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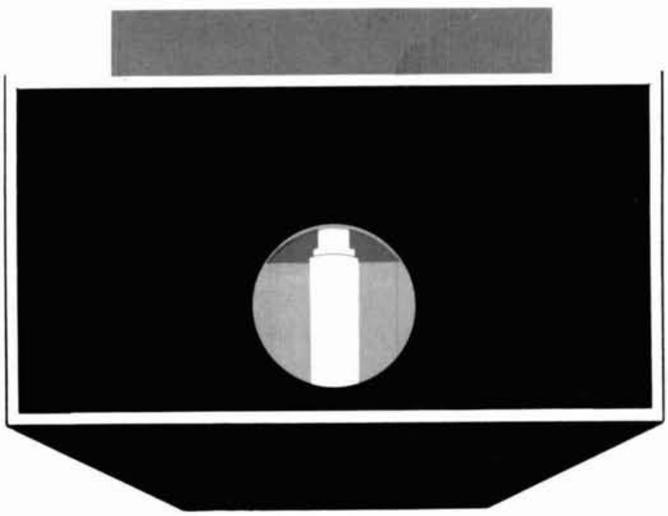
ham radio magazine

SEPTEMBER 1980

GUNN OSCILLATOR DESIGN FOR THE 10-GHZ BAND



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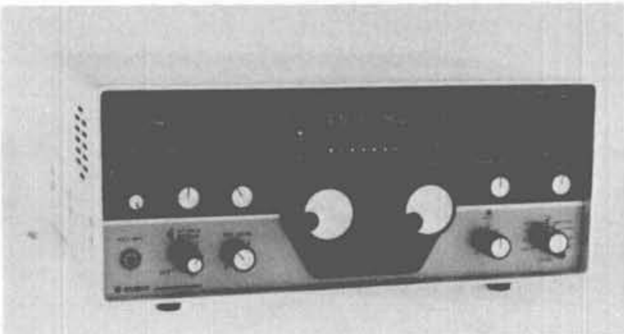

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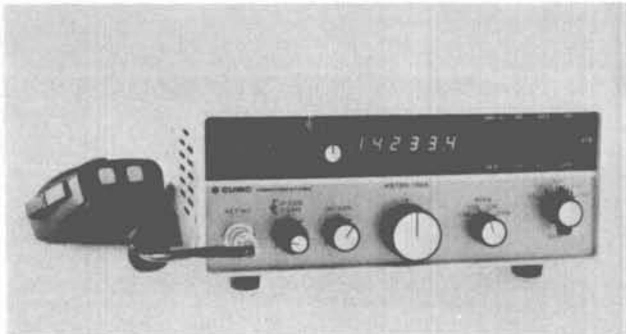

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ham radio magazine

SEPTEMBER 1980

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contents

Observations & Comments

While browsing through RSGB's *Radio Communications* (August, 1979, page 751) I came across an interesting letter from H. Herzer, DL7DO,* which had been subsequently picked up by *QST* in their July, 1980, issue (page 73). The subject of Herr Herzer's letter was the antiquated RST signal-reporting system and what might be done about it.

Herr Herzer presents a pretty good case for eliminating the RST system in terms of today's standards of equipment sophistication, band conditions, and operating methods.

During the early years of Amateur Radio, as Herzer points out, the RST reporting system was a valuable aid to operators, as most (if not all) equipment was homemade. Amplitude levels were low, power-supply filter systems were primitive or nonexistent, and measurement equipment was crude. So the RST system was a means of evaluating on-the-air signals. If your rig used, say, a type 45 tube with 90 volts on the plate, a report of S7 meant that the signal was "fairly strong." A tone report of T8 meant that something had to be done in the power-supply filter department to bring the signal to T9: "pure dc note." If you received a readability report of R3, this could mean just about anything, from poor propagation conditions to faulty adjustment of receiving equipment at the other end of the circuit.

Herzer recommends a "Q System" to replace the antiquated RST reporting system for both CW and radiotelephone communications. In Herzer's system "Q" stands for *transmission quality*, which accounts for the one and only parameter relevant to successful information transfer by Amateur Radio stations. It all boils down to: "Has the message been received and understood?" The Q reporting system used only three variables:

Q1 At no time has there been sufficient transmission quality; that is, no copy has been received.

Q2 Sufficient transmission quality has occurred some of the time; that is, partial copy has been received.

Q3 Sufficient transmission quality has occurred at all times; that is, full copy has been received.

Intermediate stages of reporting, such as Q1/2 or Q2/3, might be used. You can always ask the other operator for an explanation.

Herzer feels that such a system will result in a considerable reduction in QRM, contest reporting, and logging. I agree. Today's RST reporting system is not only redundant but meaningless.

Of course, this means changing your QSL cards to show the new system, and it will probably take a long time to implement on a worldwide basis. But the system has real merit in reducing "on-the-air" pollution in today's Amateur bands.

The other night I worked several stations, foreign and domestic, on the low end of the 14-MHz CW band. All reports were RST 579. During contest operation, all stations reported RST 599. Utterly meaningless! Why bother with such a signal-reporting system? In fact, why is any signal-reporting system useful today?

What do you think? We'd like your opinion.

Alf Wilson, W6NIF
technical editor

**Radio Communications*, Journal of the Radio Society of Great Britain, August, 1979, page 751.

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

Frequency Coverage: 50~54MHz **(143.8~148.19MHz)**

RF Output Power:
SSB 10W PEP
(1~10W adjustable) **(10W)**
CW 10W
(1~10W adjustable) **(10W)**
AM 4W
(0~4W 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
1μV
FM* 0.4μV **(0.4μV)**

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)
FM* More than ±7.5KHz at -6dB
Less than ±15KHz at -60dB

Dimensions: 111mm (H) × 241mm (W) × 311mm (D)

Weight: 6.1kg **(5kg)**

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comments

microphones

Dear HR:

Readers should be cautioned to avoid using the type of audio cables suggested by W1OLP in his article on microphones and speech processing in the March issue of *ham radio*. Only the audio line should be inside the shield. Having the audio line plus the PTT line and/or battery line inside the shield will, in many cases, result in unwanted noises in the transmitted audio: hum, switching noises, and stray rf. It is preferable to use microphone cable which has one shielded line for audio, with the PTT and other control lines outside the shield.

Buddy Massa, W5VSR
New Orleans, Louisiana

Hallicrafters story

Dear HR:

The fine story by W6SAI about Bill Halligan's HT-4 (BC-610) in the November, 1979, issue of *ham radio* surely brought back a flood of memories — Utah Beach, Ste. Mere Eglise, Carentan, Isigny.

We had five mobile units in ADSEC (Advanced Section Communications Zone), 3rd Army, and only one of them was an SCR-299, the others were SCR-399s. I would like to call your attention to the incorrect caption with the photograph on page 24. The mobile unit is an SCR-299, not 399; the 399 differed mainly from the 299 in that it was not housed in a panel truck; it was provided with an HO-17 plywood shelter, designed to fit the equally famous 6x6 International truck. The shelter could be lifted off, complete with its equipment, and operated on the ground as a fixed station.

Our first team's unit was installed in a "Duck" amphibious version of the 6x6, with two little PE-75 gas generators connected in parallel on the aft deck. They were let down from the LST *before* H-hour on D-day and made an attempt to scramble ashore, but they were met by severe mortar fire and forced to withdraw. Later they gained an exposed position on the beach and made contact with our station in England that had been set up in late May.

I was the Platoon Sergeant of the fifth team and some days later we had all units dispersed several miles apart and well camouflaged in the Normandy apple orchards. They were tied with field wire keying lines (duplexed for telephone) to a radio center in the loft of an old French barn. In the stable below were the TC-10 telephone boards and the Message Center. All operation was manual and long press dispatches were cleared between items of military traffic. Later three more SCR-399s arrived and eight were operated for a few days. Then one-by-one the ADSEC units began to leave to follow the action. Some months later one was in Namur, Belgium, in contact with besieged Bastogne.

I should add that the operators who accompanied these units were highly specialized, having worked in the signal center at the Pentagon before leaving the States. Likewise, many of the technicians were specially trained on SSB multi-channel (AFSK) high-power transmitters (40 kW). They formed the nucleus of the Paris communications center in the "Block House" about a block from the Arc de Triomphe de L'Etoile on Rue Wagram. Some followed the action; some had very important missions elsewhere.

We experienced only one real trouble with our BC-610Es. The high-voltage in the modulation transformer would break down to ground, killing the rig. We found that we could set the transformer up on four short standoff insulators supplied in the spares chest and be back on the air in half an hour. Information was sent up through channels on this fix and apparently others had experienced a similar failure because a Field Change Bulletin was put out by the Signal Corps directing that this modification be installed in all BC-610Es.

After VE-day, returning to France and to the Signal Depot at Mohn near Meziers, SCR-399s were stashed in the fields around the buildings as if we were operating a trailer park. I wonder now what happened to all of them. Some were trans-shipped to the Pacific Theater, but most were left behind.

Clifford O. Field, WA2JVD
Fair Haven, New York

more Hellschreiber

Dear HR:

E.H. Conklin, K6KA, is not quite right in describing the Hellschreiber as a wideband system (Comment, March, 1980). Admittedly, the bandwidth of any keyed system is a function of the keyed element rise time, but with proper pulse shaping as practiced by the majority of the PAØ and German Amateurs the amount of spectrum space occupied by a Hellschreiber signal is only marginally greater than that of 45.5-baud RTTY.

The bandwidth necessary for Hellschreiber may be quite easily computed by reference to CCIR Recommendations, which in Appendix 5 of Radio Regulations state that this is

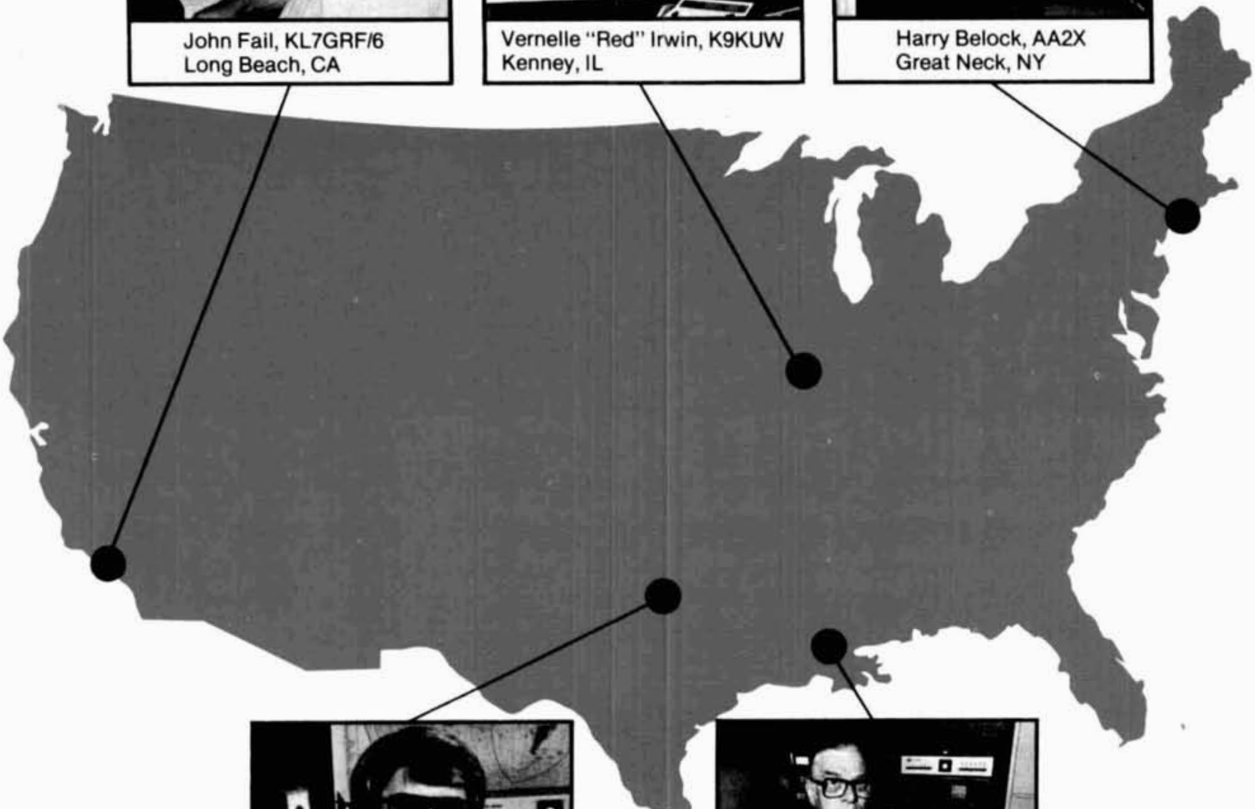
(Continued on Page 66)

WHEN OUR CUSTOMERS TALK... WE LISTEN.

John Fail, KL7GRF/6
Long Beach, CA

Vernelle "Red" Irwin, K9KUW
Kenney, IL

Harry Belock, AA2X
Great Neck, NY



Tom Gentry, K5VOU
Dallas, TX

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More Details? CHECK—OFF Page 94

presstop

FREEDOM TO LISTEN TO RADIO communications would be drastically curtailed by a bill introduced in the House of Representatives. HR-7747, by Rep. Richard Preyer (D-North Carolina) and several co-sponsors, would "prohibit the unauthorized reception of private radio communications." By prohibiting reception this new bill goes much further than Section 605 of the Communications Act, which is generally interpreted to ban the divulging of the content of private radio communications.

Pressure To Ban Reception of various radio services has been around for a long time, most recently in connection with radar speed monitor detectors. Now pressure is coming from satellite TV distributors, worried that do-it-yourselfers will cut into their revenues.

HR-7747 Has Been Referred to the House Committee on Interstate and Foreign Commerce and Judiciary for further action.

FCC'S AMATEUR EXAM PASS RATE has been climbing steadily in recent months, after having remained fairly constant for years, an FCC analysis of Field Office reports shows. Recently available study materials containing exact exam questions and answers is the only apparent reason for the shift, which is the cause of concern in both FCC and Amateur circles. There is considerable feeling that the increased pass rates are due mainly to memorization rather than understanding, which would result in unqualified applicants receiving Amateur licenses and the license itself being cheapened as a result.

The New Exams That Came into use in June should at least temporarily curb the problem. If, however, the West Coast publisher who's been selling the question-and-answer sheets gets the new exams, a considerable investment of time and taxpayers' money will have gone for nought. The FCC's rules apparently have no provisions for protecting the security of its examinations.

PETE HURD, K4NSS/N1SS, HAS JOINED HAM RADIO as assistant publisher. Pete, who's long been involved with many activities that concern Amateur Radio's future, arrived in Greenville on July 14 to begin his new role on a commuting basis. As assistant publisher he'll be responsible for day-to-day management of the Ham Radio Publishing Group and its three publications.

Pete's Part In The ACAR operation, where he served as the executive secretary for the Advisory Committee for Amateur Radio that so successfully formulated the strong pro-Amateur Radio U.S. WARC position, has made him well known to high-ranking Amateurs both domestically and abroad. Professionally, Pete served with the Air Force until his retirement a year ago, where he was a Colonel with the Department of Defense working in research and engineering management. There he not only worked closely with the U.S. electronics industry, but was often in Europe on NATO business where he became well known to many overseas Amateurs.

AN AMATEUR EQUIPMENT RIPOFF in the Washington, D.C., area has also victimized some dealers in other areas as well. Using the expired call W5SRZ as part of his "identification" and sometimes on the air as well, an individual from northern Virginia has been able to purchase quite a bit of equipment without paying for it.

He Orders By Telephone, using a valid credit card number not his own. Since Master Charge's credit OK is only for the number and not who or where the card user actually is, the sale is OKed and the merchandise delivered without hesitation when the buyer comes in to pick it up.

Distributors In California, Florida, and New York, as well as several in the greater Washington area are known to have fallen victim to "W5SRZ," who also sometimes says he's an airline captain. Although some criminal charges have been filed against him, he's apparently still active.

Merchants Who Accept Credit card orders by phone or mail must be especially cautious, as the law makes them the losers, with little recourse, when an incorrect account is billed. One dealer first learned he'd been taken when a woman phoned to ask why he'd put $1500 in radio equipment on her charge card!

U.S. AMATEUR POPULATION SWELLED to 385,625 licensees at the end of June, an increase of 1,837 from the start of the month. FCC statistics went on to show that Amateur Radio experienced twice as much growth (12,583 new licenses issued) in the first six months of this year as it did in all of 1979 (6,119 new licenses). June's increase, in fact, nearly equaled the 2,401 licenses issued between April and December of last year!

INDECENT LANGUAGE COST WD8NLS his station license and triggered suspension of his General Class operators license for the remainder of its term. FCC engineers had monitored his use of indecent language on the air on three different occasions, plus his rebroadcasting music on another.

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Super Design. *All Solid-State and Broadbanded*—from the pioneer, Ten-Tec. *Modular* plug-in circuit boards. *Functional Styling* with convenient controls, full shielding, easy-to-use size (5¾"h x 14¼"w x 14"d).

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Full Break-In. HERCULES puts the conversation back into high power CW operation—you can hear between every character you send.

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Super Styling. Designed to match OMNI, the HERCULES has the same height as OMNI, plus matching bail and matching colors. The front panel is simplicity in itself with two push-button switches (power and mode) plus two knobs (meter and bandswitch), and a "black-out" monitor panel (when unit is off, meters are unobtrusive). Amplifier size is 5¾"h x 16"w x 15½"d.

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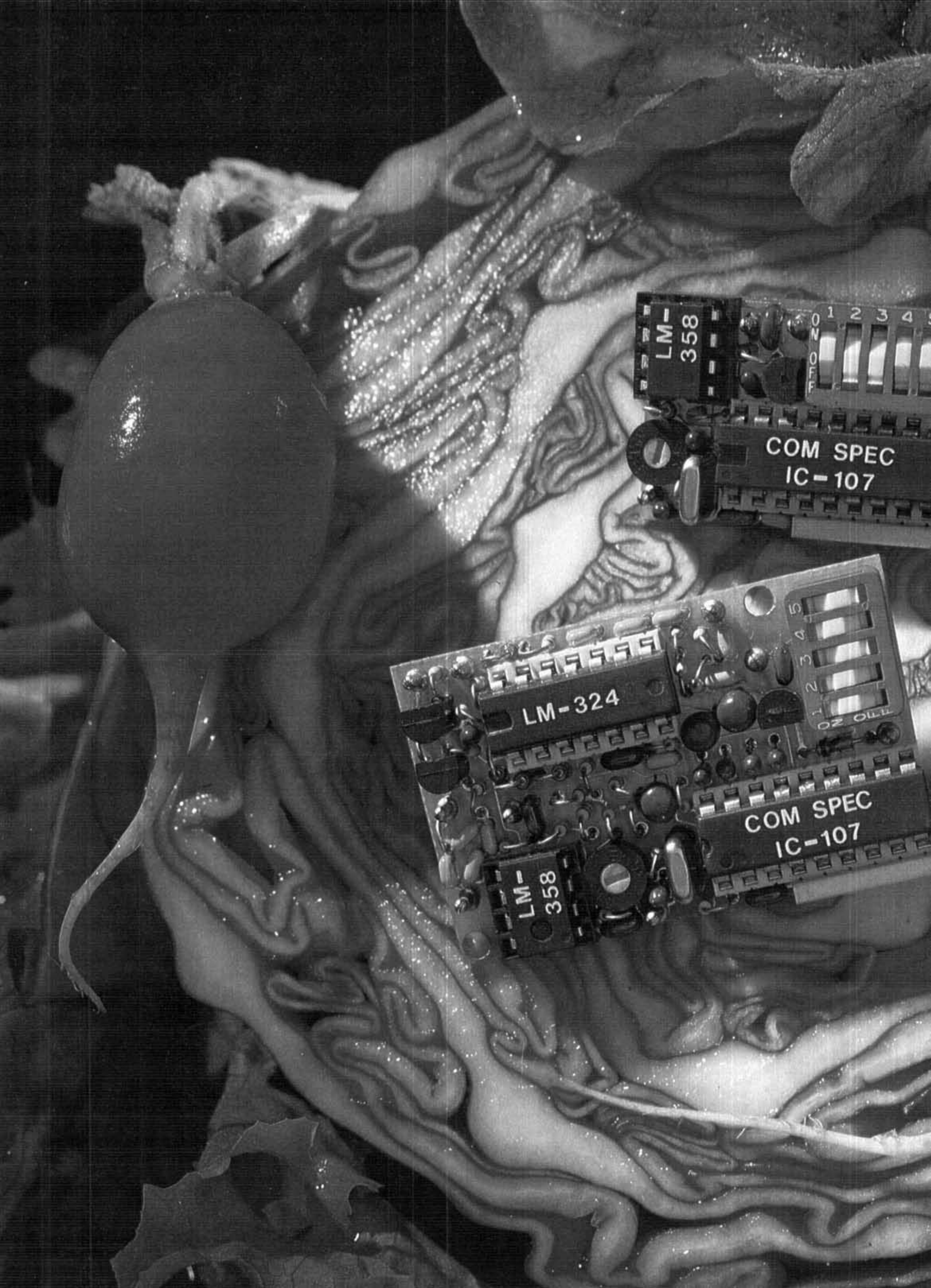
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LM-358

Gunn oscillator design

for the 10-GHz band

Gunn oscillators are an attractive alternative to reflex klystrons as a signal source in the Amateur microwave bands. However, to some, the design of stable Gunn oscillators is considered to be somewhat empirical. In practice a particular design approach is tried then perfected through cut-and-try. Much research has been done with Gunn devices as described in the literature,[1,2] leading to high-quality commercial units.

In an effort to dispel some of the mystery surrounding Gunn oscillator design, I'll set down some ground rules for a certain design approach that produced good results. I wish to emphasize at this point that what follows is only *one* of several approaches that will give results. The oscillators were designed for 10 GHz, but this design could be used and modified for any microwave band between 5 and 90 GHz. Gunn devices can be made to oscillate in a number of configurations. These can be in the form of coaxial resonators, waveguide cavities, and microstrip circuitry.[3] They can be tuned mechanically or electrically. The theory of how Gunn devices oscillate is covered in other literature.[4,5,6]

The type of Gunn oscillator presented in this article is a waveguide cavity oscillator. If you wish to tune a waveguide cavity oscillator over a wide range, the cavity volume formed between the diode mount and a movable back wall, in the form of an rf choke, is changed. D. Evans,[7] and Tsai, Rosenbaum, and Mac Kenzie[8] describe wideband mechanically tunable waveguide cavity oscillators. However, the other approach to waveguide cavity oscillator design is to use an iris-coupled waveguide cavity. Here the resonant cavity is formed by the iris and the diode mount. The backwall is adjusted close to the diode mount for optimum operation. The design is inherently narrowband, tunable over 100-300 MHz. Since operation is permitted between 10.0 and 10.5 GHz, this type of oscillator presents a practical approach. Other features of the oscillator are that it is fairly straightforward in design and can be easily reproduced.

Following is a presentation of some of the ground rules for iris-coupled waveguide oscillator design. From these, a design procedure is presented followed by testing techniques and precautions to be taken when putting the oscillator into operation.

The iris-coupled waveguide cavity oscillator is shown in **fig. 1**. The diode is mounted across the center of the broad dimension of the waveguide. Dc bias is coupled to the diode through an rf choke. The backwall, which can be moved, is usually close to the diode mount and can be adjusted for stable operation. The function of the iris is to complete the waveguide cavity while providing coupling between oscillator and load. The tuning screw located between the diode post and the iris provides a limited tuning mechanism for the oscillator.

Since the Gunn oscillator is a negative-resistance device, it will have a tendency to oscillate at more than one frequency. The idea is to reduce the number of spurious resonances and make it oscillate where you want it to. The first step in this direction is to examine what makes up the main oscillating cavity of the oscillator.

Waveguide cavity. The fundamental resonant mode of the cavity occurs when the distance between the iris and the effective backwall is one-half the guide wavelength. If the diode is mounted near a sidewall of the cavity, the effective backwall is somewhere between the diode post and the backwall. If, however, the diode is mounted in the cavity center, the effective backwall is in the plane of the diode mount.[9] This is verified by slotted-line measurements looking at the normalized impedance in the plane of the diode mount.

