a few 9-volt transistor batteries. With capacitively coupled microwave energy, the glow-lamp detector combines an acceptable signal-to-noise ratio with a broad frequency range, a wide dynamic range, and the feature of being almost indestructible — something that can't be claimed for its competitor, the crystal diode.

By Dr. Harry E. Stockman, Sercolab, P.O. Box 78, Arlington, Massachusetts 02174

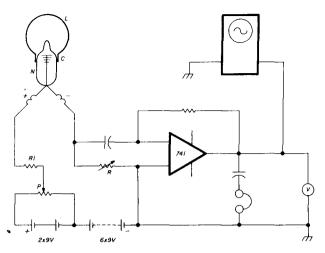


fig. 1. Experimental circuit for evaluating plasma diodes in the "starvation-current" mode. LC is the loop antenna; N is a neon lamp. Other components are discussed in the text.

Its characteristic of being able to take a heavy overload without showing any defect in operation makes the glow lamp a handy microwave detector. Furthermore, it's inexpensive. A brand-new neon lamp costs only 25 cents. (Radio Shack's NE 2 and NE 2H are good examples.)

Today, triode and tetrode glow lamps, such as the TRJ 250, are being used successfully as microwave detectors, providing the desired coupling internally.* Alternatively, an external capacitor plate can be attached to a diode, cemented to the glass envelope. In the detector described below, a diode lamp with protruding terminal wires is used with an external rfelectrode arrangement that forms a loop antenna. A dipole antenna with capacitor plates provides an alternative.

test setup

Fig. 1 shows a simple test rig. In the starvation mode, only the tip of the cathode electrode glows. The capacitor plates, C, of the loop antenna, L, are loosely coupled to the discharge. (We are looking into the glow lamp, N, in the plane of the two electrodes.) Each lamp terminal wire is formed into two solenoid turns around any temporary 3-mm (1/8-inch) core, thus providing an rf choke.

Potentiometer P is a small 25k trimpot. Variable resistor R is 250k. The values are not critical. R1 is a 5k safety resistor. A 741 IC is the detector — output amplifier; but two 741s in cascade are better, meaning that we may use a 1458 IC.

The output indicators are a scope, an electronic voltmeter, and earphones. Eight 9-volt batteries are

used, held together with a rubber band. The lampsupply current amounts to only 0.1 mA, so the life of the major part of the battery is almost the shelf life.

the antenna

The experimenter can readily make up an assortment of loop antennas by clipping metal strips from a coffee can, or better, from a neatly polished, somewhat thicker, copper sheet. The width of the strips may be 6 mm (1/4 inch).

The quickest way to obtain a properly tuned system is to try out different-size loops for a given wavelength. The inductance can be reduced somewhat by flattening the loop. The loop position relative to the glow lamp is shown in **fig. 1**. This is only one of several possible positions. Actually, the best results were obtained with the loop folded 90 degrees out of the paper (**fig. 1**), so that its axis is parallel with the direction of the electrode system. The loop is then moved up, down, and lengthwise for optimum coupling to the plasma. For the loop orientation with respect to the transmitter, the rules are about the same as for everyday shortwave work.

test equipment and adjustments

Two low-power laboratory oscillators were available for the tests — one 1.2 GHz (25-cm wavelength) and a 2.4 GHz (12.5-cm wavelength). Each was amplitude modulated with a 1000-Hz tone.

In adjusting the *receiver*, the *P* and *R* controls are set in a combination that does not promote CR relaxation oscillations, which are common in neon-lamp hookups. The adjustment of the controls is such that maximum detector audio output is provided.

Once the proper control setting has been found, the detector will remain stable. If, after a long time of operation, the lamp goes out, potentiometer P voltage is temporarily increased to make the lamp fire. Then it is decreased to its previous value. Most lamps fire in the 60-70 volt range. The extinguishing voltage is then some 10 volts or less.

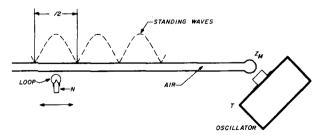


fig. 2. Plasma detector of fig. 1 implemented for plotting standing waves on a transmission line. Loop must be moved along the line at a constant distance from the line for consistent data.

^{*}The TRJ 250 tube and other glow lamps suitable for microwave detection are obtainable from General Instrument Signalite Division, 1933 Heck Avenue, Neptune, New Jersey 07753.



uses

Among practical uses of this detector is the determination of the oscillator wavelength. The glowlamp detector is placed at the side of the transmitter with just enough coupling for the modulation tone to be heard in the earphones. A meter stick is then placed vertically on the bench, and a reflector is moved up and down along the stick. Readings are taken at every sharp null on the meter or in the earphones. A suitable reflector may be made from heavy aluminum foil cemented to the back of a writing-pad cardboard.

A bigger reflector is better. One with a quartermeter side is a deluxe article. The average distance between the nulls is one-half the wavelength. Really, doing it this way we're too close to the transmitter, and if the experiment is repeated in the horizontal plane, with a larger distance between transmitter and receiver, better accuracy results.

In another experiment, we may rig up a two-wire line in air with a wire distance of 13 mm (1/2 inch), using one end for coupling and the other end either open or closed. Another arrangement is to put a bit of TV downlead on a wooden table, similarly arranged at the ends.

Fig. 2 shows how the excited end of the line is formed into a loop, coupled through the mutual impedance, Z_M , to the oscillator. With the line removed, the transmitter is tuned for minimum direct pickup by the receiver. Then, with the line in position, Z_M is assigned the compromise value that gives a clear standing-wave pattern with only sufficient rf power for good deflections on the voltmeter.

The detector is then moved along the line and the readings are tabulated, whereupon the standing-wave pattern can be plotted. Loop L must be moved at a constant distance from the line.

The two-wire air line gives sharp nulls, and the wavelength obtained agrees quite well with that measured in the preceding reflector experiment. The TV cable nulls are not equally sharp, and the measured wavelength is smaller than the true wavelength because the insulation dielectric constant is larger than unity.

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ham radio

64 hr february 1980



90-degree phase-shift network offers 2:1 bandwidth

Differential 90-degree phase-shift network built with coaxial line sections provides wide operating bandwidth

This brief note describes an easy-to-build network capable of providing 90-degree differential phase shift between two outputs. Over a 2:1 bandwidth the VSWR maximum is less than 1.2:1 and the deviation from quadrature phase is ± 2 degrees. This network should find application in antenna feed configurations and other instances where extremely broadband differential phase shift is not required.

network description

The phase-shift network is $\lambda/2$ long at band center and is shunted at midpoint with an $\lambda/8$ shorted stub and an $\lambda/8$ open stub. For a 50-ohm system the characteristic impedance of the $\lambda/2$ line and both stubs is 31 ohms. A sketch of the circuit is shown in **fig. 1**. For clarity, balanced transmission is shown although actual realization most likely will be coaxial line. The other half of the differential phase shift network is $3/4 \lambda$ length of 50-ohm transmission line. The complete network is shown in **fig. 2**.

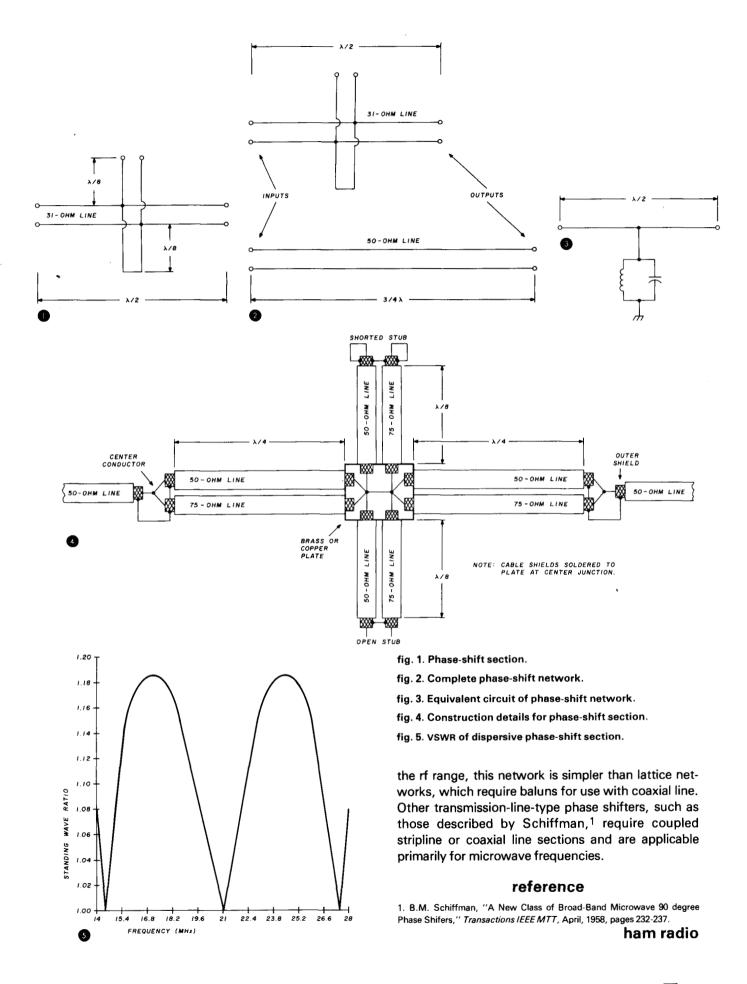
The network in **fig. 1** has a dispersive phase characteristic, which means that the phase shift through the network does not vary linearly with frequency. An equivalent circuit of a dispersive network is shown in **fig. 3**. The phase through the network is retarded below band center and is advanced above band center. At band center the parallel circuit is resonant and has no effect. In this manner, a constant 90-degree phase differential is maintained between the dispersive network and the $3/4 \lambda$ transmission line for an octave bandwidth. Optimum impedance match for the network is achieved by using 31-ohm impedance line.

Transmission line of 31-ohm impedance is not available unless specially made; however, a 30-ohm line can be realized by paralleling lengths of 50- and 75-ohm line. **Fig. 4** suggests one method you may use to build the circuit. **Fig. 5** is a computed plot of VSWR versus frequency for the dispersive network made from 30-ohm transmission line.