One end of the cavity is now defined. The other end is defined by the iris. The iris reactance can be either inductive or capacitive. **Fig. 2** shows an equivalent circuit of an iris-coupled cavity. If the iris is capacitive, the physical length of the cavity is longer than $\frac{\lambda_g}{2}$ as shown in **fig. 2(a)**. If the iris is inductive as in **fig. 2(b)**, the actual cavity length is shorter than $\frac{\lambda_g}{2}$. The relationship between cavity length and iris susceptance is:[10]

By Richard Bitzer, WB2ZKW, 3 Ray Street, Milltown, New Jersey 08850

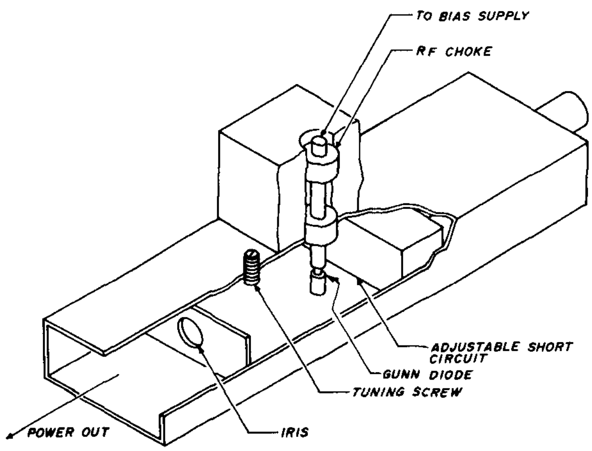


fig. 1. Simple waveguide iris-coupled oscillator. With the diode mounted across the guide center, the distance between iris and diode post is approximately one-half guide wavelength. Iris size is optimized for maximum power output while providing isolation from load mismatches. The rf choke minimizes power loss through the bias line. Backwall is adjusted to provide a stable operating point.

$$\ell = \frac{\lambda_g}{2\pi}\left[n\pi + \tfrac{1}{2}\tan^{-1}\frac{2}{|b_e|}\right] \text{(cap.)} \quad (1)$$

$$\ell = \frac{\lambda_g}{2\pi}\left[n\pi - \tfrac{1}{2}\tan^{-1}\frac{2}{|b_e|}\right] \text{(ind.)} \quad (2)$$

where ℓ = cavity length

λ_g = guide wavelength

$= \dfrac{\lambda}{\sqrt{1-\left(\dfrac{\lambda}{2a}\right)^2}}$

n = integral number

$|b_e|$ = absolute value of normalized iris susceptance

λ = operating wavelength

a = wide dimension of waveguide

In most cases $n = 1$ to reduce the possibility of the oscillator operating at a lower frequency. Once the value for b_e is found, the appropriate expressions (**eqs. 1** or **2**) give a value for the cavity length, ℓ .

The estimated value of loaded Q for an iris-coupled cavity is given by:[11]

$$Q_L \simeq b_e^2 \frac{\pi}{2} \frac{1}{\left(1-\dfrac{f_c}{f}\right)^2} \quad (3)$$

where b_e = normalized value of iris susceptance

f_c = guide cutoff frequency, TE_{10} mode

f = operating frequency

From the previous discussion, the cavity length, ℓ , and the cavity loaded Q depend on the value given to the normalized iris susceptance.

Iris. An iris is an obstruction placed into a waveguide system that electrically has a value of reactance or its inverse, susceptance. Irises can take various shapes (**fig. 3**) and can be inductive, capacitive, or resonant. From the standpoint of ease of construction, a centered circular aperture is the simplest form to make. The circular iris is inductive and its value can be calculated; however, it's easier to use the graph[12] in **fig. 4**.

The graph gives a good approximation for the value of normalized susceptance, $\frac{B}{Y_0}$ (that is, b_e) as a function of the ratio of iris diameter to the broad dimension of the waveguide. This is done for various waveguide aspect ratios and various ratios of operating-wavelength-to-guide width. The values of b_e from the chart are in good agreement with measured values. Now that the iris susceptance is calculated using **fig. 4**, cavity length can be found.

The question is raised as to how large a hole should be made in an iris plate. Critical coupling, the point where power output is maximum, occurs when the iris area is about 25 per cent of the waveguide cross-sectional area.[13,14] If the hole is further enlarged, the oscillator cavity will be overcoupled with a resultant drop in output power and poor stability. On the other hand, by making the hole smaller, the cavity will become undercoupled resulting in a drop in oscillator output power.

So in determining cavity length, choose an iris area between 20 and 25 per cent of the waveguide cross-sectional area. Since the circular iris area is $\pi d^2/4$, the diameter can be found. The next step is to design the cavity to operate at the highest frequency of interest. The cavity can always be tuned downward by a dielectric screw tuner. The operating frequency determines the wavelength. Knowing the a and b dimensions of the guide, the value of the normalized susceptance, b_e, can be found from **fig. 4**. This value is then substituted into **eq. 2** to find cavity length ℓ.

Rf bias-choke system. Erratic operation, spurious responses, and power loss can be caused by an improper bias-choke design. The diode must be operated with a dc bias while decoupled from the

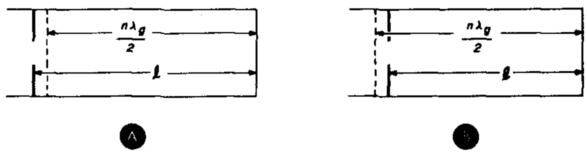


fig. 2. Equivalent circuit of an iris-coupled cavity. When the iris is capacitive, A, the actual cavity length is longer than multiples of one-half guide wavelength; when inductive, B, the actual cavity legnth is shorter than multiples of one-half guide wavelength.

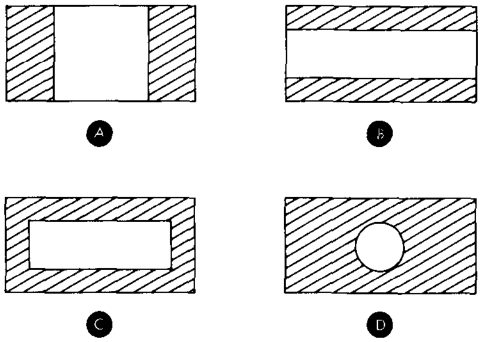


fig. 3. Various iris configurations. A and D are inductive; B and C are capacitive and resonant respectively. The circular iris in D is easiest to make.

bias supply. Early designs operated erratically and even had diode failure because of poor bias-choke design. Two common types of chokes are the radial-line choke and "dumbbell" choke. The radial-line choke requires a large circular plate parallel to the broad side of the waveguide with small separation. The radius of the radial line is approximately $\lambda/4$, with a configuration of an open-ended quarter-wave transmission line. The feedpoint impedance at the diode end becomes very low.

Although this type of bias-choke system is easy to

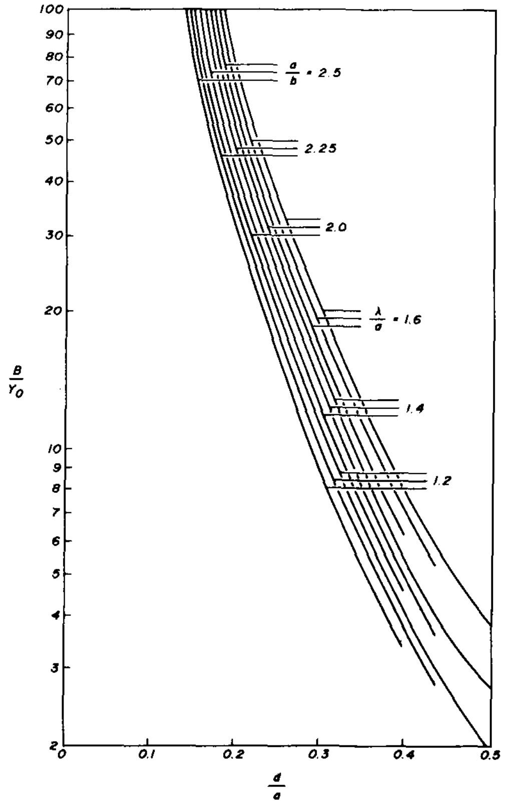


fig. 4. Relative susceptance of a centered circular aperture as a function of the ratio of iris diameter to waveguide broad dimension.

build, some radiation occurs from the open end. A more popular choke arrangement is the "dumbbell" choke, **fig. 5**. As the name implies, sections A, B, and C resemble a dumbbell. This choke design is basically a series of quarter-wave-long coaxial line transformers, alternating between low and high Z_0 sections. The design transforms a wide variety of impedances at the feed point to an extremely low impedance at the Gunn-diode mounting point. This is necessary since the waveguide wall is a current-carrying surface.

The characteristic impedance of a coaxial line is:

$$Z_0 \simeq \frac{60}{\sqrt{\epsilon_r}} \ \ln \frac{r_2}{r_1} \qquad \textbf{(4)}$$

where r_2 = inside radius, outer conductor
r_1 = outside radius, inner conductor
ϵ_r = relative dielectric constant

Since sections A and C must have close spacing with respect to the wall, a dielectric material such as Mylar tape can be used to prevent the choke from shorting the bias supply. Sections A and C will be shorter in length than section B to account for the difference in phase velocity caused by the different dielectric constant of the tape. The velocity that the wave propagates is given as:

$$v = \frac{c}{\sqrt{\epsilon_r}} \qquad \textbf{(5)}$$

where v = velocity of propagation
c = speed of light = $3(10^8)$ meters/second
ϵ_r = relative dielectric constant

To calculate the $\lambda/4$ sections, operating wavelength, λ, is found from:

$$\lambda = \frac{v}{f} \qquad \textbf{(6)}$$

where λ = operating wavelength
v = propagation velocity
f = operating frequency

Eqs. 5 and **6** will allow you to calculate the lengths of sections A, B, and C, while **eq. 4** gives the characteristic impedance of each section. The diameter of the cylindrical hole into which the choke section slides can be between 1/4 inch (6.35 mm) to 1/2 inch (12.7 mm) for X band. A compromise is to use 3/8 inch (9.5 mm).

Diode-mounting configurations. The diode can be mounted on a post centered in the waveguide. The post and mounted diode can excite TEM modes in the vicinity of the post, resulting in spurious responses that can cause oscillator turn-on problems. Post reactance can be eliminated if guide height is reduced to that of the diode package. However, results have shown that the tuning range is narrowed and power output drops.[15] If it's desirable to broadband the oscillator, a tapered or stepped post can be used. Again, because the device is a negative-resistance oscillator, broadbanding in this manner can lead to oscillation in unwanted modes and turn-on problems.

The resonant frequency of the TEM modes can be equated to post and diode height, being approximately a half-wavelength long. Mode frequency can be doubled by centering the diode on the post. The diode will cause a null in the fields at the point that makes the original post-height sections halved. However output power will drop, since mounting the diode in the post center decouples the diode. In many cases diode location on the post is left up to the experimenter.

Optimum post diameter is discussed in the literature.[16] Theoretical calculations verified by experimental results show that a post 0.125 inch (3 mm) in diameter gives the oscillator the broadest bandwidth. Post diameters greater than 0.150 inch (3.8 mm), reduce bandwidth over which the oscillator can operate. The power output, however, increases with post diameters greater than 0.150 inch (3.8 mm).

Cavity backwall. In some Gunn-oscillator designs the cavity backwall is fixed and located approximately one-half guide wavelength behind the diode post. For this configuration, some form of matching network is ahead of the diode. In the iris-coupled waveguide oscillator the backwall can also be fixed; however, this must be done by experimentation to obtain optimum results.

Once the backwall position is determined for optimum power output and stability, the position can then be fixed. From a flexibility standpoint, I found that a movable backwall permitted greater freedom of adjustment. This is particularly evident when different Gunn diodes are used in the same cavity.

Fig. 6 illustrates different movable backwall designs. **Figs. 6A** and **B** are quarter-wave choke sections, while **fig. 6C** is a close-fitting block that can be secured by locking screws once optimum operation is established. In **fig. 6A**, the quarter-wavelength sections are separated from the wall by nylon bearings. The mechanism is spring loaded, so that a constant pressure is exerted onto the choke assembly. An adjusting knob, riding on a threaded shaft, moves the choke assembly in and out.

Fig. 6B is identical to that of **fig. 6A**, except a dielectric is used around the entire sections that come in close contact with the waveguide walls. Note that these sections are narrower than the corresponding sections in **fig. 6A** because the insulated sections form a dielectric loaded guide.

Fig. 6C shows a simple block that is close-fitted in the guide. No attempt is made to insulate here; wall contact is desired with this design. In many cases, the block is cut to a depth of a quarter guide wavelength. A threaded rod acts as a handle for block adjustment.

The choke-section depth is based upon the idea that the space between guide sidewall and choke can be considered as a guide beyond cutoff (assuming TE_{10} mode). The space between the guide broad wall and choke is basically a reduced-height guide propagating in the TE_{10} mode. The relationship for λ_g is the same regardless of guide height. The choke depth is one-quarter guide wavelength. When air is in the dielectric, $\lambda_g/4$ is computed from:

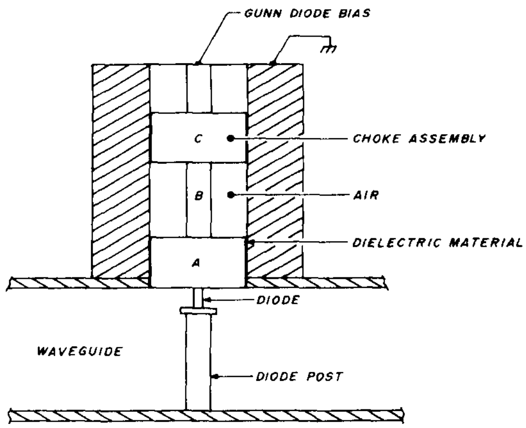


fig. 5. Cross section of the dumbbell rf choke. Sections A and C are identical. The dielectric insulates the center section from the wall. Sections A, B, and C form a coaxial transmission-line system of quarter-wavelength transformers with different characteristic impedances.

$$\lambda_g = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{2a}\right)^2}} \quad for\ TE_{10}\ mode \qquad (7)$$

where λ_g = guide wavelength
λ = operating wavelength
a = broad guide dimension

From **eq. 7** the value for λ_g is divided by four, and each section in **fig. 6A** is $\lambda_{g/4}$ long.

In **fig. 6B**, the value of λ_g for the two sections having dielectric tape wrapped around them is:

$$\lambda_{gd} = \frac{\lambda}{\sqrt{\epsilon_r-\left(\frac{\lambda}{2a}\right)^2}} \quad for\ TE_{10}\ mode \qquad (8)$$

where λ_{gd} = guide wavelength in dielectric guide
ϵ_r = relative dielectric constant

Here the first and third choke sections are $\lambda_{gd}/4$,

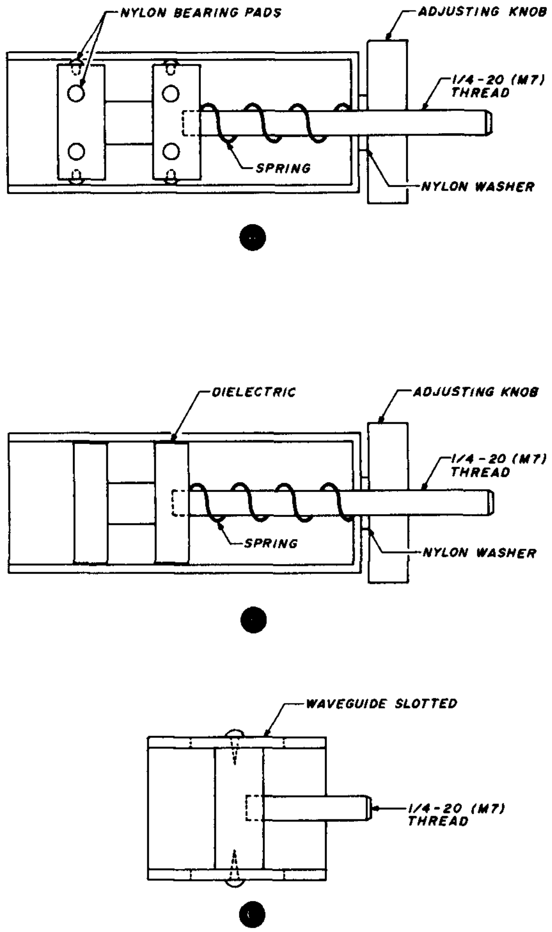


fig. 6. Different moveable backwall designs. An adjustable backwall choke assembly, using nylon pads for isolation, is shown in A. B shows a design using dielectric tape around the two sections. Wall contact is desired in the block design in C. The close-fitting adjustable backwall is locked into place by locking screws on each side of the waveguide.

while the middle section is $\lambda_g/4$. **Eq. 8** is an approximate solution, assuming that a dielectric with a low-loss tangent is used.

Matching section. This device is an adjunct to the oscillator in that, if there's a considerable mismatch between oscillator and load, some sort of matching device is needed. Several techniques are used in impedance matching. Devices such as E-H tuners and slide-screw tuners, have been fairly common. However, they've been replaced in many applications by a ferrite isolator. This device exhibits low attenuation in the forward direction and high attenuation in the reverse direction. They can be made broadband and present a constant load to the oscillator despite wide variation of load mismatch. However, they're rather expensive, so a simple approach is used.

While it's narrow band, a three-screw tuner will match over a limited range of mismatch conditions. **Fig. 7** is an example of a three-screw tuner. The screws are one-quarter guide wavelength apart and are placed along the center line of the broadwall of the guide. If a wider range of impedances is to be matched, 3/8 and 5/8 guide wavelength separations can be used if space permits. The distance between the iris and the first screw of the tuner can be made 3/8 guide wavelength.

oscillator design and assembly

Since an iris-coupled waveguide cavity oscillator can tune only over a portion of the 10-GHz band, several factors affect the choice of frequency. First, the frequency will tune downward with increasing bias for most diodes. Diodes are available where frequency increases with bias; these are employed where temperature compensation is required. Second, the frequency will tune downward as the tuning screw penetrates the cavity. Third, the frequency decreases with increasing temperature.

Because the effects of bias, temperature, and mechanical tuning lower the frequency, choose a frequency of, say, 50 MHz or so above the desired operating frequency. With an AFC system, the oscillator can stay locked onto either the incoming received signal or a reference signal.

Design example. In the following example an operating frequency of 10.350 GHz was chosen. To compensate for the effects mentioned previously 10.400 GHz became the design frequency. An iris diameter of 0.25 inch (6.35 mm) was used. Before the cavity length can be calculated from **eq. 2**, λ_g and b_e must be found:

$$\lambda = \frac{c}{f} = \frac{3 \times 10^8 m/s}{10.4 \times 10^9 Hz} = 1.13\ inch$$

$$(0.02885\ meters\ or\ 28.85\ mm)$$

$$\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{2a}\right)^2}} = \frac{28.85\ mm}{\sqrt{1 - \left(\frac{28.85\ mm}{2 \times 22.86\ mm}\right)^2}}$$

$$= 1.464\ inch\ (37.18\ mm)$$

From **fig. 4**, the value for b_e can be found, providing the following ratios are calculated:

$$\frac{d}{a} = \frac{6.35\ mm}{22.86\ mm} = 2.8$$

$$\frac{a}{b} = \frac{22.86\ mm}{10.16\ mm} = 2.25$$

$$\frac{\lambda}{a} = \frac{28.85\ mm}{22.86\ mm} = 1.26$$

Using these numbers in **fig. 4**, b_e is 14. Next substitute the values for b_e and λ_g in **eq. 2** and calculate cavity length:

$$\ell = \frac{\lambda_g}{2\pi}\left[\pi - \frac{1}{2}\tan^{-1}\frac{2}{|b_e|}\right]$$

$$\ell = \frac{37.18\ mm}{2\pi}\left[\pi - \frac{1}{2}\tan^{-1}\frac{2}{14}\right]$$

$$0.717\ inch\ (18.2\ mm)$$

The tuning screw can be placed anywhere between the diode post and the iris; however, a position of 0.22 inch (5.6 mm) from the iris was chosen, since placing it too close to the diode post would cause unstable operation.

Bias choke and diode post. Referring to **fig. 5**, the lengths of sections A, B, and C are calculated. Since sections A and C are sections of coaxial line with Mylar insulation, propagation velocity will be altered by $\sqrt{\epsilon_r}$ of the Mylar. Using **eqs. 5** and **6**:

$$v = \frac{c}{\sqrt{\epsilon_r}} = \frac{3\times 10^8\ m/s}{\sqrt{2.8}}$$
$$= 5.9 \times 10^8\ feet/second$$
$$(1.79 \times 10^8\ meters/second)$$

$$\lambda = \frac{v}{f} = \frac{1.79\times 10^8\ m/s}{10.25\times 10^9 Hz}$$
$$= 0.688\ inch\ (17.5\ mm)$$

In the above instance, the frequency for the middle of the 10-GHz band was chosen, i.e., 10.25 GHz. The length of sections A and C are $\lambda/4$, therefore:

$$\frac{\lambda}{4} = \frac{17.5\ mm}{4} = 0.17\ inch\ (4.4\ mm)$$

Section B (**fig. 5**) has air dielectric, therefore $\sqrt{\epsilon_r} = 1$ and its length becomes:

$$\lambda = \frac{c}{f} = \frac{3\times 10^8\ m/s}{10.25\times 10^9\ Hz} = 1.15\ inches\ (29.27\ mm)$$

$$\frac{\lambda}{4} = \frac{29.27\ mm}{4} = 0.28\ inch\ (7.3\ mm)$$

The characteristic impedance must be low for section A and C and high for section B. Based on choke design appearing in D. Evan's articles,[17,18,19] the outer diameter of the choke cylinder is 0.375 inch (9.525 mm). Sections A and C are chosen so that the spacing accommodates one or two layers of Mylar tape. The diameters of A and C are then 0.360 inch (9.144 mm). The diameter of B is approximately 1/3 of that of A and C. Section B diameter is chosen as 0.125 inch (3.175 mm). Using these diameters in **eq. 4**, the characteristic impedances of the sections are:

Sections A and C:

$$Z_0 = \frac{60}{\sqrt{\epsilon_r}}\ \ell n \frac{r_2}{r_1}$$

$$Z_0 = \frac{60}{\sqrt{2.8}}\ \ell n \frac{9.525\ mm/2}{9.144\ mm/2} = 1.46\ ohms$$

Section B:

$$Z_0 = \frac{60}{\sqrt{1}}\ \ell n \frac{9.525\ mm/2}{3.175\ mm/2} = 65.9\ ohms$$

The section above C maintains the same diameter as in section B and is made long enough to connect to a BNC connector. The Gunn diode connects to section A. **Fig. 8** is an outline dimension for a typical low-to-medium power Gunn diode package. Each end is 0.061 inch (1.56 mm) in diameter. The mating hole in section A is made slightly larger; i.e. 0.0625 inch (1.59 mm) in diameter and at least as deep. The diode post is made from a 10-32 (M5) screw. The end opposite the head has a hole drilled out to 0.0781 inch (1.98 mm) diameter. The depth of the hole is at least 0.0625 inch (1.59 mm). The larger hole is needed to avoid fracture of the diode caused by misalignment errors.

The quarter-wave dimensions for movable backwall choke assembly are based on the design shown in **fig. 6B**. Here the first and third sections are loaded with Mylar tape. The inner guide dimension is 0.9×0.4 inch (22.86 mm×10.16 mm). The choke

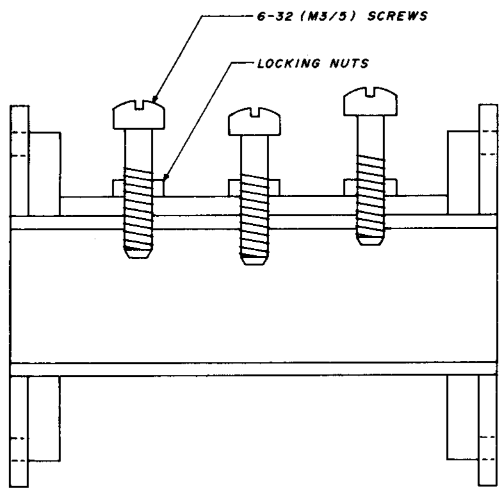


fig. 7. Cross section of the triple-screw matching section. Screws are one-quarter guide wavelength apart and are placed along the centerline on the guide broad wall. Spacings of 3/8 and 5/8 guide wavelengths can be used if space permits, since such spacings match a wider range of impedances.

block must be made smaller by the amount of the Mylar tape. If 5-mil-thick tape is used, the choke block is made with at least a 7-mil clearance on each side. These blocks then become 0.886 × 0.386 inch (22.50 mm × 9.80 mm) in their cross-sectional dimensions. Their length is computed from **eq. 8**:

$$\lambda_{gd} = \frac{\lambda}{\sqrt{\epsilon_r - \left(\frac{\lambda}{2a}\right)^2}}$$

$$= \frac{28.85\ mm}{\sqrt{2.8 - \left(\frac{28.85\ mm}{2 \times 22.86\ mm}\right)^2}}$$

$$= 0.733\ inch\ (18.62\ mm)$$

$$\frac{\lambda_{gd}}{4} = \frac{18.62\ mm}{4} = 0.183\ inch\ (4.66\ mm)$$

This value is the length of the first and third sections. The middle section length is computed from **eq. 7**:

$$\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{2a}\right)^2}}$$

$$= \frac{28.85\ mm}{\sqrt{1 - \left(\frac{28.85\ mm}{2 \times 22.86\ mm}\right)^2}}$$

$$= 1.464\ inch\ (37.18\ mm)$$

$$\frac{\lambda_g}{4} = \frac{37.18\ mm}{4} = 0.37\ inch\ (9.30\ mm)$$

The remaining calculation is that of the matching transformer or tuner. The first screw of the three-screw tuner shown in **fig. 7** is approximately 3/8 λ_g away from the iris. The distance between the screws is $\lambda_g/4$. Using 10.25 GHz, a band center, $\lambda_g = 1.5$ *inches* (38.10 mm). The distance between screws is 0.375 inch (9.5 mm), and between screw and iris is 0.563 inch (14.3 mm). When the sections were built, this dimension was closer to 0.590 inch (15 mm). The discrepancy produced no apparent problem.

construction

Now that all critical dimensions have been calculated, **figs. 9A** and **B** show the oscillator assemblies. The oscillators operate around 10.4 GHz and 10.3 GHz respectively. **Fig. 9B** differs from **A** mainly in backwall design. The construction of the oscillator follows a sequence of steps.