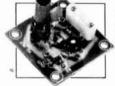
conclusions

This note describes a simple 90-degree phase-shift network, which is easy to build from readily available materials and adequate for 2:1 bandwidths. In instances where only 2:1 bandwidths are required in

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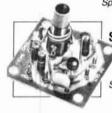
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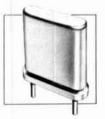
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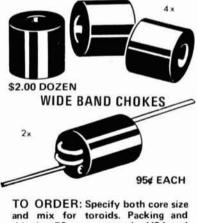
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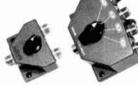
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CN-630

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comments

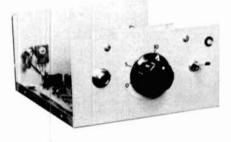
(Continued from page 6)

U4) to the top end of diode CR5 near the lower right hand corner of U13. Reconnect the clock control lead to either end of this new jumper. Finally, cut away a short segment of the foil to open the path between U13 pin 10 and CR5.

The effect of these changes is to delay the buildup at U13B pin 10 of the high on the clock control line until U3B has time to take U13B pin 9 low. The resistor and capacitor values are not too critical, and if this doesn't cure the click completely, try a slightly higher resistor value (up to about 270 ohms).

> Robert C. Cheek, W3VT Tokyo, Japan

split-band speech processing



Dear HR:

This note is prompted by Wes Stewart's article in *ham radio* for September, 1979, which I liked very much for his understanding of the subject and the unmasking of some of its secrets. However, there are a couple of points which can stand clarification; they have been alluded to in the original article describing the process (Fisk, *ham radio* June, 1976) but not discussed in detail.

The first thing I noticed in Stewart's design is the measured distortion, which is about twice as high as that of the *Vomax* unit by Maximilian Associates. The detrimental effect of distortion on the processing gain has received considerable attention in both my own (*ham radio* February, 1971) and Fisk's article; Stewart's design will probably show 2 dB less talk power gain than the commercial device, for which 10 dB is claimed. Explanations of the causes of distortion are a little tedious, but not really difficult.

Stewart's band splitting filters have a Q of 1.84; in contrast, the Vomax design value is 2.5, with the same passband ripple and slight reduction in overall bandwidth. I shall try to demonstrate the importance of the higher Q value. For a hard-clipped signal near the center of a band, the third harmonic distortion in that band due to finite filtering is 5 per cent for Vomax, 7 per cent for Stewart's design. The difference is not very significant. For frequencies above the filter peak, the numbers are essentially the same.

Unfortunately, the situation is very much worse for signals below the peak frequency of a particular band. Imagine a signal near the peak of one band, but below the peak of the next higher band by a factor of 1.73 ($\sqrt{3}$). If this signal is distorted (clipped), there will be no filtering as the third harmonic falls above the band peak by the same factor. If there is to be no distortion, then the maximum clipping must be held to whatever response is provided by a tuned circuit which is off-resonance by a factor of 1.73: this is 9.5 dB and 6.8 dB respectively for the Vomax and Stewart designs. Fortunately, this signal component amplitude is down by the same amount when combined with the lower band output, so that some distortion, i.e., clipping, is permissible.

For a total distortion limit of 10 per cent, this number is 2.5 dB, giving a total of 12 dB for *Vomax* and probably about 1 dB for a total of 8 dB for Stewart's circuit. Conversely, for the same amount of clipping, the lower Q design will produce higher distortion, approximately proportional to the square of the Q ratio. Obviously the higher the Q, the lower the distortion; this can be obtained within our design restraints, however, only by a smaller total bandwidth, higher passband ripple, or both. This distortion of signals below the peak frequency of a band increases rapidly with amplitude above a specific level, so that an effective agc of the preamplifier is an absolute necessity. For the perfectionist, I would suggest a design incorporating six or more bands permitting the use of higher Q filters for reduced distortion.

If I have one criticism of Stewart's article, it is his lack of concern for the input amplifier. The low-cost SL1626 IC is a loose-tolerance device with output level variations between samples of more than 8 dB. This necessitates individual calibration of the clipping level control for each assembly. Even worse is the effect of its output distortion, stated as 2 per cent typical with no maximum specified, on the performance of the processor.

Any distortion produced ahead of the processing circuitry will be magnified; i.e., the processor acts as a distortion enhancer. Imagine a signal near the center of the lowest band; if it contains second and third harmonic distortion, these individual components will appear near the centers or peaks of the second and third bands. With 14 dB clipping in the lowest band, the distortion components will be amplified by the same amount, or five times! Therefore, 2 per cent distortion in the input amplifier will show up as 10 per cent added to the processing distortion in the output combiner. While developing the final design for the Vomax unit, I investigated a number of available integrated circuits for use as input amplifiers with agc. I was unable to find one with acceptably low distortion (0.5 per cent), so resorted to the circuit implementation with discreet components.

To summarize, split-band processing, while an acknowledged valuable and effective technique, poses a number of design problems, which require careful attention to detail.

Walter Schreuer Maximilian Associates Swampscott, Massachusetts

I have studied Mr. Schreuer's comments, and with the exception of a couple of minor points, I find myself in agreement with them. Our major difference seems to be our different design goals. Mr. Schreuer's goal was a high quality, high performance, high cost commercial product; my goal was the design of a unit possessing similar characteristics that could be built by the average ham who could not or would not spend nearly \$200 for the Vomax.