Cavity. Referring to **fig. 9A**, the cavity assembly is made from a piece of WR-90 waveguide about 2.5 inches (63.5 mm) long. If you have a drill press, milling machine, and lathe, the job becomes easier. However, much of the construction can be done if

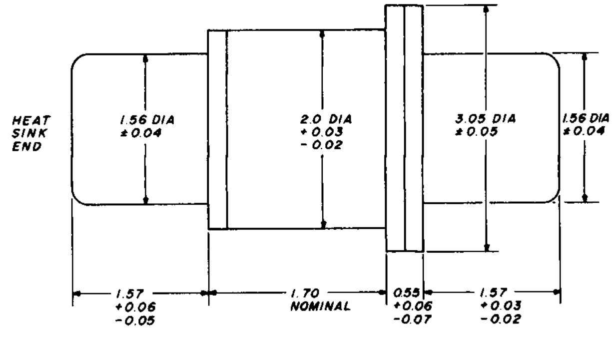


fig. 8. Outline dimensions of a typical low-power Gunn diode encapsulation. Diodes with output power ratings greater than 50 mW (such as the Alpha Industries DGB-6835C) have the cathode on the heatsink end. Those with less than 50 mW output power ratings (such as the Microwave Associates MA-49508) have the anode on the heatsink end.

you have some hand tools and a drill press.

Square off both ends of the guide and deburr. Obtain a piece of brass stock 0.75 × 1 × 1 inch (19 × 25.4 × 25.4 mm) and another 0.125 × 0.75 × 1 inch (3.175 × 19 × 25.4 mm). Fit a UG-39/U cover flange over one end of the guide and clamp the two brass pieces on each side. Make a waveguide cap for the other end of the guide out of a piece of 1 × 0.5 inch (25.4 × 12.7 mm) flat stock 60 mils thick. Place this assembly vertically on a fire brick. Rest the waveguide cap over the open waveguide. The flange end should be flush with the brick. Solder the assembly using a torch and let cool. File the flange end smooth; or if a milling machine is available, mill it smooth.

Measure a point 0.717 inch (18.2 mm) back from the flange and center it on the broad dimension of the guide on the 1 inch (25.4 mm) high block behind the flange. Drill a small pilot hole through the entire assembly using a 3/32-inch (2.38 mm) diameter drill. This ensures the alignment of the bias choke with the diode post. On the bottom side, drill and tap for a 10-32 (M5) thread. On the choke block, drill a 3/8-inch (9.525 mm) diameter hole.

A word of caution here — start by drilling progressively larger holes until the required diameter is reached. Use a slow drill speed and oil. The drill has a tendency to grab, and you may have to anchor the assembly to the work table on the drill press.

Next, locate the holes of a UG-290/A BNC connector on the top of the bias-choke housing. Drill and tap four 4-40 (M3) screws 0.5 inch (12.7 mm) deep. Drill and tap a 6-32 (M3/5) hole 0.22 inch (5.6 mm) back from the flange face and centered on the broadwall in front of the bias block for a 6-32 (M3/5) nylon tuning screw. Drill a 1/4-20 (M7) clearance hole centered on the waveguide cap to accommodate the shorting choke assembly screw mechanism.

Iris. The iris is made from thin copper foil at least 10 mils thick. One of the UG-39/U flanges can be used as a pattern as an outline to locate the center. Cut out the iris and scribe a mark locating the center of the iris hole. Carefully drill a 0.25-inch (6.35 mm) diameter hole. Then drill out the four corner holes to clear 8-32 (M4) screws. (Later, the iris plate will be clamped between the oscillator assembly and the tuner assembly.)

Bias choke. To make the bias choke, a lathe is handy since it makes the job easier but isn't absolutely necessary. Obtain a 3/8-inch (9.525 mm) diameter brass rod about 4 inches (101.6 mm) long and square off one end. Referring to **fig. 5**, mark off sections A, B, and C. These will be 0.172 inch (4.4 mm) 0.288 inch (7.32 mm), and 0.172 inch (4.4 mm) respectively. Above section C, a length of 0.236 inch (6 mm) can be cut back and attached to the bias connector.

Sections A and C are reduced in diameter 0.360 inch (9.144 mm) and B and the section above C to 0.125 inch (3.175 mm) in diameter. A 0.0625 inch (1.59 mm) diameter hole is drilled on center in the end of section A. Wrap a layer of Mylar tape around A and C. Fit the bias choke into the bias block and set it flush with the inside top of the waveguide. Cut the section of the choke above section C to mate up with the center conductor on the BNC connector and solder. The assembly is secured into place by four 4-40 (M3) screws.

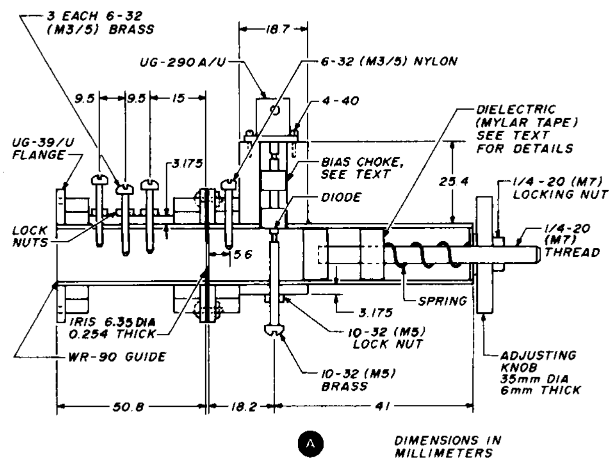
Diode post. The diode post is made from a 10-32 (M5) screw with a 0.078 inch (1.98 mm) diameter hole drilled into the center of the screw opposite the head. When the diode post is put into position, a 10-32 (M5) locking nut secures it.

Backwall. The movable backwall is made from two brass blocks 0.886 × 0.386 inch (22.50 × 9.80 mm) in cross section by 0.183 inch (4.66 mm) long, and a 2.5 inch (63.55 mm) long 1/4-20 (M7) threaded rod. Drill a hole, centered on the face of the block, and tap it for 1/4-20 (M7) thread through one block and half way through the other. Screw the rod into one block. Screw down the other block until it's about 0.366 inch (9.30 mm) from the first block. Wrap a layer of Mylar tap around each block. Slide a spring over the free end of the 1/4-20 (M7) rod.

The assembly is backed into the flange end of the oscillator cavity, and the rod goes through the clearance hole in the cap end of the waveguide. An adjusting knob, threaded for 1/4-20 (M7) is screwed on and used to move the choke assembly. A 1/4-20 (M7) is screwed on and used to move the choke assembly. A 1/4-20 (M7) locking nut is secured when the optimum position for the choke is found.

Place the diode, with heat-sinking compound, into the diode post screw. Carefully insert the screw and diode and lock into place. Do not exert too much pressure (see **"precautions"**). Lock into place with the locking nut. Insert the 6-32 (M3/5) nylon tuning screw at this point.

Tuner. The triple-screw tuner is made from a 2-inch (50.8 mm) section of WR-90 guide. The ends should be squared off and deburred. Solder a brass section 1.125 inch (28.58 mm) long by 1 inch (25.4 mm) wide by 0.125 inch (3.175 mm) thick onto the broad side of the waveguide between the two flanges. Clamp the plate to the waveguide, slide the flanges over each end of the waveguide and solder the assembly. Drill and tap three 6-32 (M3/5) holes centrally along the broad waveguide face through the block according to the locations in **fig. 9A**. Deburr the holes and



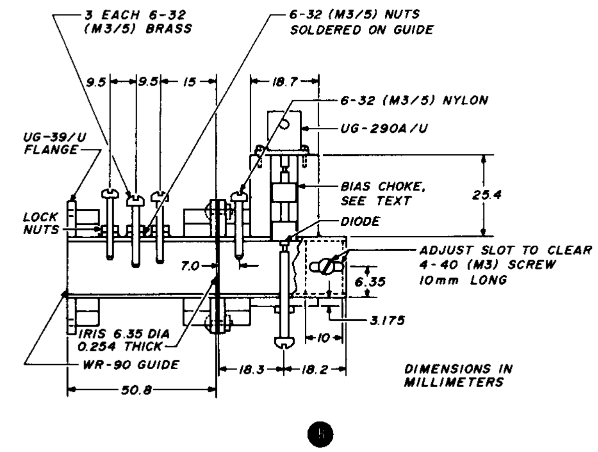


fig. 9. Assembly drawings of two Gunn oscillators. A design using an adjustable choke plunger backshort is shown in A. The oscillator in B has an adjustable contacting backshort that can be locked into place.

insert three 6-32 (M3/5) screws 0.75 inch (19.05 mm) long with locking nuts. Connect the tuner to the oscillator section. Place the iris plate as shown and use four 8-32 (M4) screws and nuts to hold the assembly together.

This completes the oscillator assembly. The unit in **fig. 9B** is constructed in the same manner except for the backwall. Here a block closely fitting the inside of the guide is clamped into place on each side by a 4-40 (M3) screw.

test techniques and results

The type of tests that can be made depends on what equipment is available. If you've been able to obtain a fair amount of X-band test equipment, including a frequency meter and a slotted line from surplus dealers or flea markets, you're in good shape. Otherwise, a minimum of test equipment can be borrowed or made. For making some of your X-band test equipment consult reference 20.

Reflectometer test. For those who have access to an X-band reflectometer, looking at the reflected power from the cavity will show where the oscillator cavity resonates and where any spurious resonances occur within the 8.5-12.4-GHz band. Moving the backwall and tuning screw gives an indication of the tuning range. However, when power is applied to the oscillator, these frequencies will shift slightly. If a coaxial-to-waveguide transition is available, a transmission test is made to determine 10-GHz leakage out of the bias choke. The results on both oscillators of **fig. 9** indicate the leakage is down by 60 dB or more.

Slotted line measurements. If a slotted line is available, the reactance of the iris can be measured. I built several irises and the differences between calculated and measured reactance were within 5 per cent. If the values of normalized reactance are calculated carefully from **fig. 4**, they should be within 5 per cent of the actual values.

Swept-bias tests. Probably the most important testing technique used on Gunn oscillator design is

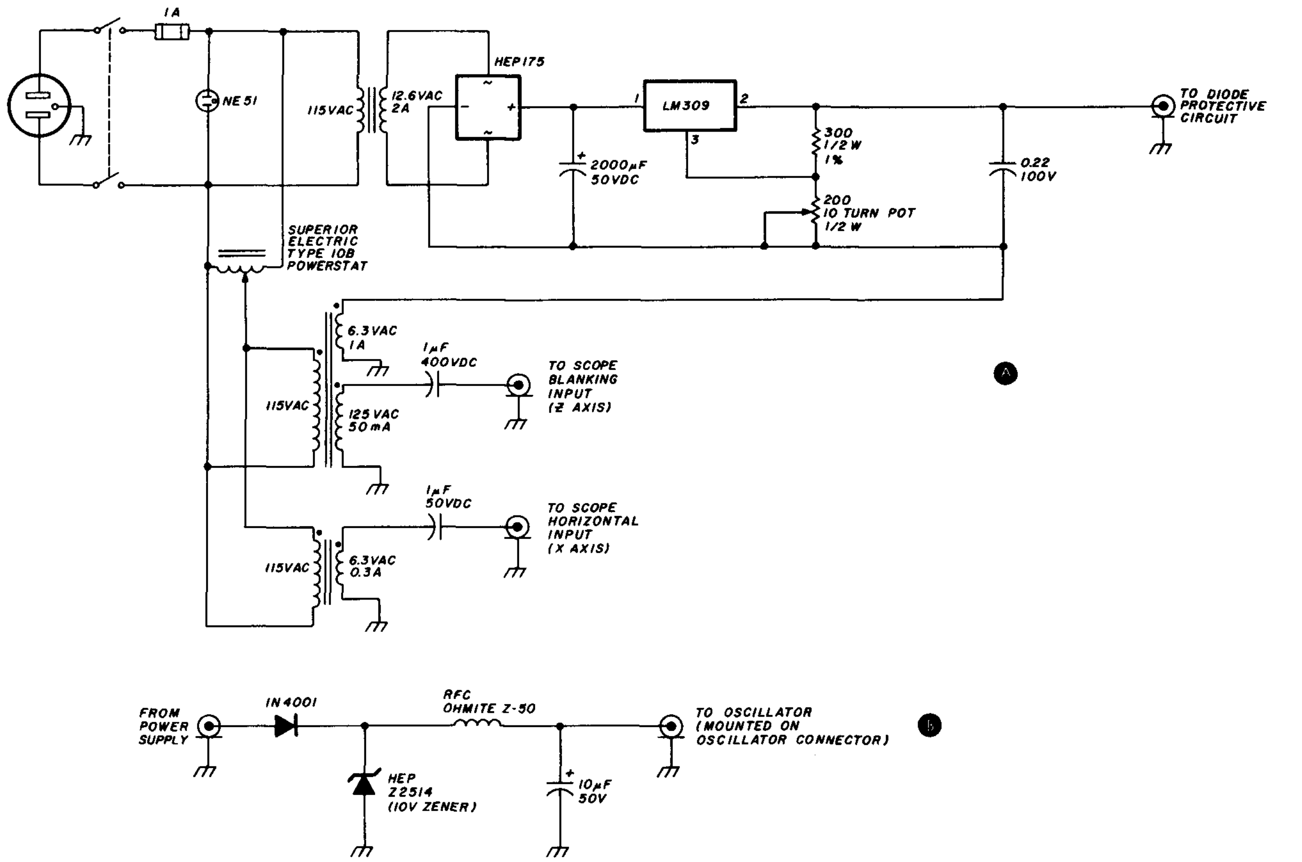


fig. 10. Schematic of the swept bias supply for testing Gunn-diode oscillators. In (A), a dc supply provides regulated dc voltages up to 500 mA. The 6.3 Vac winding of a transformer provides a variable ac supply, which is in series with the dc supply. Both supplies are connected to the Gunn oscillator through the protective circuit in (B).

the swept-bias test setup. A variable bias voltage is applied to the oscillator while its output is monitored with a detector and frequency meter. The swept response is displayed on an oscilloscope. Using such a test setup, the oscillator can be made to look into various loads while making different adjustments toward a stable operating point.

Connect the oscillator and tuner to a load through a directional coupler. Connect a frequency meter and detector to the directional-coupler coupling arm.

The swept-bias supply circuit shown in **fig. 10A** is made of several supplies. A dc supply provides a regulated 5-10 volts dc. The supply can also provide up to 500 mA of current. The 6.3 Vac winding of a transformer, providing a variable ac supply, is in series with the dc supply. Both are connected across the Gunn diode oscillator through the protective circuit, **fig. 10B**. Provisions are made for synchronized sweep voltage and a blanking voltage. Correct winding sense must be observed on the transformer windings to ensure proper sweep direction and blanking.

The protective circuit has a diode in series to protect the Gunn diode from reverse bias voltages and a zener diode to protect it from overvoltage in the forward-bias direction. The Z-50 choke and electrolytic capacitor ensure that a low impedance is presented to the oscillator and prevents buildup of dangerous voltage levels from parasitic resonances.

The protective circuit is connected to the swept source through a coaxial cable. A dummy load is then connected to the protective circuit. The dc voltage is set to 5 volts, and the sweep voltage is adjusted for 6 volts peak-to-peak if the MA-49508 diode is used, or 10 volts peak-to-peak if the DGB-6835C diode is used.

Note the powerstat settings for these ac voltages. Set the ac voltage to zero, turn off the supply, and remove the dummy load. Note: The protective circuit parts mount onto the Gunn oscillator. A matched load is used on the Gunn oscillator setup, and the nylon screw is backed out so that it's even with the inside top wall of the waveguide. The protective circuit box is connected to the oscillator. Turn on the power and advance the swept voltage to the preset position. Move the backshort until a stable swept response, shown in **fig. 11D**, is obtained. (**Figs. 11A** through **11C** show various stages of *misadjustment*.) The screws on the tuner can be adjusted for optimum output power while maintaining oscillation. If a particular antenna is to be connected, turn off the supply, connect the antenna, turn on the supply and make adjustment.

Fig. 11D shows that, under proper operation, the onset of oscillations starts abruptly once the bias voltage exceeds the Gunn-diode threshold voltage.

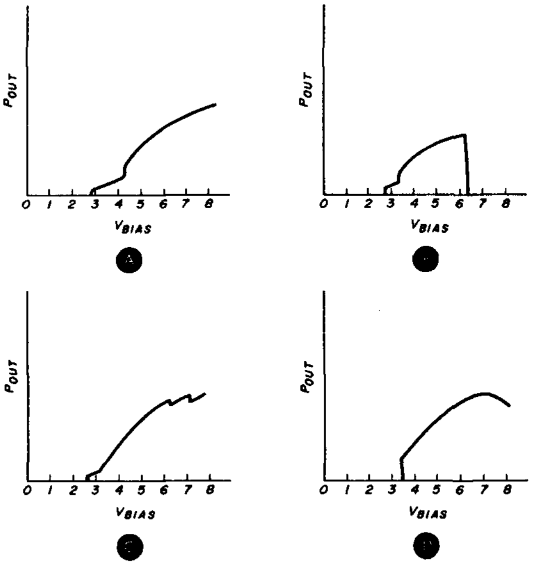


fig. 11. What to expect while running tests. Graphs A through C show respectively poor starting characteristics, oscillator-to-load mismatch, and backwall partially under the bias choke. Graph D shows the proper relationship between power output and bias.

As the bias voltage increases, output power increases to a peak and drops suddenly. Frequency usually decreases with increased bias voltage.

Modulation sensitivity tests were performed on the two types of Gunn diodes. The MA-49508 diode modulation sensitivity measurement was 15.6 MHz/volt, while the DGB-6835C modulation sensitivity measurement was 11.7 MHz/volt (both at room temperature).

For comparison, the published data on these diodes indicate an average modulation sensitivity around 7 MHz/volt at room temperature. The current through the diode remained fairly constant during these measurements.

I made a rough check of temperature effect on the oscillation frequency. The MA-49508 diode frequency drift with temperature was measured at −290 kHz/degree C compared to the maximum of −350 kHz/degree C on the data sheet. The DGB-6835C diode measured −0.87 MHz/degree C to the maximum −1 MHz/degree C on its data sheet.

some precautions

The Gunn diode, like any semiconductor device, can be damaged by an electrostatic discharge. Therefore, use care when handling the diodes. Mechanically, the diodes are fairly rugged upon compression. However, they can be damaged by shear fracture. This particular failure mode occurs when placing the diode into the oscillator cavity and tightening the diode post screw too much, especially if some axial misalignment exists in the diode-socket holes.

Check diode polarity before power is applied. In many cases positive bias with respect to ground is used. The medium- and higher-power diodes, such as the Alpha DGB-6835C, are heat sinked at the cathode end. The diode cathode end makes firm contact with the diode post, which is usually a good heatsink.

On the other hand, a lower-power diode, such as the MA-49508, is constructed with the heatsink on the anode end. It's important to note that, with this type of diode configuration, the diode is physically reversed inside the package. Therefore, if the cathode end is the grounded electrode, the package must be *physically reversed* before mounting it into the oscillator. If this is not done, the diode will be reversed-biased and will be damaged.

Another cause of diode damage is parasitic oscillation. Since the Gunn oscillator is a negative resistance device, oscillations will occur at any spurious resonance that exists from hf to the microwave frequencies. Oscillations can exceed the maximum bias voltage on the diode. If not by-passed at the bias choke cold end, any length of bias line can form a resonant system, and the oscillator may put out power at that particular frequency.

For this reason, it's desirable to mount any modulation/bias circuitry close to the oscillator. Any bias modulation circuit should present a low impedance to the Gunn oscillator. Check with the Gunn diode manufacturer for their recommended protection circuit.

One very important precaution must be mentioned regarding Gunn oscillators (or any microwave oscillator). *Do not* look into the open end of the waveguide while power is applied to the oscillator. Close up, rf power density can exceed OSHA's *10mW/cm²* safety limit. Fortunately, the rf power density falls off to a safe level a short distance from the oscillator. Your eyes are especially susceptible to damage from rf power radiation, so never look into an open waveguide or stand in front of a microwave antenna!

conclusion

I've presented some of the details in the design of a Gunn oscillator. Some of the parameters and conditions for building a working oscillator are given. Also presented is a detailed design and assembly description of the oscillator. Test results are given, and some precautions are stated since it's possible to damage these devices if not properly treated.

Detailed circuits to drive and modulate the diode were not presented, since this is a different subject. Further information can be found in references 21-27. The main purpose of this article was to present how a Gunn oscillator can be built and tested.

acknowledgments

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ham radio

measuring capacitance of electrolytic capacitors

Some simple math and a volt-ohmmeter make it possible to evaluate your electrolytics

You can build or buy many instruments that will tell you, with reasonable accuracy, the value of a capacitor. The capacitance of electrolytics is not easily determined, however, because they are polarized and tend to have sizeable leakage current. The following material provides a method for evaluating electrolytic caps.

the electrolytic capacitor

Electrolytic capacitors are made of either aluminum or tantalum electrodes and an electrolyte. Boric acid is the usual electrolyte used with aluminum, and sulfuric acid is frequently used with tantalum. The active electrode may be etched; or in some cases, tantalum is formed as a porous plug. In either case, a thin insulating layer is formed on the active electrode. This layer acts as the dielectric in which the capacitive energy is stored. In a well-behaved electrolytic cap, this thin layer looks a lot like a rectifier. That is, current flow in the polarized direction is small, while in the reverse direction it can become quite large. It is this polarized state which makes bridge-type capacitance measurements difficult.

uses

Because it's possible to obtain a large quantity of microfarads in a small package, electrolytics are used as filter elements in power supplies. Here they function as energy-storage devices, charging when the voltage is high then supplying current when the voltage starts to drop. Electrolytics are also used as coupling elements. The large capacitance acts as a low alternating-frequency impedance between two points in a circuit where a difference in direct-current potential exists. In effect, this application also uses the capability of the electrolytic to store energy.

Nothing lasts forever, and this is particularly true of electrolytic capacitors. Loss of liquid, degeneration of the dielectric film, and abuse may drastically change the basic capacitance of the unit. The question then is, How do you measure the capacitance of an electrolytic? One answer, which goes back to fundamentals, is to measure its *energy storage ability*.

mathematical derivation

Almost every Radio Amateurs' handbook gives the relationship between supply voltage, voltage across a capacitor, and time as a function of the charging resistance. This is an exponential relationship, but need not cause concern for the nonmathematically inclined. If you're interested in the basic math, finish this section; if not, just jump to the next.

First, the resistance and capacitance (RC) involved in a charge or discharge circuit, when multiplied together, give what is called the "time constant." If you have a large value of resistance, the time to charge or discharge will be long. The same may be said for capacitance. If the RC time constant is large, the time to charge or discharge is large.

We'll concern ourselves with discharge, in which the voltage across a capacitor with a shunt resistor is expressed as follows:

$$V = \frac{e_0}{exp\,(t/RC)} \tag{1}$$

The term e_0 is the voltage to which the capacitor was charged at *time zero*. *Exponent* (t/RC) is 2.718 raised to the value of t/RC. Here t is the elapsed time in seconds since the resistance and capacitance started to discharge. Note that if t is just equal to RC, we have 2.718 raised to unity power, or simply 2.718.

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With the technical details out of the way we can now attack the problem. If we measure the time required for a capacitor to reach 1/2.718 of the voltage to which it was originally charged, then we know the value of the product of the resistance and capacitance. Divide the time by the resistance and you end up with the capacitance. Simple, isn't it?

measurement example

Now for a practical example. Suppose you use a volt ohmmeter with a 20,000-ohm/volt dc sensitivity, and a capacitor has a rated voltage of 35 volts. You'll use the 50-volt scale, so the effective resistance of the meter across the capacitor will be 50 times 20,000 or 10^6 ohms. This is a rather large resistance if the capacitor is several thousand microfarads, since the time constant will be several thousand seconds. Use a 47-k resistor in parallel with the meter so that the R in the circuit is

$$R = \frac{1}{\left(\frac{1}{10^6} + \frac{1}{4.7 \times 10^4}\right)} \; ohms$$

$$= 4.49 \times 10^4 \text{ or } 44,900 \; ohms$$

(round off to *45 kilohms*)

We'll assume a) that the capacitor to be tested has been charged long enough so that it is reformed and up to capacitance, and b) we'll start with 35 volts. The voltage at which timing will be stopped will be 35/2.718, or 12.88 volts.