As mentioned in the opening paragraph of my article, the price paid for increased performance is increased complexity and cost. Even Mr. Schreuer agrees that to approach perfection the design would require six or more bands. Obviously, we must stop somewhere. Mr. Schreuer and I seem to disagree only on where to stop.

Higher Q filters can offer lower distortion; however, I feel that higher precision components should then be used in the active filters and summing amplifier. While I personally have access to precision parts, I felt the average home builder would have trouble finding them and designed accordingly. Complete design data was given in the text and appendices for those wishing to try design improvements and the experimenter is encouraged to do so.

I must add that Mr. Schreuer took the liberty of comparing the results of using higher Q using my center frequencies. If we compare results using a Q of 2.5 and the center frequencies used in the Vomax, the off-resonance attenuation computes to -8.4 dB, not the -9.5 dB claimed by Mr. Schreuer; a small point to be sure but perhaps worth mentioning.

Regarding the SL1626, the toler-

ance on output level is only 6 dB, not the over8dB claimed by Mr. Schreuer. Except in a production environment, this is not a big problem. Mr. Schreuer raises a good point regarding the distortion characteristics. Perhaps he could share his expert knowledge in this area with those who wish to go to the increased complexity of the discreet circuit.

In summary, I feel the design I presented in the September issue fills a need for an easily reproduced, inexpensive speech processor that delivers more performance enhancement for dollars spent than anything other than perhaps antenna improvements.

> Wes Stewart, N7WS Tucson, Arizona

talking clock



Dear HR:

The diagram of the Talking Clock addition in October, 1979, *ham radio* has an error: the leads connecting to pins 10 and 11 of the 74157 should be interchanged.

The Talking Readout described in the June, 1979, issue was described as it was interfaced to the Spectronics DD-1, which I understand is no longer in production. However, I have successfully interfaced the unit with the Kenwood DG-1, intended for use with the Kenwood 520S. Since the article was published, the multiplexer portion of the circuit has been made available on a separate board, which can be inserted into the DG-1, reducing the interconnection to about nine leads rather than thirty, as multiplexing is done *in situ*. The BCD data is accessed as it enters the six 7454s. Since this is latched or stored data, no gating is required, so pin 4 of the 7454 of the talking digital readout control board may be tied to +5volts.

Similarly, the Yaesu ZD-1 external readout for the FT101-Z should be readily interfaced; BCD data for each digit exist at pins 5, 10, 2, 12, of the six 74LS196s. Gating should be required but is available at Q01, which drives the LED cathodes.

Interfacing to other transceivers may be complicated by the increasing use of seven-segment counter outputs from counters. When I have located a suitable seven-segment to BCD decoder, such as the 74C915, a redesign will be undertaken.

I would be happy to answer any correspondence regarding either article from readers who include an SASE. Inquiries related to other transceivers should include a detailed diagram of the digital readout portion.

Ray Brandt, N9KV Janesville, Wisconsin

Dear HR:

I would like to express my appreciation for your series of articles on "Digital Techniques"; it expanded my knowledge of digital devices considerably. I hope that this series is continued, and I would like to see these articles expanded into the more complex integrated circuits used in microprocessors. Understanding the MCS6522 and MCS6530 would be typical.

I subscribe to *ham radio* not for the straight Amateur Radio articles but for articles on things like the phase-lock loop, power supplies, counters, operational amplifiers, and other technical subjects.

Keep up the good work.

Norbert K. Fox Guttenberg, Iowa



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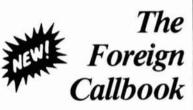
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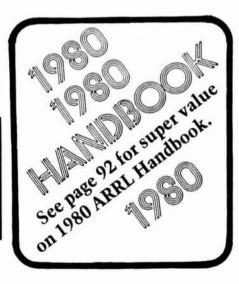
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The Bone Fone surrounds your entire body with a sound almost impossible to imagine.

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But there's no radio in sight and nobody else hears what you do. It's an unbelievable experience that will send chills through your body when you first hear it.

AROUND YOU

And nobody will know you're listening to a stereo. The entire sound system is actually draped around you like a scarf and can be hidden under a jacket or worn over clothes.

The Bone Fone is actually an AM/FM stereo multiplex radio with its speakers located near your ears. When you tune in a stereo station, you get the same stereo separation you'd expect from earphones but without the bulk and inconvenience. And you also get something you won't expect.

INNER EAR BONES

The sound will also resonate through your bones-all the way to the sensitive bones of your inner ear. It's like feeling the vibrations of a powerful stereo system or sitting in the first row listening to a symphony orchestra-it's breathtaking.

Now you can listen to beautiful stereo music everywhere-not just in your living room. Imagine walking your dog to beautiful stereo music or roller skating to a strong disco beat.

You can ride a bicycle or motorcycle, jog and even do headstands-the Bone Fone stays on no matter what the activity. The Bone Fone stereo brings beautiful music and convenience to every indoor and outdoor activity without disturbing those around you and without anything covering your ear.

SKI INVENTION

The Bone Fone was invented by an engineer who liked to ski. Every time he took a long lift ride, he noticed other skiers carrying transistor radios and cassette players and wondered if there was a better way to keep your hands free and listen to stereo music.