Connect the VOM across the capacitor and the 47k resistor, charge the capacitor to 35 volts, then note the time at which you remove the power from the circuit. Note that you must remove one lead from the power supply. Simply turning off the supply is not enough. Now watch the voltage. When it reaches 12.88 volts, again note the time. The length of time elapsed, in seconds, is the time constant. Suppose it was 83 seconds. This value divided by the resistance, 45 kilohms, is 1844 microfarads, the effective capacitance of the unit being measured.

closing remarks

This method of measurement doesn't take into account any leakage effects, so the capacitance you measure may be a bit less than the actual dynamic capacitance. Also recognize that, when dealing with high-voltage capacitors, there is danger of lethal shock and you should be extra careful of any voltage over, say, 15 volts. Another precaution concerns direct shorting a capacitor. Most units will survive the current surge from a screwdriver short; however some may not and it's better to use a resistor than to burn up a screwdriver.

ham radio

appreciating the L matching network

Theory and application of the L network in Amateur circuits

Often one looks at transmitter schematics and notices the simple two-element networks in the shape of the letter L, which are used to match a source impedance to a load impedance. These seemingly simple devices, which display high efficiencies, often baffle many Amateurs. The thought of using two completely reactive components to transform the value of a resistor by itself seems intriguing. Such simple circuits find their way into almost every piece of communications equipment; matching transistor input and output, matching a reactive antenna to a flat line, and matching a driver stage to a final amplifier. One drawback of the L network is that it is an *exact match at only one frequency*.

Let's look into the L network to try to understand it better. But we'll try to use less mathematics and more intuition to derive the equations. Once we've done this, we'll then be able to confidently use the networks not only to transform values of purely resistive loads to match a source resistance, but we'll also be able to match any complex impedance as well. The first step in understanding the L network is to examine the series-to-parallel impedance conversions.

series-parallel impedance conversions

A series combination of a resistor and an inductor or capacitor may be transformed into an equivalent parallel circuit. If the impedance of two components inside a black box were measured at a single frequency, you would have two answers: a series circuit and its equivalent parallel combination, **fig. 1**.

To understand the series equivalent of a parallel circuit or *vice versa*, begin by writing the simple expression for the impedance across two parallel elements, Z_p, and later try to shape it into a series circuit.

$$Z_p = \frac{(jX_p)(R_p)}{R_p + jX_p} \qquad (1)$$

where j indicates the imaginary or reactive component ($j = \sqrt{-1}$).

To obtain a real denominator, multiply both numerator and denominator of **eq. 1** by the complex conjugate $R_p - jX_p$. The resultant is

$$Z_p = \frac{(jX_pR_p)(R_p - jX_p)}{R_p^2 + X_p^2} \qquad (2)$$

where $j^2 = -1$.

Next, separate the equation into the real (resistive)

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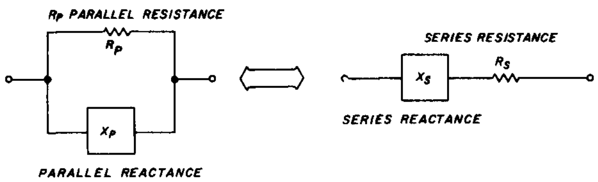

fig. 1. A series combination of resistance and reactance can always be found that exhibits the equivalent impedance of any given parallel combination of resistance and reactance.

and imaginary (reactive) components by grouping all terms with js by themselves.

$$Z_p = \left(\frac{X_p^2 R_p}{R_p^2 + X_p^2}\right) + j\left(\frac{X_p R_p^2}{R_p^2 + X_p^2}\right) \quad \textbf{(3)}$$

Thus an equation in the form of a series combination of a real (resistive) component and an imaginary (reactive) component results:

$$Z_s = R_s + jX_s = \text{series equivalent impedance} \quad \textbf{(4)}$$

$$R_s = \frac{X_p^2 R_p}{R_p^2 + X_p^2} = \frac{R_p}{1 + (R_p^2/X_p^2)} \quad \textbf{(5)}$$

$$X_s = \frac{X_p R_p^2}{R_p^2 + X_p^2} = \frac{X_p}{1 + (X_p^2/R_p^2)} \quad \textbf{(6)}$$

If circuit Q is introduced, the expressions are simplified. The Q for a parallel circuit is R_p/X_p, whereas that for a series circuit is X_s/R_s. Thus, the series equivalent elements are

$$R_s = R_p/(1 + Q^2) \quad \textbf{(7)}$$

$$X_s = X_p/(1 + 1/Q^2) \quad \textbf{(8)}$$

A transformation from a series to the equivalent parallel circuit can be made similarly:

$$R_p = R_s[1 + (X_s^2/R_s^2)] = R_s(1 + Q^2) \quad \textbf{(9)}$$

$$X_p = X_s[1 + (R_s^2/X_s^2)] = X_s(1 + 1/Q^2) \quad \textbf{(10)}$$

The sign of the reactive component doesn't change; the series equivalent of a parallel capacitance is still a capacitor.

deriving the L network

Two types of matching L networks are shown in **fig. 2, A** and **B**. One resistance, $R1$, is matched to a smaller resistance, $R2$. The shunt element is across the larger resistor, and the series element is connected to the smaller resistance. The method of deriving the equations will be to transform all the elements into the same form and make the reactances cancel.

For the circuit in **fig. 2A**, begin by transforming the series combination of $L1$ and $R2$ into a parallel equivalent circuit, which then forms the parallel circuit of $R1$, $C1$, $L1_p$, $R2_p$ (**fig. 2C**). The parallel equivalent of $R2$ is:

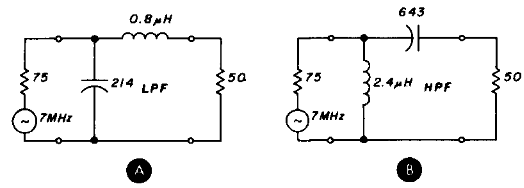

fig. 3. Components for matching networks using a (A) lowpass filter do not match those of the highpass arrangement, (B).

$$R_p = R_s[1 + (X_s/R_x)^2] \quad \textbf{(11)}$$

$$R2_p = R2[1 + (X_{L1}/R2)^2] \quad \textbf{(12)}$$

This transformed value of $R2$ must equal $R1$ for impedance matching. If **eq. 12** is solved for X_{L1} (the inductive reactance of $L1$), substituting $R1$ for $R2_p$ in the equation yields

$$X_{L1} = R2\sqrt{(R1/R2) - 1}\,;\; L1 = X_L/2\pi f \quad \textbf{(13)}$$

Thus the value of $L1$ is based on the resistor ratios. Next transform inductance $L1$ into its equivalent parallel value:

$$X_p = X_s[1 + (R_s/X_s)^2] \quad \textbf{(14)}$$

$$X_{LP} = X_L[1 + (R2/X_L)^2] \quad \textbf{(15)}$$

$$X_{LP} = R1/\sqrt{(R1/R2) - 1} \quad \textbf{(16)}$$

Now notice that we have a parallel circuit (**fig. 2C**). The transformed parallel inductance, X_{LP}, can be cancelled by an equal and opposite parallel capacitive reactance:

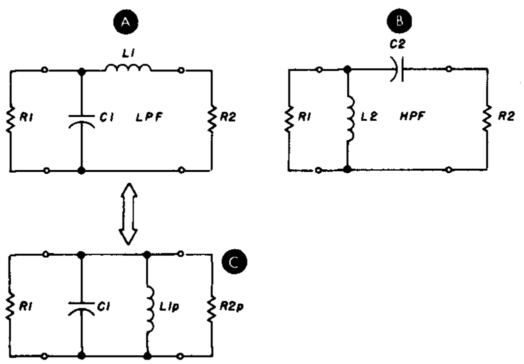

fig. 2. The L network may assume the form of either (A) lowpass or (B) highpass filter. A parallel equivalent circuit is shown in (C).

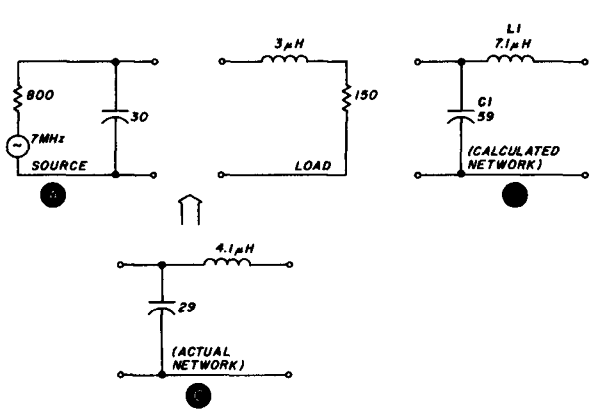

fig. 4. At 40 meters the source and load impedance, (A), partially absorb the L calculated network reactances, (B), resulting in a new network (C).

$$X_{C1} = X_{LP} = R1/\sqrt{(R1/R2)-1}\,;$$
$$C1 = 1/2\pi f X_C \qquad (17)$$

Thus series inductor $L1$ increased $R2$ resistance until it matched $R1$, then shunt capacitor $C1$ cancelled the reactance of the inductor, as in a parallel tank circuit. For the network in **fig. 2B**, simply transform the series arrangement of $C2$ and $R2$ into a parallel-equivalent circuit, with the capacitance value determined by the necessary resistance transformation and with inductance $L2$ of sufficient value to cancel the transformed capacitive reactance.

For each of these networks we could have alternatively transformed the parallel combination on the left into a series-equivalent circuit to match $R2$ impedance. The value of the other reactance would cancel the series equivalent.

resistive matching

Now we'll match a resistive generator to a resistive load. An example might be matching a 75-ohm generator, such as the output impedance of an oscillator, to a 50-ohm load (input impedance of a grounded-grid amplifier) on 40 meters. To accomplish this use either configuration **fig. 2A** or **2B**. The larger resistance is always designated $R1$.

$$R1 = 75\,ohms, R2 = 50\,ohms, f = 7\,MHz$$

Network 2A:

$$X_{L1} = R2\sqrt{(R1/R2)-1} = 50\sqrt{(75/50)-1} = 35.3\,ohms$$
$$L1 = X_{L1}/2\pi f = 0.8\,\mu H$$
$$X_{C1} = R1/\sqrt{(R1/R2)-1} = 75/\sqrt{(75/50)-1} = 106\,ohms$$
$$C1 = 1/2\pi f X_{C1} = 214\,pF$$

Network 2B:

$$X_{L2} = R1/\sqrt{(R1/R2)-1} = 75/\sqrt{(75/50)-1} = 106\,ohms$$
$$L_2 = X_{L2}/2\pi f = 2.41\,\mu H$$
$$X_{C2} = R2\sqrt{(R1/R2)-1} = 50\sqrt{(75/50)-1} = 35.3\,ohms$$
$$C2 = 1/2\pi f X_{C2} = 643\,pF$$

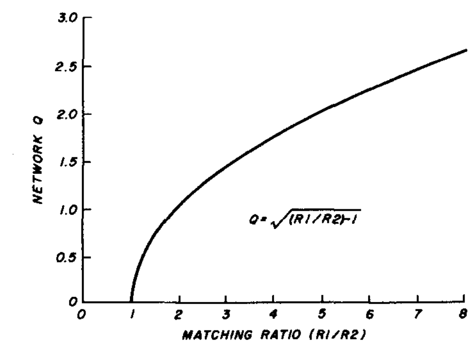

fig. 6. Circuit Q is determined by the ratio of the resistors to be matched.

Thus the same impedances can be matched with two different sets of components, **fig. 3**.

The principal difference between networks **2A** and **2B** (**fig. 2**) is that network **2A** is a lowpass filter and network **2B** is a highpass filter. For most applications it's desirable to suppress harmonics; thus network **2A** is chosen. It turns out that network **2A** is also easier to construct, because one end of a variable capacitor can be physically grounded to decrease lead inductance.

complex impedance matching

Now that we can match one resistor to another, we'll try to match complex impedances. Begin by designing an L network for matching the resistive part of a source to the resistive part of the load, then

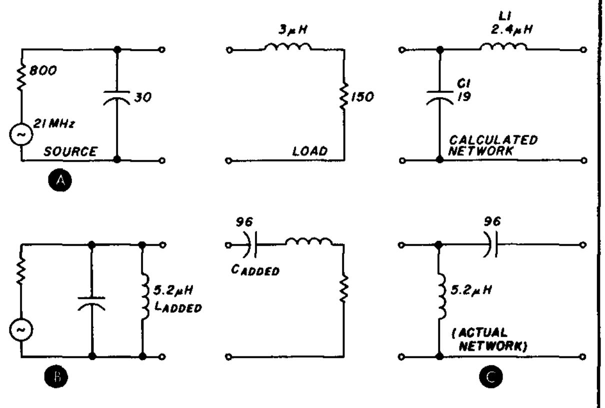

fig. 5. At 15 meters the computed L network (A) reactances must be completely absorbed by the source and load impedance. Additional compensation is required (B) to result in a totally new network (C).

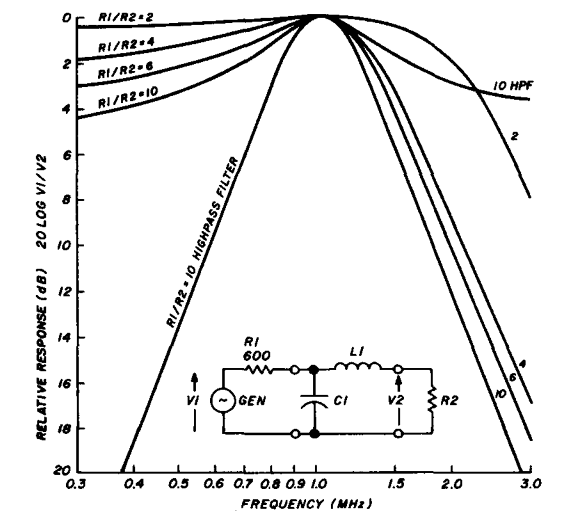

fig. 7. Several L networks were measured using various matching ratios.

modify the component values of the L network to take into account the reactive parts of source and load. Here's an example using network **2A**, because we want to suppress harmonics.

40-meter-band example. Decide whether the sources resistance is higher or lower than the load resistance. The higher resistance is labelled $R1$. Let's match a 40-meter final amplifier, which has a plate resistance of 800 ohms with a shunt capacitance of 30 pF, to a 150-ohm load with a series inductance of 3 μH, **fig. 4A**.

$$R1 = 800\,ohms, R2 = 150\,ohms, f = 7\,MHz$$

Use the equations for network **2A**, which states:

$$X_{L1} = R2\sqrt{(R1/R2)-1} = 312\,ohms$$
$$L1 = X_{L1}/2\pi f = 7.1\,\mu H$$
$$X_{C1} = R1/\sqrt{(R1/R2)-1} = 384\,ohms$$
$$C1 = 1/2\pi f X_{C1} = 59\,pF$$

If this network, **fig. 4B**, is inserted between the two resistances, we would have matched the resistances at 7 MHz. Note, however, that we must also compensate for the source shunt capacitance and the load series inductance. First, capacitance $C1$ of the matching network is in parallel with the source shunt capacitance. Therefore, $C1$ must be equal to 29 pF. The remaining 30 pF required for matching is already present in the source shunt capacitance. The same thing can be done with the load impedance. Instead of adding a 7.1-μH inductor as calculated, use the 3-μH inductance present in the load and add 4.1 μH to make up the difference, **fig. 4C**.

15-meter-band example. The previous example was easy in that the source and load impedances turned out to be directly absorbed into the matching network. What if we wanted to match the same load and source on a different band, say 15 meters?

Using the same equations at the new frequency, **fig. 5A**, $f = 21\,MHz$, $C1 = 19\,pF$, $L1 = 2.4\,\mu H$, the value of $C1$ will be totally absorbed into the source shunt capacitance with 11 pF left over, which must be cancelled. An inductor could be inserted across the source shunt capacitance to cancel only 11 pF of capacitance, **fig. 5B**. The inductive reactance of the added coil must equal the capacitive reactance of the excess shunt capacitance:

$$\begin{aligned} X_{L(added)} &= X_{C(excess)} = 1/(2\pi f_{excess}) \\ &= 1/[2\pi f(30pF - 19\,pF)] \\ &= 1/(6.28)(21\,MHz)(11\,pF) \\ &= 688\,ohms \end{aligned}$$

$$\begin{aligned} L\,added &= [X_{L(added)}/2\pi f] \\ &= 688/2\pi f \\ &= 5.2\,\mu H \end{aligned}$$

The amazing thing is that, when we build the matching L network, instead of a shunt capacitance, $C1$, we'll actually have a shunt inductance, L_{added}, **fig. 5C**.

Now for matching the load impedance. The load already has 3 μH of inductance, which is greater than the 2.4-μH series inductance required, leaving an excess of 0.6 μH to be cancelled. A series-capacitive reactance, just large enough to cancel the 0.6 μH of excess inductance, is needed (**fig. 5B**).

$$X_L(excess) = 2\pi f L(excess) = 2\pi f(0.6\,\mu H) = 79\,ohms$$
$$C(added) = 1/2\pi f X_L(excess) = 96\,pF$$

The L network, **fig. 5C**, looks more like the high-

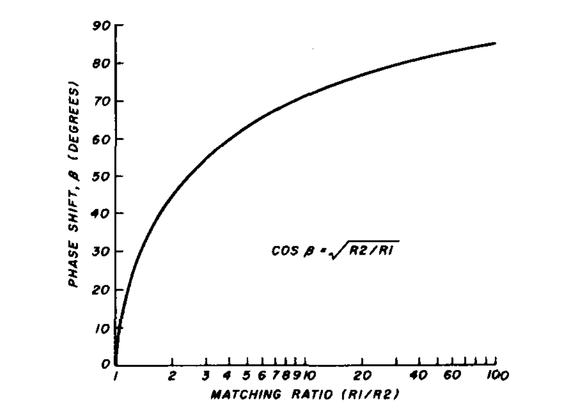

fig. 8. Phase shift of the L network depends also on the matching ratio, ($R1/R2$).

pass filter shown in **fig. 2B** than what we started out to make. Our new filter merely subtracts enough reactance so that the inherent reactance remaining in the source and load perform the actual matching. The total effect is, however, basically that of a lowpass filter.

The only problem now remaining in applying the L network to any impedance is that source and load might not be in the right form. What is needed is for the impedance on the shunt side of the L network to be described in a parallel or shunt form, and that on the series side to be described in a series configuration. If this is not the case, the impedances must be converted using the series-parallel conversion rules discussed earlier. Once this is done source and load reactances can be easily cancelled.

bandwidth

The Q or quality factor of the matching network determines the bandwidth between the upper (f_1) and lower (f_2) 3-dB frequencies.

$$Q = \frac{f_0}{f_1 - f_2} = \frac{operating\ frequency}{bandwidth} \quad \textbf{(18)}$$

The Q for an L matching network is:

$Q = \sqrt{(R1/R2) - 1}$ as shown in **fig. 6**.

Thus Q, or selectivity, is determined by the ratio of the impedances to be matched and cannot be selected independently, as in the more complex π and T matching networks. As the matching ratio increases, so does the circuit Q. The efficiency of an L network is usually greater than 95 per cent:

$$efficiency = [R2/(R2 + R_{coil})]\,[100] \quad \textbf{(19)}$$

some experiments

I put a few L networks together and measured them to see what they could do, **fig. 7**. I decided to match a 600-ohm generator to various loads at 1 MHz. How the Q affects attenuation at the second and third harmonics is apparent. For a matching ratio greater than 4 (Q of 1.7), the slope of the attenuation quickly approaches 12-dB/octave. Every time the frequency is doubled, attenuation increases by 12 dB. The series inductor contributes 6-dB/octave, while the shunt capacitor completes the additional 6 dB. Thus you can expect about 12 dB of attenuation at the second harmonic and 19 dB at the third. I also compared the response of an L network for a matching ratio of 10:1 using either the highpass or lowpass version of the L network. They appear symmetrical about the center frequency.

The ability of an L network to match impedances is also limited by the characteristics of the components. As the frequency is sufficiently increased, the series inductance in the leads of a capacitor predominate, making it look more like an inductor than a capacitor. The same thing happens to inductors at high frequencies, with the capacitance between coil windings tending to shunt the effects of inductance.

The phase shift, β, for an L network as shown in **fig. 8A** is:

$$\cos\beta = \sqrt{R2/R1} \quad \textbf{(20)}$$

when β is measured in degrees, as shown in **fig. 8**. The lagging phase shift approaches 90 per cent as the matching ratio exceeds 10:1. As the matching ratio is increased, the necessary shunt capacitance, $C1$, and series inductance, L1, decrease for a constant input impedance. Therefore, as the matching ratio increases, the network behaves more like a capacitor. This phase delay must be taken into account when matching into different elements of a phased array antenna.

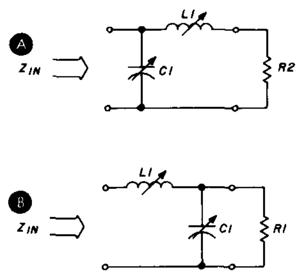

fig. 9. Input impedance varies as the matching elements are adjusted. In *A* a high impedance is matched to a lower impedance. *B* shows matching to a higher impedance.

Finally, let's take a brief look at the input impedance of an L network as series inductor, $L1$, and shunt capacitor, $C1$, are varied. For the case of matching a high impedance downward, **fig. 9A**, note that the real part of the transformation is controlled mostly by varying series inductor $L1$. Remember that matching circuits have been designed for the resistance ratio, then modified slightly to remove reactances. Increasing $L1$ increases the real part of the input impedance. The value of shunt capacitance $C1$ controls the reactive part of the impedance presented to the input. Increasing capacitance increases the $-j$ term.

For the case of matching to a higher impedance, **fig. 9B**, decreasing $C1$ causes the real part of the input impedance to increase. Increasing $L1$ makes the input impedance appear more inductive.

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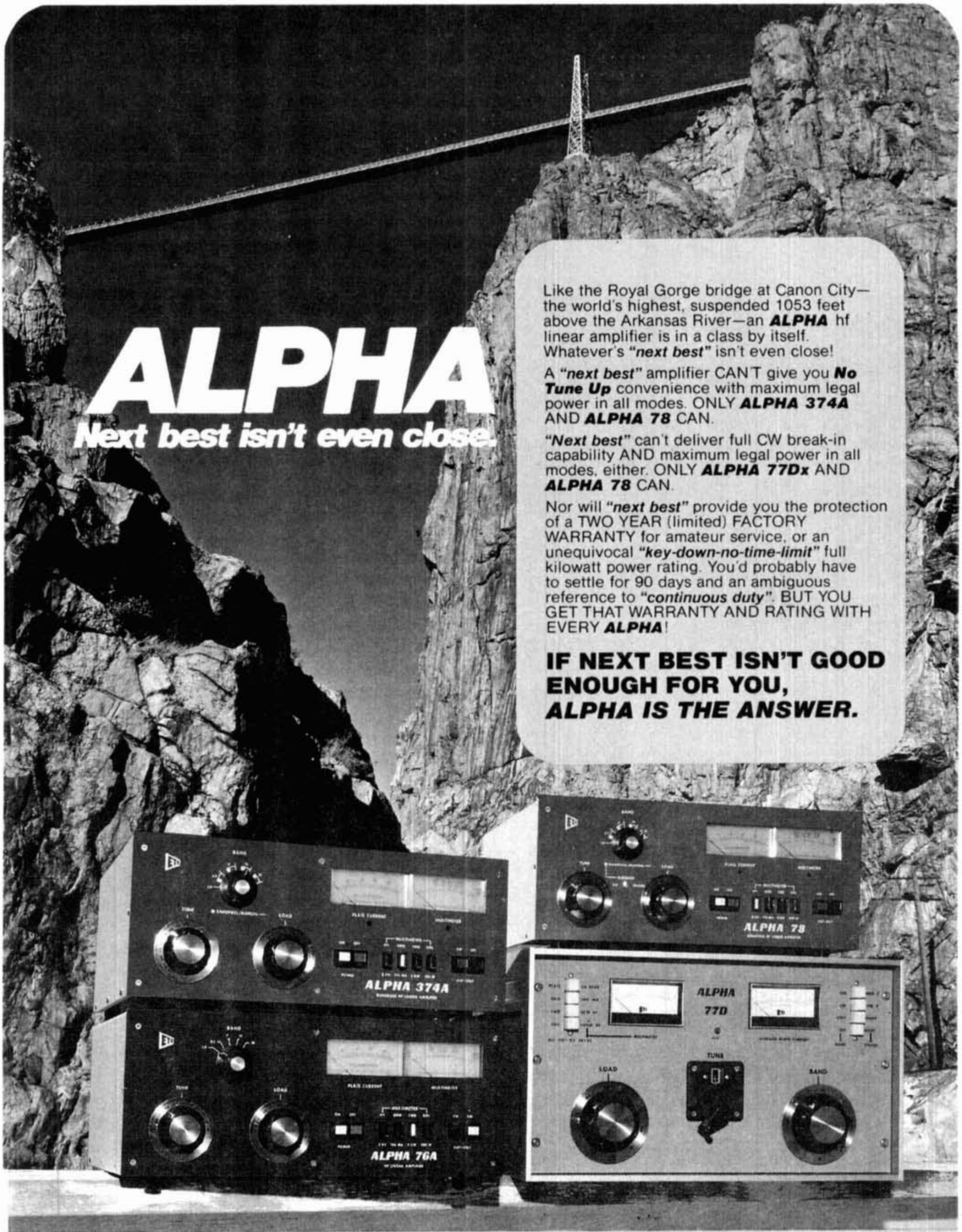

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the half-wave balun: theory and application

Some notes on this versatile tool for impedance matching

The half-wave balun, in which the inverted portion of the balanced output signal is obtained by using a half-wave section of transmission line, is perhaps the simplest and most economical balun available; because of this simplicity and low cost, it's one of the most popular. Its design is straightforward — so simple, in fact, that equations are hardly necessary. Just measure a half wavelength of transmission line and you have a balun.