So he invented the Bone Fone stereo. When he put it around his neck, he couldn't believe his ears. He was not only hearing the music and stereo separation, but the sound was resonating through his bones giving him the sensation of standing in front of a powerful stereo system.

AWARDED PATENT

The inventor took his invention to a friend who also tried it on. His friend couldn't believe what he heard and at first thought someone was playing a trick on him.

The inventor was awarded a patent for his idea and brought it to JS&A. We took the idea and our engineers produced a very sensitive yet powerful AM/FM multiplex radio called the Bone Fone.

The entire battery-powered system is selfcontained and uses four integrated circuits and two ceramic filters for high station selectivity. The Bone Fone weighs only 15 ounces, so when worn over your shoulders, the weight is not even a factor.

BUILT TO TAKE IT

The Bone Fone was built to take abuse. The large 70 millimeter speakers are protected in flexible water and crush resistant cases. The case that houses the radio itself is made of rugged ABS plastic with a special reinforcement system. We knew that the Bone Fone stereo may take a great deal of abuse so we designed it with the quality needed to withstand the worst treatment.

The Bone Fone stereo is covered with a sleeve made of Lycra Spandex-the same material used to make expensive swim suits, so it's easily washable. You simply remove the sleeve, dip it in soapy water, rinse and let the sleeve dry. It's just that easy. The entire system is also protected against damage from moisture and sweat making it ideal for jogging or bicycling.

The sleeve comes in brilliant Bone Fone blue-a color designed especially for the system. An optional set of four sleeves in orange, red, green and black is also available for \$10. You can design your own sleeve using the pattern supplied free with the optional kit.

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Several people could be in a car, each tuned to his own program or bring the Bone Fone to a ball game for the play by play. Cyclists, joggers, roller skaters, sports fans, golfers, housewives, executives-everybody can find a use for the Bone Fone. It's the perfect gift.

Why not order one on our free trial program and let your entire family try it out? Use it outdoors, while you drive, at ball games or while you golf, jog or walk the dog. But most important-compare the Bone Fone with your expensive home stereo system. Only then will you fully appreciate the major breakthrough this product represents.

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To order your Bone Fone, simply send your check or money order for **\$69.95** plus \$2.50 postage and handling to the address shown below. (Illinois residents add 5% sales tax.) Credit card buyers may call our toll-free number below. Add \$10 if you wish to also receive the accessory pack of four additional sleeves.

We'll send you the entire Bone Fone stereo complete with four AA cell batteries, instructions, and 90-day limited warranty including our prompt service-by-mail address.

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Discover the freedom, enjoyment, and quality of the first major breakthrough in portable entertainment since the transistor radio. Order a Bone Fone stereo at no obligation, today.





fm transceiver, the Azden PCS-2000



Amateur-Wholesale Electronics is proud to announce its superior new Azden PCS-2000 2-meter fm transceiver. The PCS-2000 covers 144-148 MHz in 5-kHz steps (eight hundred channels). It features six memory channels, and scanning of memory, or the full band in "free," "busy," and "vacant" modes. All frequency control functions are performed by a microcomputer.

Upon inspection, the most striking feature is the absence of a large knob for frequency control. In place of a knob, there is a twelve-button microcomputer control keyboard. The desired frequency is programmed into the radio digit-by-digit. Simplex, -600 kHz, or +600 kHz operation is selected by pushing a keyboard button. Using a front-panel rotary switch, three additional offsets become available: +400 kHz, +1 MHz and +1.6 MHz.

The unit comes apart into two pieces; the control head separates from the main unit. With an optional connecting cable, the two units can be located as much as 15 feet apart. This allows great flexibility for mobile and portable operation.

The microphone contains a volume

and squelch control, two frequencycontrol buttons, and a button for instant recall of memory channel 1. By using these controls, the necessity of reaching down to the control panel while driving is greatly minimized.

The PCS-2000 has a huge 1/2-inch LED display that makes frequency determination easy. The S/rf meter is digital, using LEDs instead of the usual often-troublesome mechanical movement. There are two selectable power output levels: 5 watts and 25 watts. Low power is internally adjustable from 3 to 7 watts. Frequency deviation is ± 5 kHz maximum. Azden units limit spurious emissions significantly better than required by FCC regulations.

An external speaker jack is provided on both the control head and the main unit. Optional accessories include external speaker, remote cable, desk microphone, and tonepad microphone kit. Both the desk microphone and the tone-pad kit provide the same remote-control functions as the standard microphone. The Azden PCS-2000 is priced at \$369.00 and carries a 90-day warranty. For additional details, contact Amateur-Wholesale Electronics, 8817 S.W. 129 Terrace, Miami, Florida 33176.

Heathkit transmitter and receiver



Heath Company introduces a new Amateur transmitter kit and matching receiver kit. The new HX-1681 CW transmitter combines solid-state technology with vacuum tube finals to give a transmitter capable of 100 watts minimum output on 80 through 15 meters, and 75 watts out on 10. It features full break-in CW operation, (QSK), built-in VFO, solid-state TR switching, sidetone output with adjustable tone and level, and receiver muting. Keying is provided for the addition of an external power amplifier.