Despite the popularity of this type of balun, equations that predict its *performance* are generally unavailable. The only previous work on this type of balun of which I am aware was done by Woodward.[1,2]

The usual schematic of the half-wave balun is given in **fig. 1**. For analysis it's convenient to redraw **fig. 1** as shown in **fig. 2**.

The balanced load is assumed to be perfectly balanced; admittedly this isn't always true in practice, but it would not be fair to blame the balun for imperfections caused by the load. The impedance seen by the generator at the unbalanced port is $Z_L/2$ in parallel with $Z_L/2$, or $Z_L/4$, as expected.

In most baluns, the two most important characteristics are the impedance match and the degree of balance, both versus frequency.

impedance balance

One desirable characteristic of a balun is that it presents an essentially constant impedance over a wide bandwidth. Since the half-wave balun will be exactly one-half wavelength at only one frequency, it's useful to examine the impedance presented to the generator in **fig. 2**. The input impedance in **fig. 2** is composed of one-half the load impedance, Z_1, in parallel with the input impedance of the half-wave balancing line. This line, in turn, is terminated in the other half of the load impedance, Z_2. The impedances Z_1 and Z_2 will both be assumed to be independent of frequency. The input impedance of the balancing line, however, will depend on the characteristic impedance and the electrical length of the line, and hence on frequency.

The impedance presented at the input of the half-wave balancing line is given by the familiar transmission line equation:

$$Z_{in} = Z_c \left[\frac{Z_2 \cos\theta + jZ_c \sin\theta}{Z_c \cos\theta + jZ_2 \sin\theta} \right] \qquad (1)$$

where Z_{in} = input impedance of half-wave balancing line
Z_c = characteristic impedance of balancing line
Z_2 = load impedance of the balun; equal to one-half balanced load

After some arithmetical manipulation, which I'll explain in the appendix, and setting $k = Z_c/Z_2$, we obtain the ratio of input impedance to load impedance for the resistive (real) component:

$$\frac{Z_{in}}{Z_c} = \frac{k^2}{(k \cos\theta)^2 + (\sin\theta)^2} \qquad (2)$$

Eq. 2 is plotted for various values of k over a range of θ from 90 to 270 degrees; see **fig. 3**. This represents a frequency range from one-half to one and one-half times the design frequency. The ordinate gives the input impedance in terms of the load impedance.

As shown, the ordinate is matched $(Z_{in} = Z_c)$ for all values of characteristic impedance only when $k = 1$, i.e., when the characteristic impedance equals the load impedance. If you stop and think about it, both these results should have been expected.

For values of $\theta = 90$ and 270 *degrees*, the balanc-

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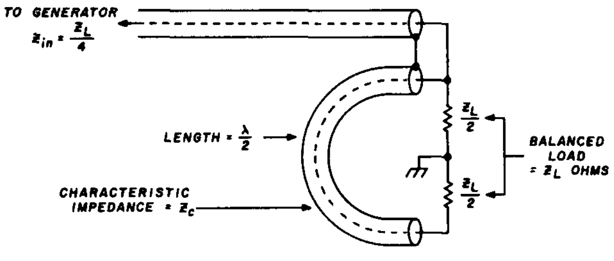

fig. 1. Basic half-wave balun circuit using lengths of transmission line.

ing line is an odd number of quarter-wavelengths; thus the balancing line inverts the load impedance about the characteristic impedance squared. For these line lengths, the ratio Z_{in}/Z_2 is equal to k^2.

Fig. 4 shows the reactive (imaginary) component of balancing line input impedance. In this case the reactance is zero for all frequencies only when $k = 1$, again the matched case.

The balun input impedance is, of course, the impedance given from **figs. 3** and **4** in parallel with $Z_1 = (Z_L/2)$. (This latter impedance has been assumed to be independent of frequency, but in the real world it may also vary with frequency.)

balance bandwidth

In addition to the impedance bandwidth, it's desirable to know balance bandwidth. Balance bandwidth means that the frequency range over which the currents through the two halves of the balanced load are equal in magnitude and opposite in phase with a specified tolerance. Following Woodward,[1,2] we use as the criteria of balance:

$$balance = \left| \frac{I_1 - I_2}{I_1 + I_2} \right| \quad (3)$$

From **fig. 2** current I_1 is given by $\frac{e_0}{Z_1}$. Current I_2 is given in terms of the load resistance and characteristic impedance by the well-known transmission line equation[3]

$$e_0 = e_2 \cos\theta - jI_2 Z_c \sin\theta \quad (4)$$

After arithmetical manipulations as shown in the appendix, we obtain:

$$I_2 = \frac{e_0}{Z_2 (\cos\theta + jk \sin\theta)} \quad (5)$$

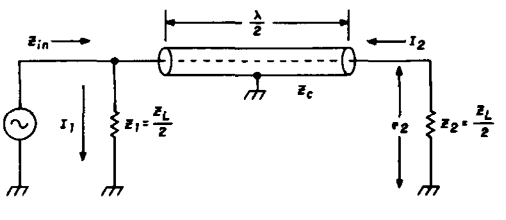

fig. 2. Equivalent of half-wave balun shown in *fig. 1*.

from which $$\left| \frac{I_1 - I_2}{I_1 + I_2} \right| = \left[\frac{(\cos\theta + 1) + jk \sin\theta}{(\cos\theta - 1) + jk \sin\theta} \right] \quad (6)$$

where $k = \frac{Z_c}{Z_2}$ as before.

After rationalizing, separating into real and imaginary parts, and finding the magnitude (described in the appendix), a plot of this equation is given in **fig. 5**; the system is balanced when the plot equals zero. As can be seen, a perfect balance occurs only when $\theta = 180$ degrees as might be expected. The balance bandwidth improves for small values of $k = Z_c/Z_2$. Thus, for best balance bandwidth, the load impedance should be large compared with the balancing-line characteristic impedance. This is contrary to impedance bandwidth considerations, in which the widest impedance bandwidth is obtained for the matched case.

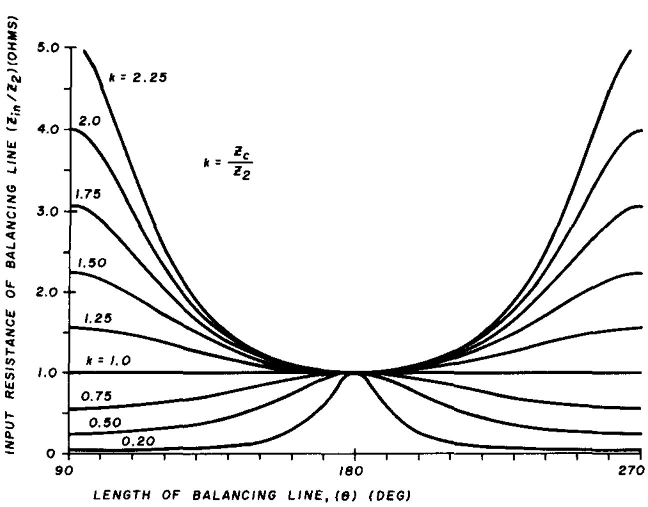

fig. 3. Input resistance of the half-wave balancing line as a function of line length for various values of characteristic impedance.

Increasing the load impedance to values greater than about ten times the characteristic impedance $(k \leq 0.1)$ gives little additional improvement in balance bandwidth. For small values of k, **eq. 6** approaches $(\cos\theta + 1)/(\cos\theta - 1)$, which is independent of k.

phase shift

One final factor to consider is the change in phase with frequency caused by propagation delay through the balancing line. At the design frequency, the relative phase at the output of the line is 180 degrees with respect to the input. This phase shift will vary directly with frequency. Assuming that the amplitude (current or voltage) is constant with frequency, the phase change will cause reduction in the in-phase component (referenced to 180-degree phase-shift

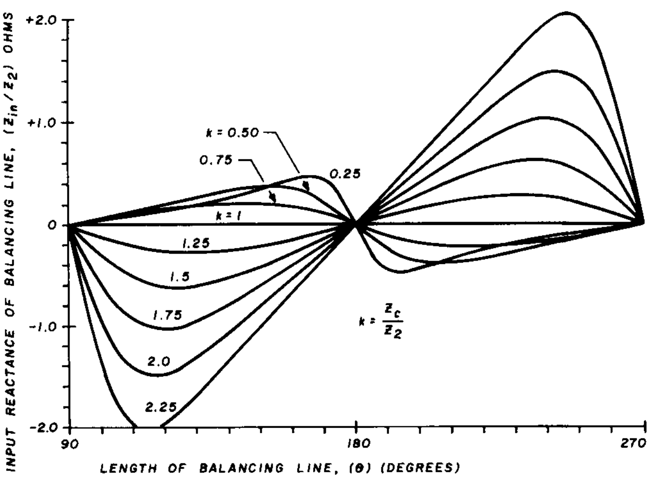

fig. 4. Input reactance of the half-wave balun as a function of line length for various values of characteristic impedance.

from the balancing-cable input) and the generation of a quadrature component.

The in-phase and quadrature components will be proportional to

$$\cos\theta \text{ (in-phase) and } \sin\theta \text{ (quadrature)} \qquad (7)$$

respectively, where θ is the deviation from 180 degrees with respect to the input. From **eq. 7**, it can be seen that a deviation of 26 degrees (14.5 per cent) of design frequency will cause a 1-dB reduction in output amplitude through one-half of the balanced load caused by phase change. Similarly, a 6.24 degree deviation (3.4 per cent of design frequency) will cause a 1-dB quadrature component to be generated.

The effect of the reduction in the in-phase component is an unbalance through one-half the load. The effect of the quadrature component will depend upon the balanced-load characteristics. If the balanced load is an antenna, two possibilities are a) generation of unexpected side lobes, or b) generation of a circular component to the antenna response.

In other situations the quadrature component may or may not be important. The effect of the quadrature component should be considered for each application.

example

A typical application of the half-wave balun is to couple a balanced folded dipole with a nominal impedance of 300 ohms to a 75-ohm coaxial cable. This represents a 4:1 impedance transformation. For convenience, the same 75-ohm coax is generally also used for the balancing section. In this case, $Z_2 = Z_1 = 150$ *ohms* equals one-half the balanced load, so that with 75-ohm coax, $k = 75/100 = 0.5$. From the $k = 0.5$ curve, it can be seen that the frequency range is not particularly good — only about ±25 degrees or about 12 per cent for an impedance change to the 70 per cent point. Use of higher-impedance coax for the balancing section would give a wider bandwidth.

It can also be seen that the half-wave balun will give better performance when used to couple a circuit to a higher impedance balanced load, such as a folded dipole, rather than to a lower impedance antenna, such as a Yagi antenna.

conclusions

I have discussed the performance of the half-wave coaxial balun. As often occurs in engineering problems, the performance characteristics of a device can have very subtle implications, even when the device itself is extremely simple. It has been shown that the best impedance characteristic is obtained when the impedance of the balancing line is equal to one-half the balanced load impedance. On the other hand, the best balance versus frequency characteristic is obtained when the balanced load impedance is high compared with the balancing line impedance. There is no requirement that the characteristic impedance of the transmission line feeding the balun/load be the same as that of the balancing half-wave section.

Whether or not the bandwidth limitations of the half-wave balun are the limiting factor in overall system performance depends on other factors. In particular, if the balanced load is an antenna, as it often is, its performance may deteriorate faster than that of the balun so that it may be the limiting factor.

Despite the bandwidth limitations described, I believe the half-wave balun will continue to be widely used because of its simplicity and economy.

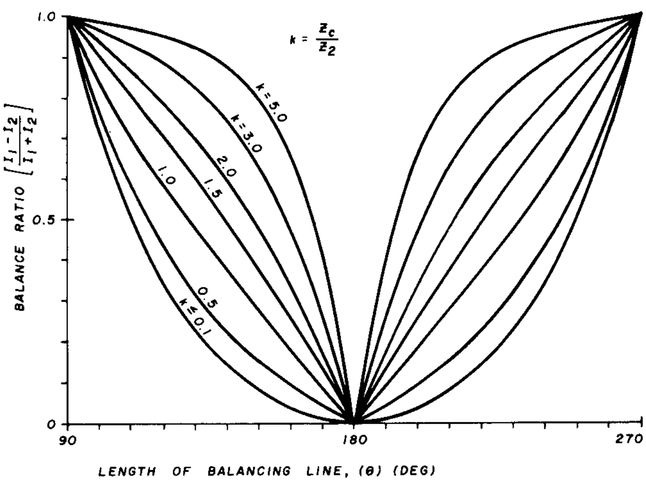

fig. 5. Balance ratio as a function of balancing-line electrical length. Perfect balance is obtained when the balance ratio is zero.

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1. O.M. Woodward, Jr., "Balance Measurements on Balun Transformers," *RCA License Bulletin LB-872*, July 22, 1952. RCA Laboratories Division, Princeton, New Jersey.
2. O.M. Woodward, Jr., "Balance Measurements on Balun Transformers," *Electronics*, Vol. 26, No. 9, September, 1953, pages 188-191.
3. Terman, *Radio Engineering*, 3rd Edition, McGraw-Hill Book Company, Inc., 1947, figure 4-1, page 75, and Eq. (4-32a), page 92.

appendix

Because many of *ham radio*'s readers are mathematically inclined, I'll present a brief derivation of the more important equations used in this article.

Beginning with the basic transmission line equation, **eq. 1**, divide the numerator and denominator of the right-hand side by Z_2; also divide both sides of the equation by Z_2. This gives:

$$\frac{Z_{in}}{Z_2} = \frac{Z_c}{Z_2}\left[\frac{\cos\theta + j\frac{Z_c}{Z_2}\sin\theta}{\frac{Z_c}{Z_2}\cos\theta + j\sin\theta}\right] \qquad \textbf{(A-1)}$$

$$= k\left[\frac{\cos\theta + jk\sin\theta}{k\cos\theta + j\sin\theta}\right] \qquad \textbf{(A-2)}$$

Since we defined $\frac{Z_c}{Z_2} = k$

Now rationalize **eq. A-2**.

$$\frac{Z_{in}}{Z_2} = k\left[\frac{\cos\theta + jk\sin\theta}{k\cos\theta + j\sin\theta}\right]\left[\frac{k\cos\theta - j\sin\theta}{k\cos\theta - j\sin\theta}\right] \qquad \textbf{(A-3)}$$

Separate **eq. A-3** *into real and imaginary parts:*

$$\frac{Z_{in}}{Z_2}(real) = k\left[\frac{k\cos^2\theta + k\sin^2\theta}{(k\cos\theta)^2 + (\sin\theta)^2}\right]$$

$$= \frac{k^2}{(k\cos\theta)^2 + (\sin\theta)^2} \qquad \textbf{(A-4)}$$

which is plotted in **fig. 3**.

Note that when $k = 1$, the matched case, $\frac{Z_{in}}{Z_2}$ (real) is independent of θ and equal to unity.

For the imaginary part,

$$\frac{Z_{in}}{Z_2}(imag.) = k\left[\frac{k^2\cos\theta\sin\theta - \cos\theta\sin\theta}{(k\cos\theta)^2 + (\sin\theta)^2}\right]$$

Factoring,

$$\frac{Z_{in}}{Z_2}(imag.) = k\left[\frac{(k^2 - 1)\cos\theta\sin\theta}{(k\cos\theta)^2 + (\sin\theta)^2}\right] \qquad \textbf{(A-5)}$$

which is plotted in **fig. 4**.

In this case, when $k = 1$, the imag. part of $\frac{Z_{in}}{Z_2}$, which represents the reactive component, is zero regardless of θ.

To derive the balance equation, **eq. 6**, refer to **fig. A-1**. From **fig. A-1**, the well-known transmission line equation can be obtained:[3]

$$E = E_r\cos\theta + jI_r Z_0\sin\theta \qquad \textbf{(A-6)}$$

Eq. A-6 gives voltage at the input of the transmission line in terms of the load current and characteristic impedance of the line.

As $E_r = I_r Z_r$, **eq. A-6** can be rewritten as

$$E = I_r Z_r\cos\theta + jI_r Z_0\sin\theta \qquad \textbf{(A-7)}$$

Solving for I_r,

$$I_r = \frac{E}{Z_r\cos\theta + jZ_0\sin\theta} \qquad \textbf{(A-8)}$$

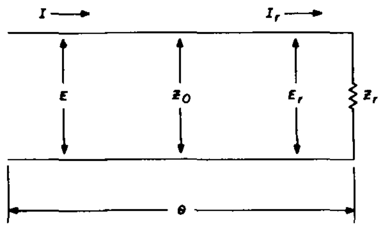

fig. A-1. Basic transmission-line circuit.

Rewrite **eq. A-8** using the nomenclature of **fig. 2**:

$$I_2 = \frac{-e_0}{Z_2\cos\theta + jZ_c\sin\theta} \qquad \textbf{(A-9)}$$

The minus sign is obtained since I_2 of **fig. 2** flows in the opposite direction from I_r.

Since $k = \frac{Z_c}{Z_2}$,

$$I_2 = \frac{-e_0}{Z_2(\cos\theta + jk\sin\theta)}$$

which is **eq. 5** in text.

Current I_1 in **fig. 2** is given by $I_1 = \frac{e_0}{Z_1}$. But since $Z_1 = Z_2$, we can write $I_1 = \frac{e_0}{Z_2}$. Hence the ratio $\left|\frac{I_1 - I_2}{I_1 + I_2}\right|$ is given by

$$|ratio| = \left|\frac{\frac{e_0}{Z_2} - \frac{-e_0}{Z_2(\cos\theta + jk\sin\theta)}}{\frac{e_0}{Z_2} + \frac{-e_0}{Z_2(\cos\theta + jk\sin\theta)}}\right|$$

$$|ratio| = \left|\frac{(\cos\theta + 1) + jk\sin\theta}{(\cos\theta - 1) + jk\sin\theta}\right| \qquad \textbf{(A-10)}$$

Rationalize **eq. A-10** and obtain

$$|ratio| = \frac{[(\cos\theta + 1) + jk\sin\theta]\,[(\cos\theta - 1) - jk\sin\theta]}{(\cos\theta - 1)^2 + (k\sin\theta)^2}$$

from which the real and imaginary parts are given by

$$real = \frac{\cos^2\theta - 1 + (k\sin\theta)^2}{(\cos\theta - 1)^2 + (k\sin\theta)^2}$$

$$imaginary = \frac{-2k\sin\theta}{(\cos\theta - 1)^2 + (k\sin\theta)^2}$$

The magnitude is given by the square root of the sum of the real and imaginary parts squared, or

$$magnitude = \frac{\sqrt{[\cos^2\theta - 1 + k^2\sin^2\theta]^2 + [-2k\sin\theta]^2}}{(\cos\theta - 1)^2 + (k\sin\theta)^2} \qquad \textbf{(A-11)}$$

This equation is plotted in **fig. 5**. It can be seen from **eq. A-11** that when $k \leq 0.1$, the magnitude approaches

$$magnitude = \left|\frac{\cos\theta + 1}{\cos\theta - 1}\right|$$

and is independent of k.

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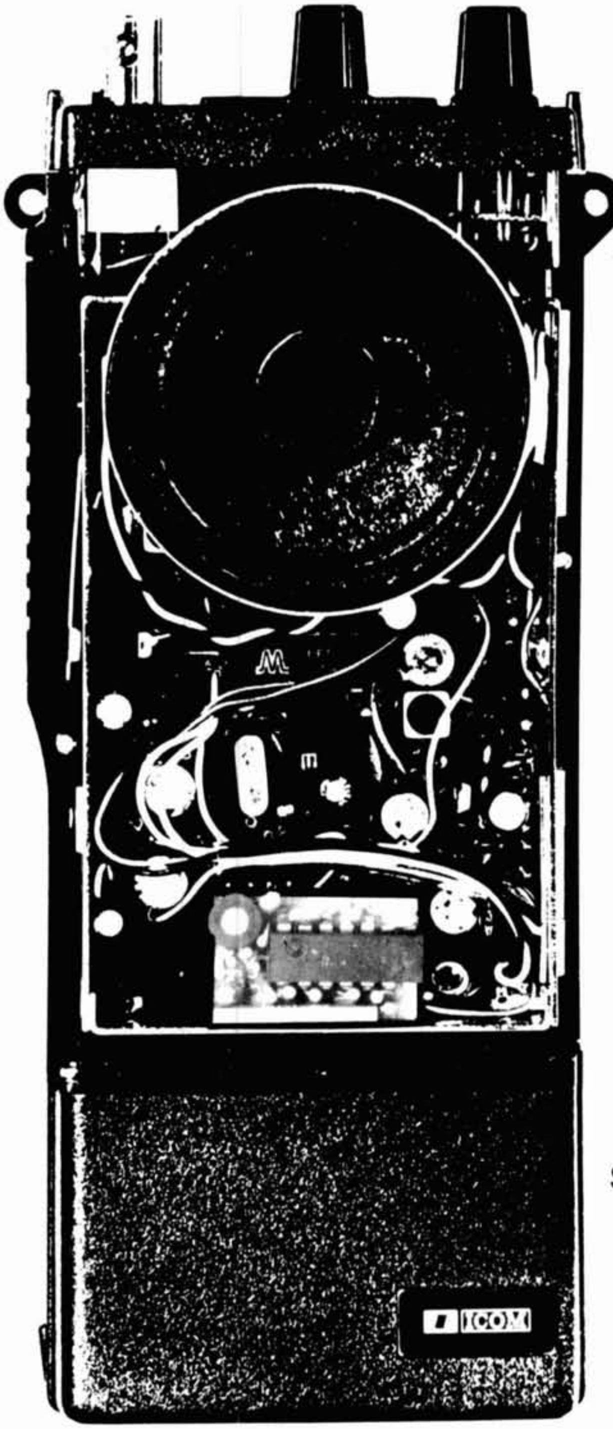

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More Details? CHECK—OFF Page 94

Yagi antenna design: quads and quagis

A discussion of the one wavelength in circumference loop including examination of gain and loop shape

Up to this point in this series on Yagi antennas, we have considered only linear cylindrical elements about one-half wavelength long with a small diameter compared to wavelength; there is in common use, however, a radiating "element" consisting of a loop of wire about one wavelength in circumference. There are many such loop configurations: a triangular loop (commonly known as a delta loop), a square loop with two sides parallel to the earth (known as the quad), and a square loop oriented so that two diagonal corners are perpendicular to the earth (known as a diamond loop). Loops can also be made that are not equilateral. Triangles can be isosceles or even with three different sides; four-sided loops can be rectangles, and, indeed, loops can have more than four sides or can even be round.

The one-wavelength loop can be used either as a driven radiator or a parasite. To drive the loop it is opened at some point on its circumference where it is excited with a suitable voltage or current. As a parasite the loop is left closed; current will flow depending upon the induced voltages from other loops or elements and the self-impedance of the loop. To help understand the behavior of these loops it is useful to model the loop in terms of dipole elements.

Let's first consider a model for a square quad driven loop. The model starts with two driven half-wavelength dipoles in free space as shown in **fig. 1**. The dipoles are voltage driven at their centers (at xx and yy) and in this model are separated (vertically) by a quarter wavelength ($\lambda/4$). If the excitation is equal at xx and yy (the same voltage and phase), the dipole currents will also be equal and, therefore, the voltages at the ends of the dipoles will be equal. The outer $\lambda/8$ sections of each dipole may now be bent as shown by the dotted line, forming a square. Since the voltages at dipole ends are still equal, the bent dipoles can be connected together without changing any currents in the square.

Since for each dipole there is a unique relationship between voltage at its ends to current at its center, the end connection of both dipoles in the square configuration insures that the center dipole currents will be equal even though excitation voltages at xx and yy may be different! Indeed, we can remove voltage excitation at yy (shorting the terminals) and still be sure that whatever dipole currents flow from excitation at xx, they will be equal in the two dipoles. One can look at the excitation in this case as a dipole *current* feed at xx; the dipole centered at yy is *voltage* fed at its ends. Note that to realize the same loop current the sum of the voltages supplied to xx and yy must be constant. This leads to the well-known fact that the driving point impedance at xx and yy shorted will be just twice the impedance at xx when yy and xx are excited equally.

Fig. 2 shows the square quad excited at the center of the bottom section and also shows the relative magnitudes and directions of the currents which flow. Note that the horizontal sections of the square show currents in the same direction as those of the original dipoles of **fig. 1**; therefore these sections will provide radiation fields at long distances; since they are shorter than the original half-wavelength dipoles,

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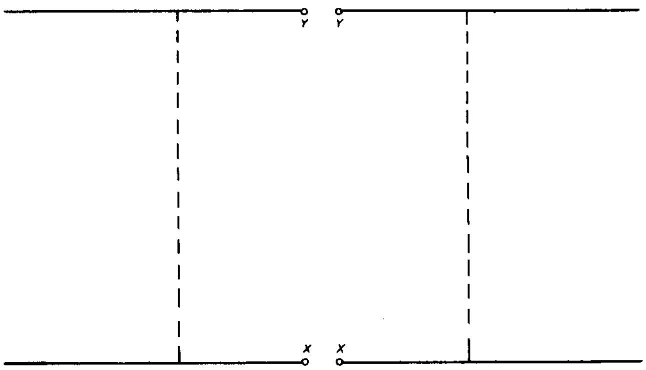

fig. 1. Diagram of two driven dipoles with λ/4 vertical spacing. Bending the ends of the dipoles together, as shown by the dotted lines, forms the square quad loop.

however, they require more current to produce a given radiation field, *i.e.*, the radiation resistance will drop. Moreover, the directivity and gain of each segment will be somewhat smaller than that of a full half-wavelength dipole. Both of these effects will be discussed shortly.

The vertical portions of the square quad shown in **fig. 2** are quite different. Note that for each of these vertical segments the top and bottom portions have identical currents but in opposite directions. This symmetry insures complete cancellation of the far (radiation) field. In other words, the vertical sections do not radiate; they simply act as a capacitive "top hat" loading (or tuning) for the horizontal radiating segments. Thus the radiating properties of the quad square of **fig. 2** will be identical to the (capacity loaded) resonant shortened or truncated horizontal segments alone. These will, of course, produce only horizontally polarized radiation.

Take another look at the two half-wavelength dipoles of **fig. 1**. The two (broadside) dipoles separated by a distance of $\lambda/4$ will produce a gain increase over one dipole. This separation or *stacking* gain can be easily calculated by using known self and mutual impedances; the result is shown in **table 1** and also in **fig. 3** where the overall gain in dBi is plotted against the separation, S, in units of wavelength, λ. Note that the separation gain improvement peaks when S is about $5\lambda/8$, leading to the often-quoted (but generally incorrect for other than single elements) "optimum stacking separation."

Fig. 3 can be understood qualitatively by remembering that the effective capture cross section area for a single half-wavelength dipole is about $0.13\lambda^2$. Thus, when the separation between the dipoles is large enough, the capture areas are basically independent (and thus additive), leading to an effective gain improvement of a factor of 2 or 3 dB. For smaller separations, however, the capture areas overlap. In this overlap region both constructive and destructive interference can occur; for very small separations the combined capture areas reduce to that of a single dipole but for some intermediate regions (such as $S = 5\lambda/8$) constructive interference occurs, making the gain improvement more than a factor of 2. This constructive interference is due to a favorable reduction in the driving point resistance of the dipole, which is a direct result of the behavior of the real part of the mutual impedance of the two dipoles.

Another qualitative way of understanding this entire phenomenon is to view the (transmitting) dipoles as an excitation of a vertical aperture. Broadening this aperture by separating the dipoles is tantamount to narrowing the H-plane pattern, which will increase the gain. When the dipole separation becomes large enough, however, the quasi uniform illumination disappears and the vertical aperture acts like two independent (small) apertures giving rise to a diffraction pattern in the H-plane with maxima corresponding to a gain improvement of exactly a factor of 2. In any case, the actual calculated values of gain are shown in **table 1**. Note that for the quarter-wavelength separation of **fig. 1** the gain of the stacked dipoles is 3.236 dBi, or just 1.085 dB above that for a single half-wavelength dipole. This gain increase is all due to *beam* pattern narrowing in the H-plane; the E-plane pattern beamwidth remains the same as that for a single half-wavelength dipole.

The square quad loop of **fig. 2** is very similar to the stacked dipoles of **fig. 1**, but there are two significant changes. Because of the truncated or shortened elements, the driving-point impedances of the elements are reduced and the E-plane pattern width is somewhat increased, resulting in a somewhat reduced gain. To calculate these two results I will use the same method outlined by Kraus[1] (page 139-143).

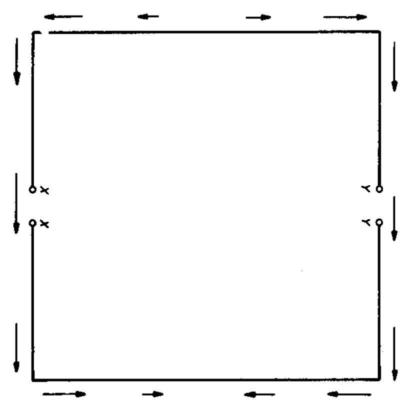

fig. 2. Outline of the square quad loop showing the current distribution in the horizontal and vertical members. Horizontal polarization results with the feedpoints as shown.

Kraus's calculation applies to a thin radiating element which is not capacitance loaded; to make the calculation apply to a single truncated element fed at the maximum current point, the limits of integration must be suitably altered; **fig. 4** shows the essential geometry.

The retarded value of the current at point z on the antenna referred to a point at distance, S, is

$$I = I_o \left[\cos \frac{2\pi}{\lambda} \cdot z\right] e^{jw(t - \frac{S}{c})} \qquad (1)$$

where the quantity in brackets is the form factor for the current on the antenna. Following the Kraus development, the antenna can be viewed as a string of infinitesimal dipoles of length dz. The far fields, dE_θ and dH_ϕ, at a distance, S, from an infinitesimal dipole, dz are

$$dH_\phi = j \frac{I \sin \theta \, dz}{S\lambda} \qquad (2)$$

$$dE_\theta = 120 \cdot \pi \cdot dH_\phi \qquad (3)$$

where ϕ is the azimuth angle around the z axis. The total field from the entire antenna is then

$$H_\phi = \int_{-L/2}^{+L/2} dH_\phi \qquad (4)$$

From **eq. 1, eq. 2,** and **eq. 4,** (5)

$$H_\phi = \frac{j I_o \sin \theta \, e^{jwt}}{2\lambda} \int_{-L/2}^{+L/2} \frac{1}{S} \cos \frac{2\pi}{\lambda} \cdot z \cdot e^{-j\frac{wS}{c}} dz$$

Note that $S = r - z \cos \theta$ (6)
and also note that at long distances the amplitude of S is the same as the amplitude of r, so that we may write:

$$H_\phi = \frac{j I_o \sin \theta \cdot e^{j\omega(t - \frac{r}{c})}}{2\lambda r} \int_{-L/2}^{+L/2} \cos \frac{2\pi}{\lambda} \cdot z \cdot e^{j \frac{\omega \cos \theta}{c} \cdot z} dz \qquad (7)$$

Let $\beta = \omega/c = 2\pi/\lambda$; **eq. 7** may be rewritten:

$$H_\phi = \frac{j \beta I_o \sin \theta \cdot e^{j\omega(t - \frac{r}{c})}}{4\pi r} \int_{-L/2}^{+L/2} e^{j\beta z \cos \theta} \cos (\beta z) \, dz \qquad (8)$$

Since $\int e^{ax} \cdot \cos (bx) \, dx = \dfrac{e^{ax}[a \cos (bx) + b \sin (bx)]}{a^2 + b^2}$

and if $a = j \beta \cos \theta$ and $b = \beta$, then **eq. 8** becomes:

$$H_\phi = \frac{j \beta I_o \sin \theta \cdot e^{j\omega(t - \frac{r}{c})}}{4\pi r} \qquad (9)$$

$$\left| \frac{e^{j \cos \theta \cdot \beta z}}{\beta^2 (\sin^2 \theta)} (j \beta \cos \theta \cdot \cos \beta z + \beta \cdot \sin \beta z) \right|_{-L/2}^{+L/2}$$

Evaluating this expression at both limits and collecting terms:

$$H_\phi = \frac{j [I_o]}{2\pi r} \cdot F(\theta)$$

where $[I_o] = I_o e^{j\omega(t - \frac{r}{c})}$

and $F(\theta) =$ (10)

$$\left\{ \frac{\cos (\frac{\beta L}{2} \cdot \cos \theta) \sin \frac{\beta L}{2} - \cos \theta \cdot \sin (\frac{\beta L}{2} \cdot \cos \theta) \cos \frac{\beta L}{2}}{\sin \theta} \right\}$$

$F(\theta)$ is often referred to as the field pattern factor. Thus the far fields of the truncated element can be written:

$$H_\phi = \frac{j [I_o]}{2\pi r} \cdot F(\theta) \qquad (11)$$

$$\text{and } E_\theta = \frac{j\, 60 \, [I_o]}{r} \cdot F(\theta) \qquad (12)$$

Note that if there is no truncation $(L = \frac{\lambda}{2})$ **eq. 10** reduces to the well-known expression for a $\lambda/2$ dipole:

$$F(\theta) = \frac{\cos (\frac{\pi}{2} \cdot \cos \theta)}{\sin \theta} \qquad (13)$$

Kraus has shown that the self-radiation resistance, R_{11}, of such a linear element can be computed by equating the integral of the Poynting vector over a large sphere (total power radiated) to the driving point (current maximum) total power supplied. The result is (see Kraus,[1] 5-90, page 143):

$$R_{11} = 60 \int_o^\pi F^2(\theta) \cdot \sin \theta \, d\theta \qquad (14)$$

For the truncated element it is now a simple matter to insert the value of $F(\theta)$ from **eq. 10** into **eq. 14** and integrate. The integration is quite easily done numerically with a simple computer program. The result of such an integration is shown in **table 2** where the radiation resistance of a truncated dipole element of length L (in terms of wavelength λ) is shown.

The directivity or gain can also be easily computed. Once $F(\theta)$ is given, the directivity is simply the ratio of the *maximum* value of $F^2(\theta)$ to the *average* value of $F^2(\theta)$ over the entire 4π solid angle. That is, the directivity D is:

$$D = \frac{F^2 (\theta = \frac{\pi}{2}) \cdot 4\pi}{2\pi \int_o^\pi F^2(\theta) \sin \theta \, d\theta} \qquad (15)$$

or from **eq. 14**

$$D = \frac{120}{R_{11}} F^2\left(\theta = \frac{\pi}{2}\right) = \frac{120}{R_{11}} \cdot \sin^2 \left(\frac{\beta L}{2}\right) \qquad (16)$$

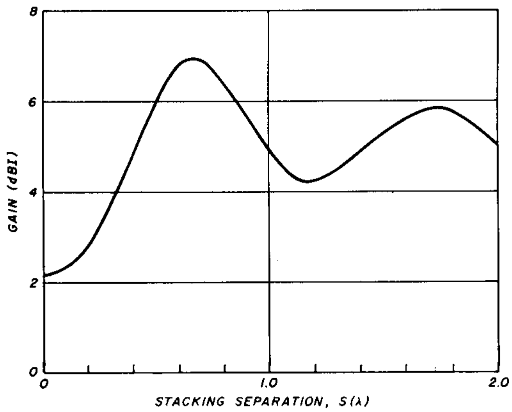

fig. 3. "Stacking" gain *vs* separation for two side-by-side dipoles.

The calculated directivity, D, and the related gain, expressed in dBi, are also listed in **table 2**.

Note that for small truncations where L is only slightly shorter than $\lambda/2$ there is not much reduction in self-radiation resistance and not much reduction in directivity; this is to be expected because the small ends of the dipole which are truncated carry little current, so do not contribute greatly to element performance. For heavy truncation, however, both self-radiation resistance and directivity decrease significantly; the limiting case where the length goes to zero is the well-known infinitesimal dipole whose directivity is just 1.500 and gain is 1.761 dBi.

To compute the driving-point resistance of the full quad square loop it is necessary to know the mutual resistance between two (truncated) elements separated by $\lambda/4$. From quite fundamental considerations Hurwitz[2] has shown a mathematical expression for the real part of the complex mutual impedance between two elements, R_{21}:

$$R_{21} = 60\int_0^{\pi} F^2(\theta)\cdot \sin\theta \cdot J_o\,(\beta S\cdot \sin\theta)\,d\theta \qquad (17)$$

where $F(\theta)$ is the pattern function and J_o is the Bessel function whose argument is the product of element separation S and $\beta\cdot\sin\theta$. If S is measured in units of λ the argument is simply $(2\pi S\cdot \sin\theta)$. Note that for very small separations J_o approaches unity; for very close separations $R_{21} \cong R_{11}$.

Shown in **table 3** are values of mutual resistance vs separation (λ) of truncated elements. Note that the mutual resistance behaves very similarly to the values for full half-wavelength dipoles[3] but the magnitudes are much smaller; a careful comparison shows that the reduction factor varies somewhat with separation.

The properties of the square quad loop can now be computed. **Table 2** shows that the gain of a *single* (truncated) $\lambda/4$ element is 1.922 dBi and that $R_{11} = 38.547$ *ohms*. For the full quad loop of **fig. 2** there are two "elements;" if they are both equally driven, they will produce the same far-field strength and power density as a single driven "element" carrying twice the current. However, the driving-point resistance of each one of the driven elements, when both elements are equally excited, is $R_{11}+R_{12}$. Therefore, total input power (both elements) is:

$$W_2 = 2\cdot I^2(R_{11}+R_{12}) \qquad (18)$$

While for the single element alone (same far field) is:

$$W_1 = (2I)^2\cdot R_{11} \qquad (19)$$

If the directivity of a single element is designated as D_1 and the directivity of the full quad loop as D_2, then

$$D_2 = D_1\,\frac{W_1}{W_2} = 2D_1/(1+R_{12}/R_{11}) \qquad (20)$$

From **tables 2** and **3** values for the square quad loop are: $R_1 = 38.547$ *ohms*; $R_2 = 21.729$ *ohms*; D_2/D_1 becomes 1.279 or 1.069 dB. Note that this *stacking* gain of the two truncated elements is nearly identical to the stacking gain of half-wavelength dipoles at a separation of $\lambda/4$ (1.085 dB, see **table 1**); the difference is due only to the details of mutual resistance. The driving-point resistance, R, of the total loop, of course, is twice the value for a single element (when both are driven):

$$R = 2(R_{11}+R_{12}) = 120.6\ ohms \qquad (21)$$

Since D_1 (see **table 2**) is 1.557,

$$D_2 = 1.991 \text{ and} \qquad (22)$$

$$Gain = 2.992\ dBi \qquad (23)$$

These properties of the square quad loop have been obtained rather rigorously; the main assumption in the model is total neglect of far-field radiation from the vertical sections, which are assumed to act as capacitance sinks for the current at the ends of the horizontal radiating segments. Moreover, the assumption is made that the current distribution on

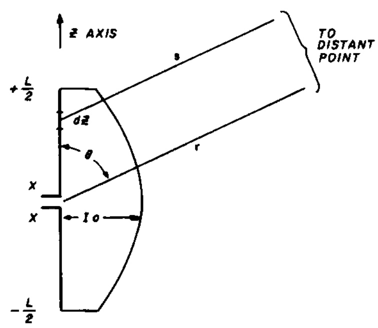

fig. 4. Current relationships for the symmetrical, thin center-fed truncated antenna of length L. This is the relationship necessary to determine driving-point impedance and H- and E-plane patterns of the square quad loop.

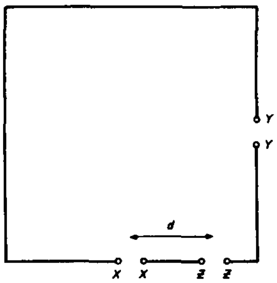

fig. 5. Example of a square loop with the excitation applied at one of three different places.

the horizontal segments is strictly sinusoidal; this is valid for very thin elements. Most quad loops are built with wire, which is thin compared to a wavelength, so one can be quite confident of the model.

It is now easy to understand qualitatively the radiation pattern of the loop. The H-plane profile does narrow, compared with a single linear element, because of the "illumination" of a wider vertical aperture; quantitatively, this profile narrowing is almost the same for truncated quad "elements" and the equivalently separated half-wavelength dipoles. The E-plane profile, however, is not as narrow for the truncated element as for a half-wavelength dipole; this factor accounts for the somewhat reduced directivity (see **table 2**). The loss in gain is about 0.23 dB for the $\lambda/4$ truncated element.

other driven loops

Before I consider a multiloop quad system I will briefly discuss other forms of driven loops. First, I will determine the performance of the square quad loop when it is driven or excited at a point other than the center of the bottom segment. **Fig. 5** shows the square loop with three different feed points: the center of the bottom leg xx, the center of the right vertical leg yy, and at an arbitrary point, zz, placed at a counter-clockwise distance d (from xx) around the loop. It has already been shown that if excitation is applied at xx (with yy and zz shorted), the loop will produce a totally horizontally polarized far field; its gain is 2.992 dBi. Similarly, if excitation is applied at yy (with xx and zz shorted), only vertically polarized far-field radiation will occur; gain is again just 2.992 dBi. Note that excitation at xx and yy are basically independent, i.e., unit current flow due to excitation at xx has a null current at yy and vice versa.

If we now excite the loop at zz with the same unit current, $I_0 \cos \omega t$, it is easy to see that current flow at xx is just $I_0 \cos (\beta d) \cos \omega t$ and at yy is $I_0 \sin (\beta d) \cos \omega t$. These currents will produce orthogonal far fields which must be added vectorially to obtain the total far field; the total far field has exactly the same magnitude as that which is produced by the same unit current at xx alone. In other words, the square loop gain and drive-point resistance are totally independent of the feed point; only the polarization changes (from totally horizontal if the feed is at xx to totally vertical if the feed is at yy). This simple theorem can be easily proved by the same type of argument for *any* equilateral one-wavelength loop.

It is now interesting to consider moving the feedpoint zz to the lower right-hand corner of the square loop. Since we are considering free space loop gain for which rotation is unimportant, it is clear that this configuration produces *exactly* the same result as the familiar square *diamond* loop fed at the bottom or top corner. Thus we now know that the quad square and the diamond square have *exactly* the same gain and *exactly* the same drive-point resistance. Similarly, the gain and drive-point resistance of *any* equilateral (one-wavelength) loop is *totally* independent of the position of the drive point on the loop.

I shall now return to the horizontally polarized square of **fig. 2**. We have shown that a square loop (relative to a half-wave dipole) has a somewhat enhanced gain (+0.84 dB), made up of an increase (+1.07 dB) due to the vertical separation of the radiating segments (H-plane narrowed somewhat) and a decrease (−0.23 dB) due to the shortened or truncated radiating segments (E-plane broadening). Let's now explore the performance of a *rectangular* one-wavelength loop; **fig. 6** shows some examples.

Rectangle **A** is a wide but low loop which is recognized as a folded dipole loop; **B** is a narrower and higher loop than the square, and **C** is a high but very narrow loop. For all of these loops the sum of the width, W, and height, H, measured in wavelengths is constrained to be just 0.5. We are now in a position

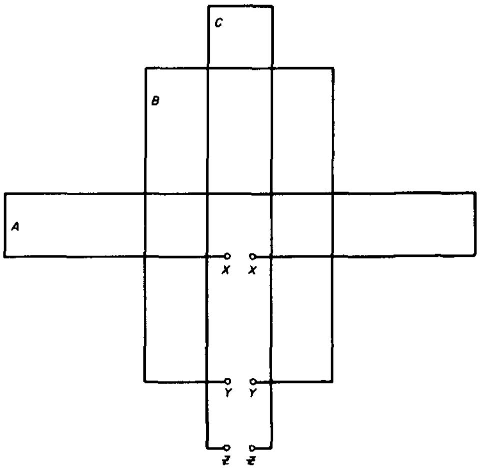

fig. 6. Diagram of three different rectangular configurations. In each case the feed point is located for horizontally polarized radiation.

table 1. Stacked half-wavelength dipoles showing gain vs. dipole separation.

separation S(λ)	gain (dBi)	stacking gain (dB)
0.000	2.151	0.000
0.125	2.425	0.274
0.250	3.236	1.085
0.375	4.514	2.363
0.500	5.978	3.827
0.625	6.938	4.787
0.750	6.758	4.607
0.875	5.831	3.680
1.000	4.929	2.778
1.250	4.373	2.222
1.500	5.275	3.124
1.750	5.844	3.693
2.000	5.097	2.946

to compute loop performance. **Table 4** shows the results of calculations for rectangular loops by the same methods as used for the square. D_2/D_1 is the H-plane directivity increase due to the vertical separation of radiating segments (see **eq. 20**), while D_1 is the directivity of a single horizontal radiating segment (see **table 1**). R is the loop driving point resistance (see **eq. 22**). Note that the limiting case of the folded dipole ($W = 0.5, H = 0$) shows a gain of 2.151 dBi (identical to a single half-wave dipole) and a driving-point resistance of 292.5 ohms (just four times that of a single half-wave dipole). Since reactance effects of the full loop are double that of a single "element," the Q of the folded dipole will be just one-half the Q of a single half-wave dipole, leading automatically to a bandwidth twice as large. As one reduces the loop width from this limiting case the gain increases monotonically. However, this favorable increase in gain is automatically accompanied by a significant reduction in driving point resistance; since reactance effects are essentially identical for all loops, the circuit Q increases commensurately. Thus the potential gain obtainable from high narrow loops is always offset by unfavorably high values of Q and corresponding narrow bandwidths!

table 2. Self-resistance and gain for one truncated element.

element length (λ)	R_{11} (ohms)	directivity	gain dBi
0.05	1.955	1.502	1.768
0.10	7.590	1.510	1.789
0.15	16.255	1.522	1.823
0.20	26.966	1.537	1.868
0.25	38.547	1.557	1.922
0.30	49.780	1.578	1.980
0.35	59.562	1.599	2.040
0.40	67.023	1.619	2.094
0.45	71.612	1.635	2.134
0.50	73.130	1.641	2.151

Let us now consider other loop shapes. For equilateral shapes, we have already seen that gain and driving-point resistance are independent of feed-point resistance on the loop. Moreover, a reasonably rigorous solution has been obtained for a square loop. Because of the independence of properties on feed-point position, which is equivalent to independence on loop rotation with feed point fixed at the bottom, it seems reasonable to assume that gain from all equilateral loops is approximately equal to that of a circular loop having the same enclosed area. This in turn can be equated to an equivalent square of the same area. Such an intuitive model, together with a model of how E-plane directivity varies with element length (see **table 2**) and how H-plane directivity varies with element separation (see **table 1**), allows a reasonable estimate of equilateral loop gain. **Table 5** lists these values for several equilateral loops. The values for the square are the ones already computed (see **table 4**)l; all others are estimated by this simple model. It is quite easy to see that the popular triangle or delta loop is slightly lower in gain (by 0.3 dB) than the square (quad or diamond) and that the loop with the highest gain is a circle. However, note that the gain of the circular loop is estimated as 3.28 dBi which is 1.13 dB larger than that of a half-wave dipole. This is not quite as large an increase as the approximately 2 dB which was quoted by Lindsay,[4] but I believe this discrepancy is probably not outside of the experimental accuracy range of Lindsay's measurements combined with the estimation accuracy range of the simple model I have used.

table 3. Mutual resistance, R_{12}, for λ/4 truncated elements vs. separation in λ.

separation (λ)	mutual resistance (ohms)	separation (λ)	mutual resistance (ohms)
0.00	38.547	0.80	−9.783
0.05	37.782	0.85	−7.165
0.10	35.536	0.90	−4.194
0.15	31.951	0.95	−1.142
0.20	27.251	1.00	1.730
0.25	21.729	1.10	6.083
0.30	15.721	1.20	7.713
0.35	9.589	1.30	6.481
0.40	3.686	1.40	3.190
0.45	−1.659	1.50	−0.787
0.50	−6.170	1.60	−4.008
0.55	−9.636	1.70	−5.450
0.60	−11.930	1.80	−4.806
0.65	−13.010	1.90	−2.518
0.70	−12.919	2.00	0.446
0.75	−11.780		

I shall now consider a multiloop square quad array where not only driven elements but parasitic loops as well are used. From what we have just seen, the table 4. Properties of the rectangular loop.

width $w(\lambda)$	height $H(\lambda)$	D_1	D_2/D_1	D_2	gain dBi	R ohms
0.05	0.45	1.502	2.073	3.114	4.933	3.77
0.1	0.4	1.510	1.815	2.741	4.379	16.7
0.15	0.35	1.522	1.596	2.430	3.856	40.8
0.2	0.3	1.537	1.419	2.181	3.386	76.0
0.25	0.25	1.557	1.279	1.991	2.990	120.5
0.3	0.2	1.578	1.172	1.850	2.672	169.8
0.35	0.15	1.599	1.094	1.750	2.430	217.7
0.4	0.1	1.619	1.041	1.686	2.267	257.5
0.45	0.05	1.635	1.010	1.652	2.179	283.6
0.5	0.0	1.641	1.000	1.641	2.151	292.5

quad array should behave just about like an equivalent stacked Yagi array separated (vertically) by 0.25λ and having elements bent together to form the individual square quad loops. However, to proceed with a computation of the properties of the parasitic quad array, we must know all complex self and mutual impedances of the truncated elements. Up to this point it has been possible to rather rigorously describe the properties of a driven loop because only the real part of self and mutual impedances were required to obtain gain, driving-point resistance, and pattern information (both in the E- and H-planes).