The matching solid-state HR-1680 receiver covers 80 through 10 meters plus the lower 1 MHz of the 10 meter band. It features a preselector-tuned dual-conversion front end for .05 μ V sensitivity, as well as solid-state diode bandswitching, built-in 100 kHz calibrator and switchable wide/narrow active audio circuitry for SSB or CW operation. The transmitter kit sells for \$239.95 (requires separate power supply) and the receiver is priced at \$209.95. For more information write Heath Company, Department 350-940, Benton Harbor, Michigan 49022.

new MFJ 24-hour digital clock



The MFJ-101 is a new 24-hour, solid-state, digital clock. It features pleasant blue, 0.6-inch digits that are easy on the eyes, yet bright enough to see all the way across the room.

The MFJ-101 has an ID timer that alerts you every nine minutes after you tap the ID/doze button. This nine-minute timer gives you a full minute to identify after the timer sounds and still be legal.

The alarm feature will remind you of that important sked, or wake you in the morning with a pleasant but persistent chirping sound. The fast/slow set buttons make setting time and alarm simple, while the lock function prevents accidental missetting.

The alarm has an indicator which lights up when the alarm is on. If power has been interrupted, the digits flash on and off until the time is reset. The MFJ-101 is UL approved, and operates on 110 Vac, and can be switched for 50/60 Hz for operation in European and Asian countries.

MFJ provides a 30-day, moneyback trial period. If you are not satisfied, you may return it within 30 days for a full refund (less shipping). MFJ also provides a one-year limited warranty.

The MFJ-101 is available from MFJ Enterprises, Inc., for \$29.95 plus \$3.00 shipping and handling.

To "order, call toll-free 800-647-1800, or mail order with check or money order to MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762.

mobile antennas



A completely new line of mobile antennas has been introduced by Signals Communications Corporation. The new line covers 30 through 512 MHz. These high-performance, lowprofile antennas feature plated whips, high-impact bases, and optional springs. These antennas will fill the needs of most professional and government applications. The Matchmate mobile-mount adapter system permits the mounting of Signals antennas on all popular mounts.

The mobile antennas feature solid dielectric coils enclosed in highimpact base covers. This configuration eliminates variation from coil to coil, thereby ensuring consistent high quality. The antennas are rated at 200 watts with less than 1.5:1 VSWR, and 50 ohms nominal impedance.

The antenna whips are long enough to permit cutting to frequency. They are made of quadraplated 17-7PH stainless steel. This results in lower skin effect losses and greater radiation.

Contact Signals Communications Corp., 1 Signals Park, P.O. Box 4833, Manchester, New Hampshire 03108.

rf speech processors for Drake

The new Sherwood RF speech processors provide no-compromise rf/i-f envelope clipping for high intelligibility and unexcelled talk power. A specially designed eight-pole i-f crystal filter drives a highly effective active hard limiter (clipper). Great care has been exercised in developing reliable, ultra-fast, high-isolation pindiode switching of filters and the clipper stage. Full eight-pole, low-leakage receive and transmit capability is realized. The processors are rig powered, and are easy to interconnect with your equipment. No drilling of holes is required.

The 4-SP for the 4-Line transmitters (T-4X, T-4XB, and T-4XC) has separate LSB and USB eight-pole filters of carefully chosen bandwidth for proper filtering ahead of its clipper. The highly versatile 7-SP for the TR-7 not only improves your transmitted signal, but offers automatic transmit/receive switching as well as the option of selecting the processor's special crystal filter for 16-pole, 1.9- kHz receive bandwidth. In addition, circuitry is provided for installing one of the normal accessory filters, such as the SL-500, on the processor board, allowing room for other filters in the TR-7 itself.

All units have filter selection and clipping in/out switches, plus an output control after the clipper (for proper final/PA drive adjustment from band to band, or for precisely setting linear-amplifier excitation level). Clipping is adjusted by the rig microphone gain control. Each processor is housed in an attractive charcoalcolored cabinet, accented with white lettering, that blends well with all Drake equipment. High-quality machine-screw construction is used (no sheet-metal screws to strip out), with top and bottom removable for easy access to both sides of the PC board.

Model 7-SP for the TR-7 sells for \$265.00; the 4-SP for the 4-Line transmitters sells for \$285.00. Others will be available later. Money back if not satisfied. Master Charge and VISA welcome. Please add \$3.00 shipping/handling per order. Overseas airmail add \$6.00. Write to Sherwood Engineering, 1268 South Ogden Street, Denver, Colorado 80210.

brochure from J.W. Miller

A new four-page brochure describing accessory equipment for Amateur Radio operators is now available from J.W. Miller, Division Bell Industries. Included in the brochure are descriptions and technical details of SWR and power meters, rf speech processors, precision coaxial switches, and various interference filters.

Direct reading SWR, forward power, and reflected power are provided by models CN-720 and CN-620 over the 1.8-150 MHz range; model CN-630 covers 140-450 MHz.

Rf clipping that insures low distortion is provided by models RF-440 and RF-660 speech processors.

Adjacent channel isolation of better than 50 dB at 300 MHz and 45 dB at 450 MHz is provided by two-position model CS-201 and four-position model CS-401 coaxial switches. The broadline of interference filters includes high-pass, lowpass, audio, and ac power line filters.