When parasitic elements are involved, the imaginary impedance terms are required. Computation of the imaginary impedances is a non-trivial exercise, and I am unaware of any rigorous procedure for carrying out such a calculation for truncated elements. As far as self-impedance is concerned, the reactive or imaginary value is controlled entirely by the "tuning" of the loop; that is, the relationship of wavelength to loop circumference and the effective loop Q. The complex impedance has been calculated for a linear nearly half-wave thin cylindrical element.[5] The method involves treating the metallic cylinder as a boundary-value problem (tangential components of electric field are made to vanish at every point on the conductor surface), from which an integral equation is derived. Approximate solutions of this integral equation yield the current distribution on the cylindrical element from which the input impedance, including the imaginary component, was derived.

I have been unable to find an equivalent rigorous boundary-value calculation for the square quad loop; thus we do not yet have the basis for calculating the precise reactance for a nearly one-wavelength square loop. However, it is possible to at least estimate the loop Q (but not its precise resonant frequency) by remembering that reactance changes with frequency are due to the effective inductance and capacitance of the antenna; that is, near-field stored energies, whereas the resistance (radiation) has to do with far field. Truncating the half-wave dipole changes the geometry of the element but hardly affects its (central) inductance or (end) capacitance. It would be reasonable, therefore, to assume the quad loop contains essentially the same total reactive impedance changes as two half-wave dipoles. Thus, following the argument made in a previous article,[3] we may write for the loop self-impedance:

$$Z_{11} = R_{11}+jX_{11} = R_{11}\,[1+j\cdot 2Q(F/FR-1)] \qquad \textbf{(24)}$$

but (empirically)

$$R_{11}Q = (430.3\log_{10}K-320) \qquad \textbf{(25)}$$

so that

$$Z_{11} = 120.5+j\,(860.6\log_{10}K-640)\,(F/FR-1)\ ohms \qquad \textbf{(26)}$$

As in the previous article, F is the normalized frequency and FR the normalized resonant frequency of the loop. Loop Q is readily estimated from **eqs. 25** and **21**. The only remaining problem is a determination of FR. Although there is no rigorous way of calculating FR from basic principles, it is significant to note that the region where the ends of the two dipoles are joined must have electric (capacitive) fields at right angles to the conducting cylindrical element, exactly like those of an infinitely long straight cylinder near a voltage loop. This observation implies that there should be a negligible "end effect" at the capacitive voltage loop, and, therefore, FR should be very close to the frequency at which the total loop circumference is just one wavelength.

When we consider the imaginary part of the mutual impedance between loop halves or "elements," another computational complication arises. At long distances or separations, the imaginary mutual reactance, X_{21}, must be (except for a phase shift) simply related to the real part, R_{21}, and this relationship should be unaffected by the precise "tuning" of the "elements." However, at very small separations X_{21} must approximate the value of X_{11}, which is fundamentally fixed by circuit Q and resonant frequency and not at all by R_{11}. How to correctly represent X_{21} at all intermediate spacings has not, to my knowledge, been solved quantitatively. For this reason I will model the quad array first as the equivalent Yagi array stacked at a spacing of $\lambda/4$ then, second, apply necessary corrections to direc-

table 5. Estimated equilateral loop properties.

equilateral loop type	sides	equivalent square side (λ)	D_1	D_2/D_1	D_2	gain dBi	R ohms
triangle	3	0.219	1.545	1.205	1.862	2.70	104.7
square	4	0.250	1.557	1.279	1.991	2.99	120.5
pentagon	5	0.262	1.562	1.309	2.044	3.10	126.4
hexagon	6	0.269	1.565	1.324	2.071	3.16	129.3
octagon	8	0.275	1.567	1.338	2.097	3.22	132.0
circle	∞	0.282	1.570	1.356	2.129	3.28	135.3

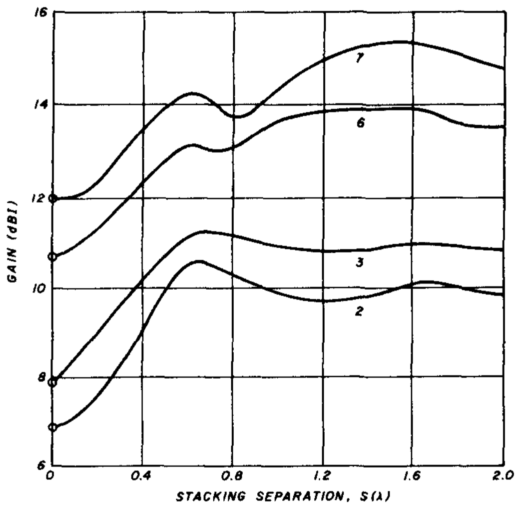

fig. 7. Representative examples of stacking gain for 2, 3, 6, and 7-element Yagis *vs* the stacking separation.

tivity and gain. Compared with stacked (λ/2 element) Yagis, the square quad will have 0.23 dB *less* directivity and gain (all due to *E*-plane pattern broadening).

It is interesting to calculate the free-space "stacking gain" of Yagis as a function of vertical separation. I chose as representative boom lengths and number of elements the following simplistic Yagis, which were found to have excellent free-space properties:[6] a two-element beam on a *0.15*-λ boom, a three-element beam on a *0.25*-λ boom, a six-element beam on a *0.75*-λ boom, and a seven-element beam on a *1.25*-λ boom. **Table 6** lists characteristics of these beams.

Fig. 7 shows the computed gains of these stacked configurations as a function of the stacking separation, *S*, in units of λ. Note that the rise in gain due to *S* is somewhat different for each case and also somewhat different than the case for dipoles shown in **fig. 3**. Two things seem to be occurring as *S* is increased. For small separations, the capture area of one Yagi essentially overlaps that of the other Yagi; therefore, the total capture area for both is essentially the same as for one alone. As *S* increases, the total capture area increases and ultimately doubles if *S* is large enough. For the larger Yagis (where the original capture area for one Yagi is large), it is easy to see that to realize a given separation, gain *S* must be relatively larger than for a small Yagi or especially a dipole. In other words, the transition from the gain of one large Yagi to the doubling of gain (3-dB increase) for two large Yagis requires a *larger* separation than is required for smaller Yagis. In addition to this rather gradual gain increase due to separation of capture areas, **fig. 7** suggests that the constructive interference due to mutual impedances noticed for the dipoles in **fig. 3** also persists in stacked Yagis. An increase in gain is noticeable at $S = 0.6$ and also at 1.6. Qualitatively, **fig. 7** shows a combination of these two effects; first, the constructive impedance gain increases at particular separations and second, the capture area separation gain increase which takes place more slowly with large Yagis.

table 6. Characteristics of representative beams. Radius of all elements = 0.0005260λ.

number elements	length (λ) refl	driven	dir.	boom length (λ)	element spacing (λ)	fig. 7 curve
2	.4937	.4705	–	0.15	0.150	2
3	.4980	.4896	.4690	0.25	0.125	3
6	.4953	.4803	.4481	0.75	0.150	6
7	.4936	.4762	.4466	1.25	0.208	7

It is interesting to compare the computed gain increase of the seven-element beam with the experimental values reported by Viezbicke[7] in his **fig. 11A**. **Table 7** shows a comparison of my computed results with Viezbicke's published experimental values. The comparison is not totally valid because Viezbicke's seven-element beam is not the same beam I have used in **fig. 7**. The differences cannot be quantified, since Viezbicke failed to include specifications for his seven-element beam. Nevertheless, **table 7** shows good qualitative agreement and even fair quantitative agreement (close to Viezbicke's stated experimental accuracy of 0.5 dB).

table 7. Stacking gain (dB) of seven-element beams.

spacing S (λ)	(Viezbicke) 7 element	computed 7 element
0.38	0.80	1.3
0.57	1.58	2.15
0.78	1.36	1.8
0.99	1.90	2.35
1.20	2.34	3.0
1.40	2.53	3.25
1.61	1.93	3.3

Computations represented by the data shown in **fig. 7** cannot really be carried down to very small separations with any confidence because the mathematical model uses mutual impedances of full λ/2 elements. When reactive parasites get very close together their mutual impedance has an imaginary value quite different from that of λ/2 dipoles.

I shall now examine the stacking gain increases for a separation of λ/4. **Table 8** shows these increases for the four computed cases (dipole data from **fig. 3**). Note that the λ/4 stacking gain for very large Yagis is *not* as large as that for dipoles;

this can easily be understood because of the (relatively) smaller increase in capture area. One might expect this gain increase to fall monotonically from the value of 1.09 dB (for dipoles) as the array length is increased, however, **table 8** shows the actual increase to vary somewhat due to the detailed way in which impedances vary.

We are now in a position to estimate the performance of a quad array. If instead of the dipole elements of the stacked ($\lambda/4$) Yagi arrays I use (bent) elements connected in square loops, I will make a quad array in which all conditions remain about the same *except* that the E-plane pattern is broadened and the gain is correspondingly reduced due to the truncated "elements." The gain performance of the quad array is therefore about 0.23 dB lower than the stacked $\lambda/4$ Yagis. These estimations are also shown in **table 8.**

table 8. Quarter wavelength stacking gain for Yagis and estimated quad/single Yagi gain.

number elements	boom length (λ)	Yagi S = $\lambda/4$ stacking gain (dB)	quad/single Yagi gain difference (dB)
1	0.0	1.09	0.86
2	0.15	1.03	0.80
3	0.25	1.38	1.15
6	0.75	0.84	0.61
7	1.25	0.65	0.42

It must be evident by now that a square quad array is very much like an equivalent Yagi. Its overall gain is expected to be somewhat higher than the Yagi because of individual loop gain, but not by very much. E-plane pattern is slightly broader due to the current distribution on the truncated "elements," whereas H-plane pattern is slightly narrower due to the "stacking gain" of the loops. Because of this similarity of Yagi and quad arrays, it is plausible that one can intermix quad loops and Yagi elements to provide a hybrid structure of roughly equal performance. Such a hybrid is known as a "quagi;" if properly constructed, it should provide a pattern and gain intermediate between a similar quad and a similar Yagi. There are obviously an enormous variety of possible quagi configurations; there will remain for some time a challenge to the quagi designer to determine preferred configurations and best dimensions for all radiating elements.

summary

A number of interesting conclusions have been reached regarding antenna arrays constructed with conducting loops roughly one wavelength in circumference.

1. A single (driven) loop will provide a free-space gain somewhat larger than that of a half-wave dipole. The gain increase comes about through a narrower H-plane pattern and a slightly broader E-plane pattern.

2. Loop gain varies significantly with shape. For rectangular loops fed in the center of the lower horizontal segment, gain depends on the ratio of height, H to width, W, varying from 2.15 dBi (equal to a half-wave dipole) if $H/W = 0$ to about 5 dBi (3.8 dB above a half-wave dipole) as H/W approaches zero.

3. For equilateral loops, gain depends only on the number of sides (see **table 5**). For the square loop, the free-space gain calculated rather rigorously is 2.99 dBi and driving-point resistance is 120 ohms. It is significant that these properties of the square loop *are totally independent of the feed point on the* square. As an example, the gain and drive-point resistance for a quad square is *exactly* the same as that of the diamond square.

4. The free-space gain of a quad array is estimated to be somewhat higher than that of an equivalent (single) Yagi. Calculations show this difference depends somewhat on the particular array (see **table 8**), but ranges from about 1 dB for short arrays to less than 0.5 dB for long arrays.

5. Quagi configurations are expected to show performance figures between those of an equivalent Yagi and equivalent quad.

6. A rigorous theory does not yet exist for self and mutual quad loop reactances. Consequently, quad (and quagi) parasitic loops must be experimentally adjusted for correct resonant frequency. It is unlikely that such experimental adjustments can be made with the same precision and confidence that Yagi element lengths can now be specified by present theory. Therefore, the slight gain advantage of the quad over the Yagi shown in **table 8** may well disappear in practice.

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ham radio

navigational aid

for small-boat operators

An idea for approaching harbor entrances using two simple beacon transmitters and a harmonic phase detector

Many small-boat operators have probably wished for a simple navigational aid during fog, darkness, or other hazards. The system suggested here provides guidance between two simple low-powered transmitters, which may be located on opposite sides of a channel, breakwater entrance, or even on a wharf that may be the goal of the boat. It's a sort of SHORAN system, with simplifications, which should put it well within the technical ability, as well as the pocketbook, of the average Amateur.

theory

Consider **fig. 1**, which shows points **A** and **B**, representing the location of two transmitters, which may be battery powered and in the milliwatt range. Our boat is located at point **C**, and our aim is to guide the boat between points **A** and **B** using signals transmitted from these two points. Transmitters at **A** and **B** operate on closely chosen radio frequencies, which differ by some arbitrary audio frequency, perhaps several hundred to one thousand Hz. The frequency separation is well within the passband of a receiver located on the boat at point **C**. Receiver output, therefore, is the beat frequency caused by this difference in the transmitted frequencies.

If now, the boat maneuvers so that the path length between it and the higher-frequency transmitter becomes relatively shorter than the path to the lower-frequency transmitter, the phase of the received beat frequency signal will advance. This effect may be generalized by stating that the received beat frequency is phase-variant with the position of point **C**, the point of reception.

analysis

A number of paths exist along which the receiving point, **C**, can move without changing the phase of the beat note. These paths are a family of hyperbolas, *all of which pass between* **A** *and* **B**. On the line **AB**, they will be a half-wavelength apart, but as they appear further from line **AB**, they separate. This, in effect will "funnel" the boat between the two target points. The number of these paths will be a function of the separation between **A** and **B** in wavelengths (**fig. 2**).

To use this phase-variant characteristic of the beat frequency, a reference signal is needed with which the beat frequency can be compared. The generation and transmission of this signal is a relatively simple process, for which two alternatives exist.

reference signal

In the antenna feedline of either transmitter, we can insert a directional coupler. While signals from the local transmitter are being radiated from this antenna, signals from the other transmitter will be

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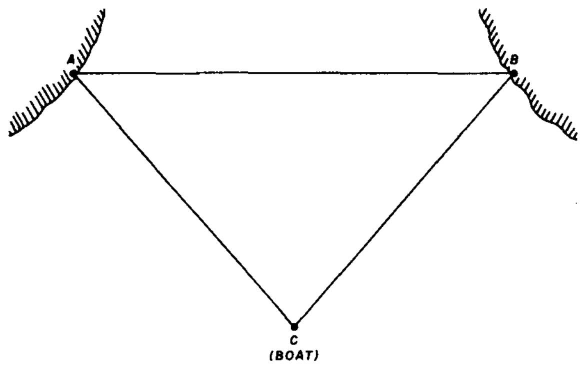

fig. 1. Example of a typical navigational problem. Beacon transmitters are at points A and B at a harbor entrance. Guidance may be provided to a small boat entering the channel using the idea in the text.

picked up. The directivity of this coupler is used to reduce the local signal, while accepting as much as possible of the received signal from the other transmitter.

When detected, a beat note is produced, which is the same frequency as that received at point **C**, but which is completely independent of the position of point **C**. In other words, here is our reference signal.

It could be transmitted to point **C** by using it to modulate a transmitter operating on a third frequency, but this would require a considerable amount of additional equipment both on shore as well as on the boat. Instead, we can double the beat frequency thus produced and use it to amplitude-modulate that transmitter, to a modest percentage, probably below 50 per cent, to avoid distortion. Appropriate filtering built into the modulator system will remove the fundamental.

The second alternative would be to divide the beat frequency by two, and with appropriate filtering, as before, use it to modulate that transmitter. Frequency division by digital methods should produce a symmetrical square wave, which would contain, theoretically, no trace of its second harmonic, which of course is the beat-note fundamental.

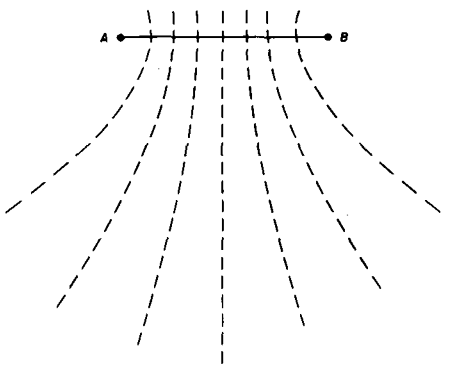

fig. 2. Geometry of equi-phase paths between points A and B of *fig. 1. In this example* AB = 4 wavelengths. Paths above AB line are the mirror image of those below.

This square wave would have to be reduced to a sine wave by some means, such as a low-pass filter to remove harmonics. A square wave contains only odd harmonics, so this doesn't appear too difficult for several stages of RC feedback filtering with a multiple op-amp.

phase detector

Now let us move to the boat, at point **C**. Its receiver will be producing two audio frequencies, having a 2:1 frequency relationship, one of which is phase-variant with the boat's position. This audio-frequency signal is now fed into a harmonic phase detector, **fig. 3**, which responds to two such harmonically related signals, giving 1) a balanced zero voltage when both signals cross the zero axis simultaneously, and 2) a positive or negative voltage, according to which signal crosses the zero axis before the other (reference 1).

The differential voltage produced by the phase

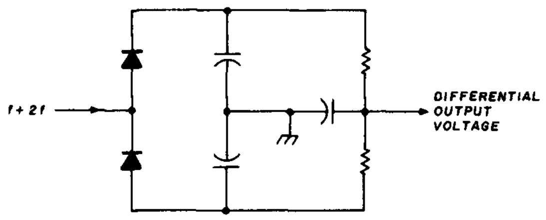

fig. 3. Harmonic phase detector produces a differential voltage that can be used to drive a zero-center meter that will deviate from center in accordance with the boat's deviation from one of the equi-phase paths in fig. 2.

detector is easily amplified to drive a zero-center meter, which will deviate from center in accordance with the boat's deviation from one of the equi-phase paths.

some final thoughts

I suspect that such a system might operate, for example, in one of the 20-kHz slots that appear in the CB spectrum. This would reduce the acqusition of new equipment by the average boat enthusiast to a bare minimum. If he stayed too late at the fishing grounds, he could call the operator of the marina where he keeps his craft and ask him to set the beacons out by the entrance through the breakwater. The marina operator should enhance his popularity by doing so, not only with his own customers, but with the boating fraternity in general.

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Easy selection.

15 memories/offset recall, scan, priority, DTMF
Touch-Tone ®

TR-7800

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TR-7800 FEATURES:

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  Entire band (5-kHz or 10-kHz steps) and memories. Automatically locks on busy channel; scan resumes automatically after several seconds, unless CLEAR or mic PTT button is pressed to cancel scan.

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optimum pi-network design

Methods for optimizing bandwidth without using the Q factor — design parameters and examples are given for this useful circuit

The familiar pi network shown in **fig. 1** is both a lowpass filter and a simple impedance-matching network. Stray reactances at source or load may be compensated for by capacitor adjustment. Traditional design methods have used a Q factor to choose correct component values. This article presents design methods for optimizing bandwidth without using the Q factor.

The main function of the pi network is to provide a resistive match of source resistance, $R1$, with a load resistance, $R2$. An infinite number of component possibilities are available; two examples are shown in **fig. 2**.

This article was rewritten by Leonard H. Anderson, who is a member of the technical staff of Rocketdyne division of Rockwell International, Inc., and is well known for his contributions to *ham radio* in the *Digital Techniques* series. The *ham radio* staff expresses its thanks to Mr. Anderson for his help in interpreting this difficult subject.

impedance transformation diagrams

Fig. 2 is useful for any simple network. Each resistance circle represents the series-form impedance of the parallel combination of end resistance with shunt reactance. This chart is used in the following manner:

1. Calculate series form $R+jX$ of one end, including the shunt reactance.

2. Find the intersection on the resistance circle for both R and X.

3. Move vertically from this intersection to intercept the other resistance circle.

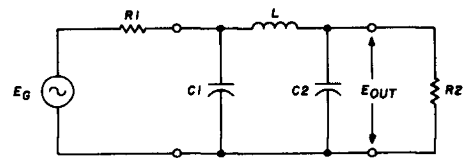

fig. 1. Classic pi network. Circuit is both a lowpass filter and impedance-matching device. Its main function is to provide a resistive match of source resistance, $R1$, with a load resistance, $R2$.

4. Measure the length of the vertical movement; this total reactance is the series-arm value.

5. Determine the reactance at the second intersection as measured from the zero reactance point (R axis passing through reactance axis).

6. Take the *opposite* reactance sign and calculate the new combination of parallel resistance and reactance.

The pi network first intersections are shown for the series impedance of $R1$ and $C1$. Total vertical movement is the reactance of series inductor L. The

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second intersections on the $R2$ circle represent the impedance presented to the inductor and $R1$, $C1$. To obtain a resistive match, the second intersections must be changed from a positive reactance to negative; this *conjugate* operation moves the Z_2 intersection to the R axis.

As long as each set of intersections falls on each resistance circle, any component combination will yield a perfect impedance match. One particular combination will give the widest bandwidth, and some network synthesis techniques can be used to find that combination.

image impedances

The image impedance of a network is the square root of the product of one input impedance with the *other end open and the same input impedance with*

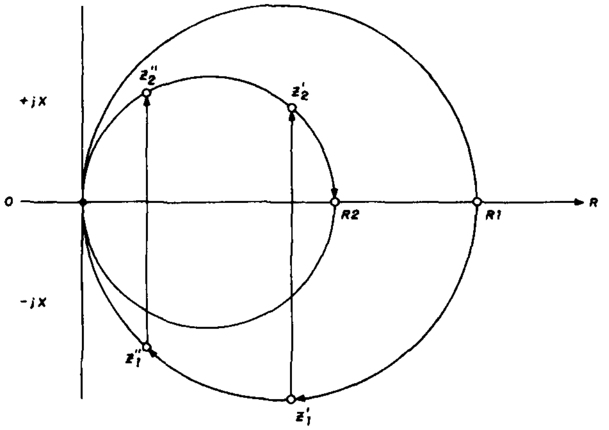

fig. 2. Impedance-transformation diagram. Each resistance circle represents the series-form impedance of the parallel combination of end resistance with shunt reactance.

the other end shorted. An image impedance is not the parameter for direct design but a mathematical tool for achieving the final solution.

The open/shorted opposite-end technique is used in microwave measurements where it's difficult to obtain a pure resistance.[1] At one particular frequency the image impedance will represent the actual input impedance when the opposite end is loaded with a resistance. An example is an infinite length of transmission line: at any frequency the image impedance is equal to the *characteristic impedance* of the line.

Fig. 3 shows the pi-equivalent circuit composed of pure reactance and susceptances. Using a matrix representation of the network, the following is true:

$$A_{11} = 1 + Z_2Y_3 \qquad A_{12} = Z_2$$
$$A_{21} = Y_1 + Y_3 + Y_1Z_2Y_3 \qquad A_{22} = 1 + Y_1Z_2$$

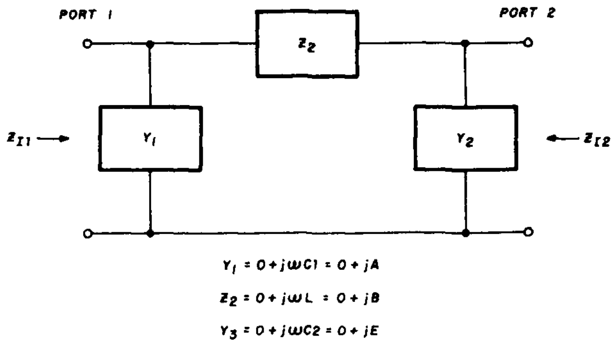

fig. 3. Pi-equivalent circuit composed of pure reactance and susceptances. These parameters at any frequency are derived in the text.

Using these matrix values, image impedance Z_{I1} becomes:

$$Z_{I1}^2 = Z_{o1}Z_{s1} = \frac{A_{11}A_{12}}{A_{21}A_{22}}$$

$$= \frac{B(BE-1)}{(AB-1)(A+E-ABE)} \tag{1}$$

Letters A, B, and E are the susceptances and reactances at any frequency as given with **fig. 3**. Z_{o1} is the impedance into port 1 with port 2 open; Z_{s1} is the port-1 impedance with port 2 shorted. The image impedance at port 2 is:

$$Z_{I2}^2 = Z_{o2}Z_{s2} = \frac{A_{12}A_{22}}{A_{11}A_{21}} = \frac{B(AB-1)}{(BE-1)(A+E-ABE)} \tag{2}$$

Z_{o2} and Z_{s2} are the port-2 impedances with port 1 open and shorted respectively. Both right-hand expressions are complex numbers with a zero-value imaginary component.

A broad spectrum plot of both image impedances is shown in **fig. 4** for a typical network. It is common to denote frequency in such plots as ω, the product of $2\pi \times$ *frequency*. The dash lines indicate a negative-real-part image impedance.

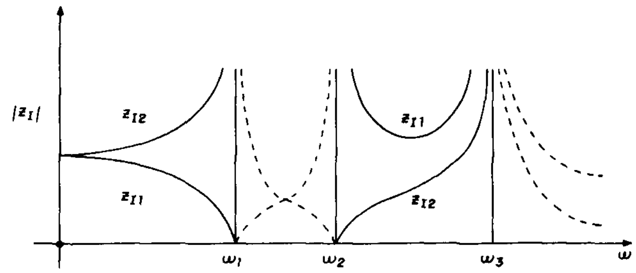

fig. 4. Image impedances at input and output ports as functions of frequency. Frequency is denoted as $\omega = 2\pi f$. Dashed lines indicate a negative-real-part image impedance.