For additional information, contact Jerry Hall, Operations Manager, J.W. Miller Division, Bell Industries, 19070 Reyes Avenue, Compton, California 90221.

multifunction frequency counter

The FC-841 is a seven-digit multifunction frequency counter that covers the 10-Hz to 50-MHz range. The tilt-view stand, all-steel case, and the 0.3-inch high LED readout facilitate troubleshooting. Supplied with four AA cells, the counter can also be plugged into an ac outlet or a car's cigarette lighter.

The FC-841 has a selectable gate time of 100 milliseconds or 1 second and a sensitivity of 30 mV rms, up to 30 MHz, decreasing to 60 mV rms at 60 MHz. The time base stability of the counter is 3 parts per million from 20° to 30°C (68° to 86°F). The instrument also features a switch for selecting the kHz or MHz ranges. The FC-841 is supplied complete with batteries, antenna, and a test lead having alligator clips on one end and a BNC connector on the other end.

Price is \$90 from Soar Electronics Corp., 200 13th Avenue, Ronkonkoma, New York 11779.

Hustler ten-meter Yagi

A new beam, designated 10-MB-4, is the result of extensive antenna design refinements by Hustler. The beam is a four-element Yagi optimized for best directivity, excellent front-to-back ratio, and maximum gain through selective element spacing, and precisely resonated element lengths. The 10-MB-4 has a gammamatch feed system and is adjustable for a 1.2:1 or better SWR.

The Hustler 10-MB-4 is designed to withstand severe weather, yet it's light enough to be turned by a TV antenna rotator. The entire antenna is built from high-strength aluminum tubing and can easily be grounded for lightning protection.

The new Hustler 10-MB-4 10meter beam has a suggested list price of \$109.95 and is available now. For further information on this or other Hustler Amateur antennas, write Hustler Inc., 3275 North B Avenue, Kissimmee, Florida 32741.

Microcraft Morse-A-Word code reader

An eight-character Morse-code reader has been introduced by Microcraft, for SWLs, beginners, and veteran Amateur Radio operators. It accepts audio signals from a communications receiver's headphone, jack, or loudspeaker, and displays the decoded characters. All text characters letters, numerals, punctuation marks, special Morse-symbols, and word spaces - are shown sequentially on the display in moving-character fashion. Code speeds of 5 to 35 WPM can be copied, depending on the setting of the front panel control. The Morse-A-Word also includes a built-in code-practice oscillator and monitor speaker for practice sessions. A complete kit is \$169.95, and a wired and tested version is \$249.95. Write to Microcraft, P.O. Box 513, Thiensville, Wisconsin 53092.

computer logging package

Snow Micro Systems, Inc., announces a logging package for Radio Amateurs. This package, written in Northstar BASIC, allows Amateurs to keep their log records on floppy discs.

The log data-file can be examined, edited, and printouts and summaries obtained of sections of the log. The entries can be listed by prefix such as G, G3, G3Z, G3ZC, or G3ZCZ. QSL card information can be automatically printed out by the computer.

Different logs can be emended, merged, deleted, renamed, searched, and summaries listed on any of the seven output devices supported by the Northstar system.

The package contains one disc, which includes sample logs and full instructions, and is available for \$15.00, postage paid.

For further information, write to Snow Micro Systems, Inc., P.O. Box 1704, Silver Spring, Maryland; or telephone (301) 622-1931.



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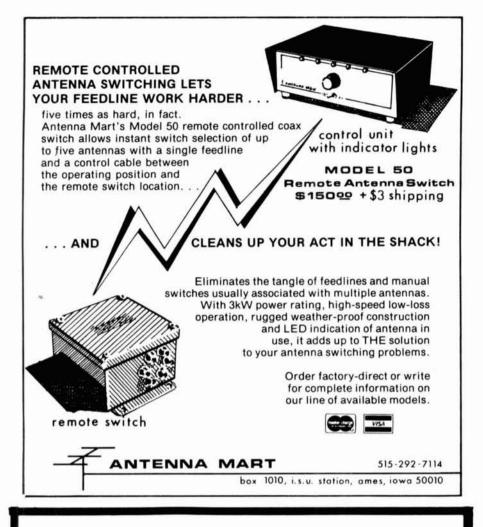
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Coming Events

MICHIGAN: Traverse City Swap 'n Shop, Saturday, February 9, 10 AM - 4 PM. Talk-in on 25/85 or 52. For more details contact: Jim Fenton, W8LVZ, 11476 Cedar Run Rd., Traverse City, MI 49684. Phone: (616) 275-7597. Good Door Prizes

MISSOURI: The Missouri Valley Amateur Radio Club, Inc. proudly announces its second annual Pony Express Days from the original stables in St. Joseph, April 5 and 6. Operating time 1000 CST to 1900 CST both days. Anyone making contact with the club station will receive the Pony Express Award. Send legal-size SASE along with personal QSL card to: Missouri Valley ARC, 401 North 12th Street, St. Joseph, MO 64501. Certificate will be stamped with original seal of the Pony Express. Operat-ing frequencies will be 28.575 and 10 kc's from the bottom of the General phone band on the other bands - 15 through 75. CW bands will be 28.150, 21.150 and 7.125. Listen for WONH from the home of the Pony Express.

VERO REACH FLORIDA HAMFEST - The Treasure Coast Hamfest will be held on March 15 & 16, 1980 at the Vero Beach Community Center, Admission is \$3 per family in advance, \$3.50 at the door. Talk in on 146.13/73 - 146.04/64 - 222.34/223.94. For information write P.O. Box 3088, Vero Beach, Fla. 32960

MARYLAND: Calvert Hall and the Baltimore Amateur Radio Club's all new 1980 Greater Baltimore Hamboree and Computerfest, Sunday, March 30. New location: The Maryland State Fairgrounds at Timonium located just off I-83, 2 miles north of I-695 north of Baltimore. Plenty of space for dealers, displays, tables, commercial exhibits and lectures for both Amateur Radio and personal computer enthusiasts. Acres and acres of space for tailgate sales. Door prizes throughout the day with hundreds of dollars in cash prizes given away at the grand drawing. Admission: \$3.00. Tables \$5.00. Special events, lectures, and demonstrations. Refreshments available. Talk in on the BARC repeaters, 146.07/67 and 146.34/94. For additional information, tickets and space reservatiosn contact: Joseph A. Lochte, Jr., 2136 Pine Valley Drive, Timonium, MD 21093. For a recorded message dial: (301) HAM-TALK.



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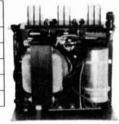
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Inside View - RS-12A

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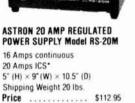
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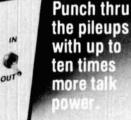
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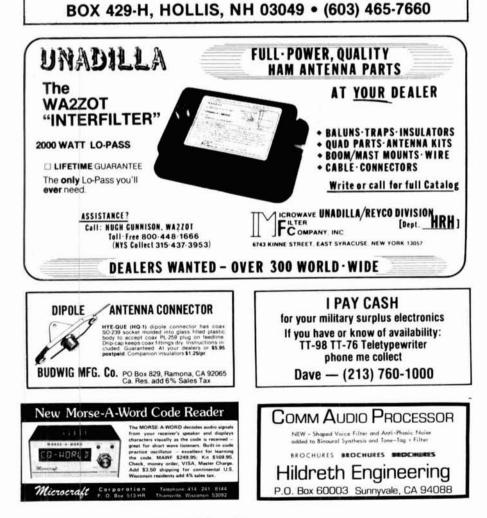
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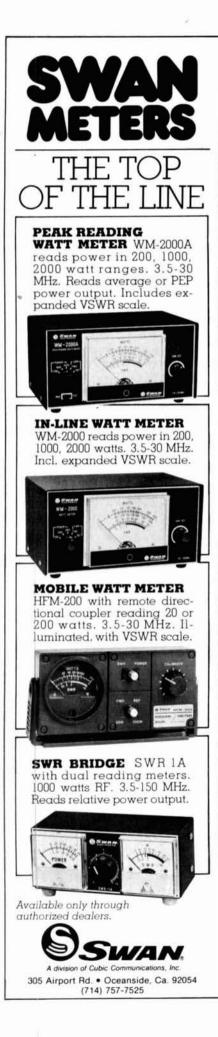
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More Details? CHECK - OFF Page 94



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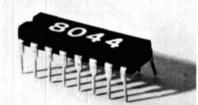
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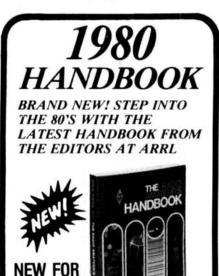
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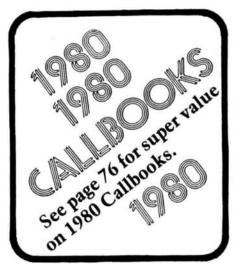
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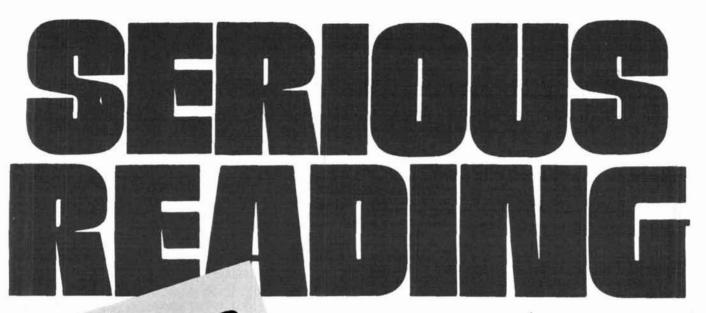
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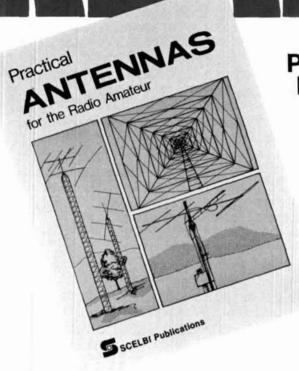
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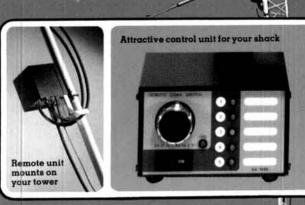
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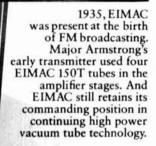
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