Synthesis of component values is not possible between ω_1 and ω_2 nor above ω_3 because of the negative real parts of the impedance. The only choice is below ω_1 or between ω_2 and ω_3.

center frequency impedance transformation

Assuming that $R1$ is greater than $R2$ will exclude any frequency condition below ω_1. The range between ω_2 and ω_3 is left and shown expanded in **fig. 5**.

Frequency ω_0 is the geometric mean of ω_2 and ω_3, expressed as:

$$\omega_0 = \sqrt{\omega_2 \omega_3} \quad \textbf{(3)}$$

The intersection of ω_0 and each image impedance plot line will give $R1 = Z_{I1}$ and $R2 = Z_{I2}$. Frequency ω_0 is the network center frequency.

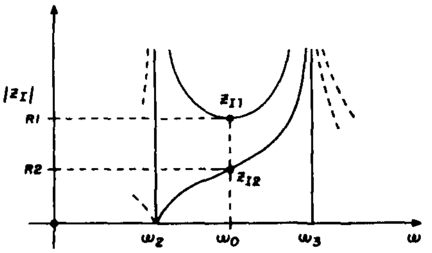

fig. 5. Impedance-transformation relationship to geometric center frequency, ω_0, which is the geometric mean of ω_2 and ω_3. Note that Z_{I1} has the least change at ω_0.

Holding to the network center frequency, ω_0, the susceptances and reactances of each pi-network arm are defined as:

$$A_0 = \omega_0 C1 \quad \textbf{(4)}$$

$$B_0 = \omega_0 L \quad \textbf{(5)}$$

$$E_0 = \omega_0 C2 \quad \textbf{(6)}$$

Each end resistance can then be related to all component values by substituting **eqs. 4, 5,** and **6** into **eqs. 1** and **2**:

$$R1^2 = \frac{B_0(B_0E_0 - 1)}{(A_0B_0 - 1)(A_0 + E_0 - A_0B_0E_0)} \quad \textbf{(7)}$$

$$R2^2 = \frac{B_0(A_0B_0 - 1)}{(B_0E_0 - 1)(A_0 + E_0 - A_0B_0E_0)} \quad \textbf{(8)}$$

Defining a ratio, m, as $R1$ divided by $R2$, an identity is obtained from **eqs. 7** and **8**:

$$m = \frac{R1}{R2} = \frac{B_0E_0 - 1}{A_0B_0 - 1} \quad \textbf{(9)}$$

The relationship of ω_2 and ω_3 frequencies must now be established.

optimizing the design-center frequency

Examination of **fig. 5** shows that Z_{I1} has the least change of value at ω_0. This can be proved by taking *the derivative of Z_{I1} as a function of frequency with* **eq. 1**. Z_{I2} has a relatively constant slope at ω_0. Choosing ω_0 as the design center frequency results in a minimum impedance change at each network port over a given bandwidth.

To find ω_0 from **eq. 3**, cut-off frequencies ω_2 and ω_3 must be found. Image impedance plots in **figs. 4** and **5** will show that Z_{I1} goes to infinity at each frequency and Z_{I2} is zero at ω_2 and infinity at ω_3. Examination of **eq. 1** shows that the denominator becomes zero if $AB = 1$. A zero denominator will yield a result of infinity. Similarly, the numerator of **eq. 2** will be zero if $AB = 1$. A relationship to ω_2 now exists, and from the expressions in **fig. 3**:

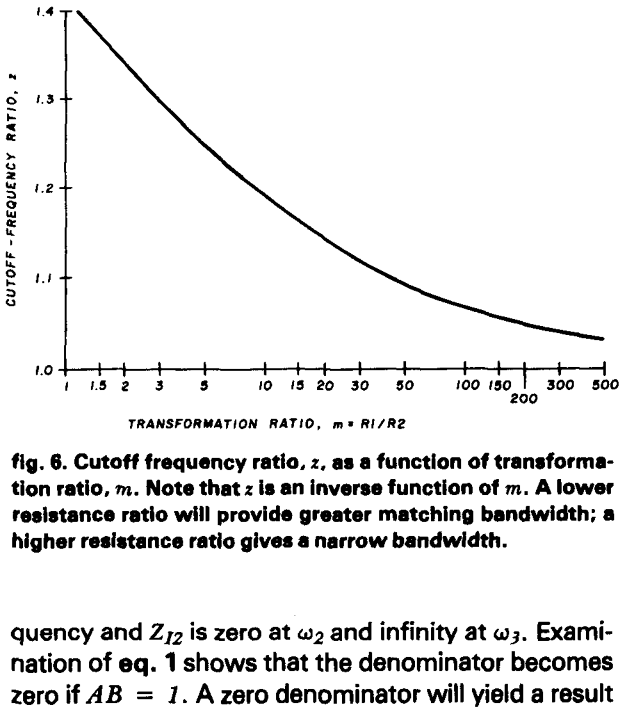

fig. 6. Cutoff frequency ratio, z, as a function of transformation ratio, m. Note that z is an inverse function of m. A lower resistance ratio will provide greater matching bandwidth; a higher resistance ratio gives a narrow bandwidth.

$$\omega_2^2 = \frac{1}{C1L} \quad \textbf{(10)}$$

Both image impedances are infinity at ω_3. This condition will result if the term $(A + E - ABE)$ is zero or $(A + E) = ABE$. This term is common to **eqs. 1** and **2**. Using the expressions in **fig. 3**,

$$\omega_3^2 = \frac{C1 + C2}{C1\,LC2} \quad \textbf{(11)}$$

Eqs. 10 and **11** are a bit clumsy to handle directly. It

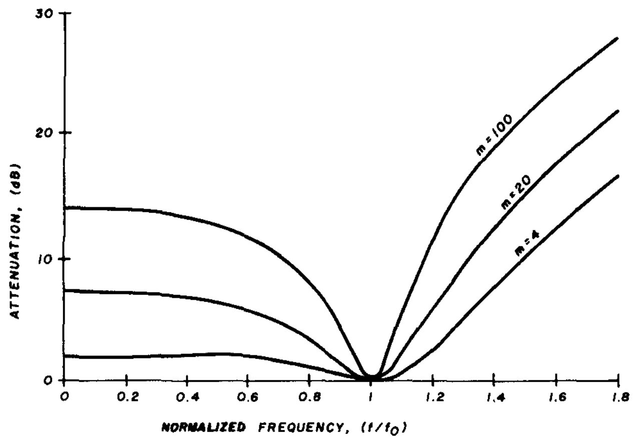

fig. 7. Normalized attenuation characteristic for various transformation ratios. Curves illustrate the difference in response caused by transformation ratio m.

will be easier to handle if another ratio is used:

$$z = \frac{\omega_3}{\omega_2} \tag{12}$$

Some algebraic manipulation of **eqs. 3** through **6** and **12** result in the identities

$$z = A_0 B_0 = \sqrt{\frac{A_0 + E_0}{E_0}} \tag{13}$$

normalizing the component values

More algebraic manipulation using **eqs. 4, 5, 6; 9** through **11;** and **13** give the following relationships:

$$A_0 = \frac{\sqrt{m(z+1)}}{R1} \quad \text{and } C1 = \frac{\sqrt{m(z+1)}}{\omega_0 R1} \tag{14}$$

$$B_0 = \frac{z\,R1}{\sqrt{m(z+1)}} \quad \text{and } L = \frac{z\,R1}{\omega_0\sqrt{m(z+1)}} \tag{15}$$

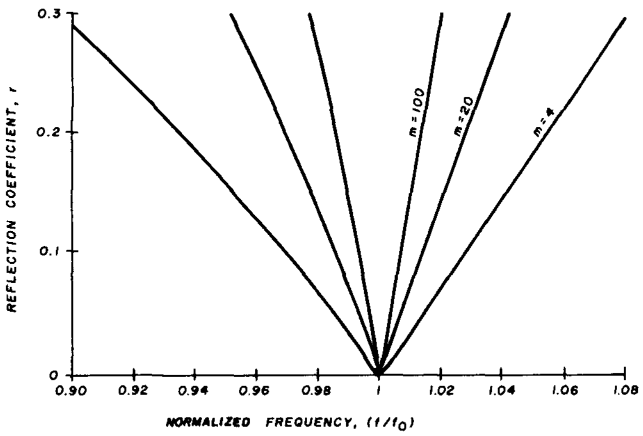

fig. 8. Normalized reflection characteristic for various transformation ratios. Best operation occurs when reflection coefficient $r < 0.1$.

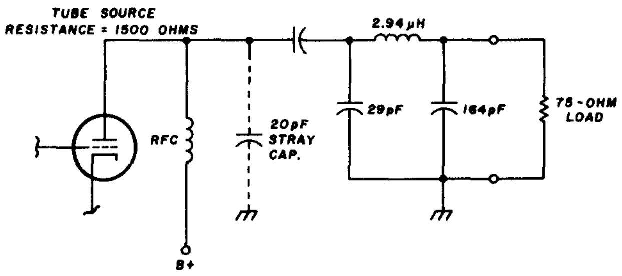

fig. 9. Pi net example for a tube amplifier. Frequency range is between 13.9-14.4 MHz. The reflection coefficient can be checked against fig. 8 for the band edges.

$$E_0 = \frac{1}{R2(z-1)\sqrt{m(z+1)}} \tag{16}$$

$$\text{and } C2 = \frac{1}{\omega_0 R2(z-1)\sqrt{m(z+1)}}$$

Cutoff frequency ratio, z, can't be selected arbitrarily. It has a definite relationship to end-resistance/transformation ratio m. Using **eqs. 9** and **13** through **16**, this is:

$$z^3 - z^2[(m-1)/m] - z[(m+1)/m] + [(m-1)/m] = 0 \tag{17}$$

Fig. 6 is a plot of cutoff frequency ratio, z, versus transformation ratio, m. **Eq. 17** is valid only for m greater than unity and $R1$ greater than $R2$.

Note that z is inversely proportional to m. A lower resistance ratio will give a greater matching bandwidth; a higher resistance ratio gives a narrow bandwidth. A check on bandwidth is possible by direct analysis of a network versus frequency, but it may be easier to test by mathematical means.

transfer functions and properties

A *transfer function* is a mathematical expression of a network that allows comparison of input to output *vs.* frequency. It can also be used to find the *reflection coefficient* at the input port as a check of matching bandwidth. This reflection coefficient is the same as that of a loaded transmission line VSWR expression.

The common transfer function, $F(\Omega)$, representing available generator power versus output power, may be expressed in the normalized form for the pi network as:

$$F(\Omega) = \left(\frac{1}{2\sqrt{m}}\right)\left[A_{11} + \frac{A_{12}}{R2} + A_{21}R_1 + A_{22}m\right]$$

$$= G + jU = \sqrt{\frac{P_o}{P_2}} = \frac{E_g}{2E_{out}\sqrt{m}} \tag{18}$$

where A_{11}, A_{12}, A_{21}, A_{22} are the matrix terms given previously and

Ω = ω/ω_0, ratio of calculation versus center frequency

G = real part of $F(\Omega)$

U = imaginary part of $F(\Omega)$

P_o = available power of generator, R1

P_2 = power absorbed by R2

The transfer function, $F(\Omega)$, may be expressed in terms of Ω, z, and m by:

$$F(\Omega) = \frac{1}{2\sqrt{m}}\left[\frac{m(z^2-1)(1-\Omega^2 z)+z^2-\Omega^2 z-1}{z^2-1} + j\,\frac{\Omega z(2z-\Omega^2-1)\sqrt{m(z+1)}}{z^2-1}\right] \quad \textbf{(19)}$$

The term $\frac{1}{2\sqrt{m}}$ is the normalizing factor. Any combination of z and m will result in unity *magnitude* of $F(\Omega)$ at ω_0; this allows a response comparison without scale shifting.

Eq. 19 may be programmed on a calculator but must be converted to polar form.* Taken directly, the term in brackets of the normalized transfer function is the generator voltage of **fig. 1** divided by the complex output voltage across $R2$. The ratio of complex-output-voltage-to-complex-input-voltage of the network is the inverse of $F(\Omega)$, with the result multiplied by $1/2$.

Normalized attenuation around design center frequency can be expressed by:

$$e^{2\alpha} = |F(\Omega)|^2 = G^2 + U^2 \quad \textbf{(20)}$$

This response is plotted in **fig. 7** showing attenuation as a function of design center frequency. Attenuation in dB is $20\ log_{10}(e^{\alpha})$. This plot illustrates clearly the difference in response caused by transformation ratio m.

Reflection coefficient, r, is the reflected voltage

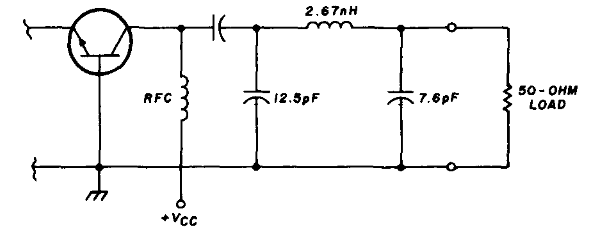

fig. 10. Example showing a transistor amplifier in the range of 1215-1300 MHz. Transistor output impedance is resistive at 12.5 ohms across the frequency band; *R1* is now the load and *R2* is the generator.

*A rectangular-form expression is usually easier to show in texts. The problem with such expressions is that the complex real part becomes negative! Conversion to polar form gives a positive magnitude with the correct phase angle.

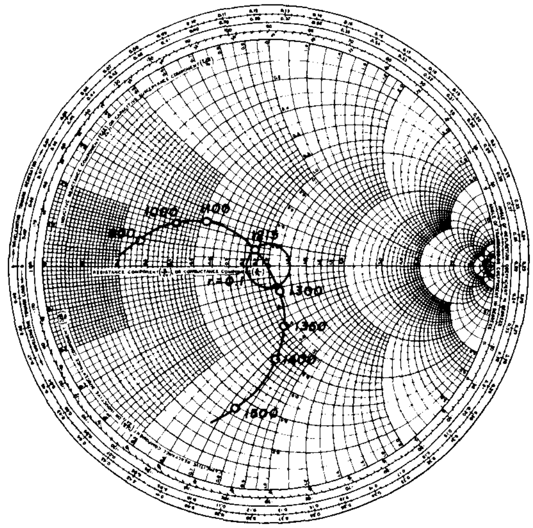

fig. 11. Smith chart showing network output impedance of a transistor amplifier and pi network. Output VSWR is less than 1.25 across the desired bandwidth.

divided by the forward voltage at the generator. *Normalized* reflection coefficient can be derived from **eqs. 19** and **20**:

$$r = \sqrt{\frac{e^{2\alpha}-1}{e^{2\alpha}}} \quad \textbf{(21)}$$

This expression is plotted in **fig. 8** for m values of 4, 20, and 100. Best operation occurs when r is less than 0.1. Bandwidth again depends on transformation ratio.

some examples

A vacuum-tube amplifier has an output impedance of 1500 ohms with stray capacitance of 20 pF. It is to be matched to a 75-ohm line over the range of 13.9-14.4 MHz. The geometric center of the range will be the design center frequency:

$$f_0 = \sqrt{13.9 \times 14.4} = 14.15\ MHz$$

$R1$ is 1500, R2 is 75; so the transformation ratio, m, equals 20. From **fig. 6**, $m = 20$; $z = 1.14$. **Eqs. 14, 15** and **16** give the component values after calculation of some common terms:

$$\omega_0 = 2\pi f_0 = 6.283 \times 14.15 \times 10^6 = 88.91 \times 10^6$$
$$\sqrt{m(z+1)} = \sqrt{20 \times 2.14} = 6.542$$

From **eq. 14**:

$$C1 = \frac{6.542}{88.91 \times 10^6 \times 1500} = \frac{6.542}{133.4 \times 10^9} = 49.06\ pF$$

Subtracting the 20-pF stray capacitance gives a component value of 29.06 pF. From **eq. 15**,

$$L = \frac{1.14 \times 1500}{88.91 \times 10^6 \times 6.542} = \frac{1.710 \times 10^3}{581.6 \times 10^6} = 2.940\ \mu H$$

From **eq. 16**:

$$C2 = \frac{1}{88.91 \times 10^6 \times 75 \times 0.14 \times 6.542}$$

$$= \frac{1}{6.107 \times 10^9} = 163.7\,pF$$

The tube and network circuit is shown in **fig. 9**. The cutoff frequencies, ω_2 and ω_3, can be checked with **eqs. 10** and **11** as 13.25 MHz and 15.11 MHz respectively.

The reflection coefficient can be checked against **fig. 8** for the band edges. The high end is *14.4/14.15 or 1.018* relative to center; low end is *13.9/14.15*, or *0.982*. The reflection coefficient is approximately 0.12, or 12 per cent.

Another example is a microwave amplifier output circuit (**fig. 10**). The desired range is 1215-1300 MHz. The transistor must match a 50-ohm load. The transistor output impedance is resistive at 12.5 ohms across the band; $R1$ is now the load and $R2$ the generator end. The design center frequency is:

$$f_0 = \sqrt{1215 \times 1300} = 1257\,MHz$$

Transformation ratio is 50/12.5 or 4, and z is 1.27 (from **fig. 6**). Remembering that $C2$ is next to the generator and $C1$ next to the load, **eqs. 14, 15**, and **16** give the following values:

$$C1 = 7.63\,pF,\ L = 2.67\,nH,\ C2 = 12.45\,pF$$

A Smith-chart plot of the network output impedance, normalized to 50 ohms, is given in **fig. 11**. Output VSWR is less than 1.25 across the desired bandwidth.

group delay and element dissipation

Group delay is the differential phase delay divided by the differential frequency across the desired band. It is the time delay of a signal through the network and is important for wideband modulation transmission and determination of network dissipation loss. Normalized group delay, $\omega_0 T_g$, can be determined from the complex transfer function of **eq. 19** as:

$$\omega_0 T_g = \frac{\left(\frac{dU}{d\Omega}\right) G - \left(\frac{dG}{d\Omega}\right) U}{G^2 + U^2} \quad \textbf{(22)}$$

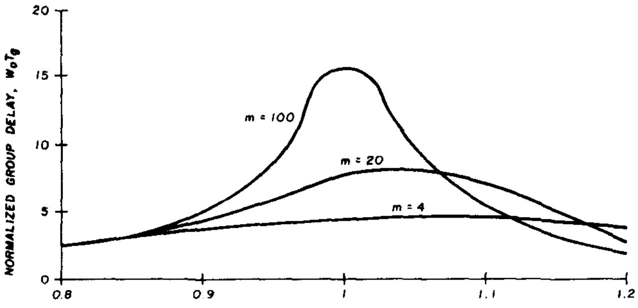

fig. 12. Normalized group delay for three transformation ratios, m.

Normalized group delay for three values of m is plotted in **fig. 12**. In general, maximum group delay occurs at rapid attenuation versus frequency.* The pi-network maximums are slightly higher than design center frequency.

Network attenuation other than the transformation ratio is determined by the unloaded Q of each reactive element. Knowing the element Q allows determination of loss through the normalized group-delay expression:

$$a_v = loss\ (in\ dB) = \frac{4.343\omega_0 T_g}{Q_u} \quad \textbf{(23)}$$

Eq. 23 is assisted by the design-center-frequency normalized group delay plotted in **fig. 13**. Assuming very high unloaded Qs of the capacitors, an unloaded Q of 160 for the inductor, and an m of 20 (from the first example), additional network loss will be 0.20 dB, a negligible amount.

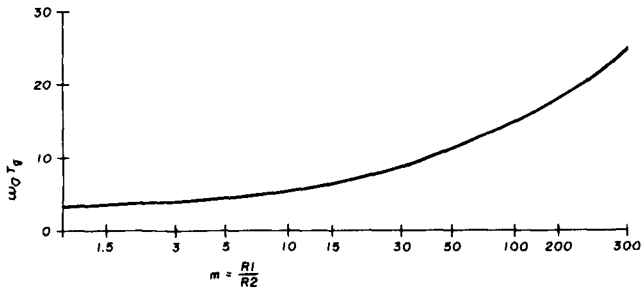

fig. 13. Normalized group delay at center frequency for various transformation ratios.

summary

The optimum design of a pi network depends on the transformation ratio. Bandwidth is inversely proportional to this ratio. Simple calculation of components is possible with the aid of a few graphs. Reflection coefficient is proportional to transformation ratio and may be used to determine if a network must be retuned for a particular bandwidth.

*Phase shift through a filter is responsible; all filters have rapid phase changes as amplitude response moves from passband to stopband regions.

a note on the appendix

Appendix A was included in Mr. Leonard H. Anderson's rewritten version of this article. It discusses the "image impedance" method of network design with respect to matrix notation. Also included are normalized component values and their relationships with respect to **eqs. 14, 15,** and **16** as well as an explanation of transfer functions with regard to the derivation of **eqs. 19, 20,** and **22.**

Author DL9LX, in his original version of this article, furnished other appended material. This includes a listing of computed pi-network elements as a function of impedance-transforming ratios for various center frequencies (**Appendix B**); computed values of $z = f(m)$ from **eq. 17** (**Appendix C**); and a table of normalized network elements for various transformation ratios useful in general network design (**Appendix D**).

Interested readers may obtain a copy of author DL9LX's appendices from *ham radio* upon receipt of a large self-addressed, stamped envelope with 28 cents postage. The material in these appendices is in the author's original notation.

Editor.

references

1. George L. Matthaei, Leo Young, and E.M.T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*, McGraw-Hill Book Company, 1964.
2. S.B. Cohn, "Dissipation Loss in Multiple-Coupled-Resonator Design Filters," *Proceedings of the IRE*, August, 1959.
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Temes and Mitra, *Modern Filter Theory and Design*, John Wiley and Sons, 1973.

appendix A

image impedance

Modern network theory tends to ignore the "image method" of design. While image methods may be disregarded for complicated structures, they are valid for simple networks and quite useful at frequencies where it's difficult to obtain a purely resistive load.

Many readers will be unfamiliar with network matrix notation. Those who are familiar may be more acquainted with "A, B, C, D" notation instead of the subscripted form. Open and short-circuit impedances are given below, referred to **fig. 3**.

Port 1 input impedance with port 2 open:

$$Z_{o1} = \frac{Z_2Y_3 + 1}{Y_1Z_2Y_3 + Y_1 + Y_3}$$

Port 1 input impedance with port 2 shorted:

$$Z_{s1} = \frac{Z_2}{Y_1Z_2 + 1}$$

Port 2 input impedance with port 1 open:

$$Z_{o2} = \frac{Y_1Z_2 + 1}{Y_1Z_2Y_3 + Y_1 + Y_3}$$

Port 2 input impedance with port 1 shorted:

$$Z_{s2} = \frac{Z_2}{Z_2Y_3 + 1}$$

Multiplication of Z_{o1} and Z_{s1} or Z_{o2} and Z_{s2} will still give the same result as with matrix notation. The fact that each image impedance expression, while complex, results in an imaginary part of zero comes about by *completing* the complex division; this can be verified by completing all steps. This also applies to **eqs. (7)** and **(8)**. Real-part-only complex expressions are common to purely reactive networks.

Image impedances show the individual resonances within the network. They are synthesis tools — not an actual input impedance *when loaded*. The "missing" expression for network resonance at ω_1 of **fig. 4** is because of the $(BE - 1)$ term of **eqs. (1)** *and* **(2)**, where:

$$\omega_1^2 = 1/(C_2L)$$

normalized component values and relationships

Eqs. (14), (15), and **(16)** could have been expressed without the frequency ratio, z. In fact, z could have been omitted, but at a price: the usable bandwidth would not be optimized, since the design center frequency could not be located for minimum reflection coefficient, or "goodness of match."

Many readers are under the false assumption that a pi network is a resonant circuit with a quality factor, Q. It is simply an impedance transformation network with the *appearance* of resonance due to the sharp cutoff above design frequency. Low-side response behaves more like a conventional asymmetrical lowpass with varying *passband* response. Since many power amplifiers are still tube types with pi networks, transformation ratios will be high and the network will *appear* to peak at center frequency. As high-frequency, high-power semiconductor technology improves, transformation ratios will decrease, and the pi network will be treated as the simple lowpass filter it really is.

transfer functions

Eq. (18) may be found in reference 1, page 37, equations (2.10-1) and (2.10-5), using ABDC matrix descriptors. The transfer function is the generator-voltage-to-load-voltage ratio. The normalizing term will yield "available voltage" at the load; that is, all power from the generator is assumed dissipated in the load, $R1$; none in source resistance, $R1$.

Eq. (19) is obtained by substitution of the network arm reactances into **eq. (18)**. The steps of substitution and simplification are too long to be included here; they have been checked independently.

Input/output complex voltage ratio is obtained by deletion of (1/2) from the normalizing term. This yields a condition in which the generator is a constant-current source with a source conductance always present at the network input.

More detail on the attenuation and delay functions may be found in reference 1, sections 3.02 and 3.03. These use the *image propagation function*, γ, expressed in general terms as:

$$\gamma = \alpha + j\beta = \ln(\sqrt{A_{11}A_{22}} + \sqrt{A_{12}A_{21}})$$

where α = image attenuation in nepers
β = image phase in radians

Eq. (20) is derived by manipulation of this basic expression in terms of **eq. (19)**. Normalized group delay, **eq. (22)**, is derived from the basic group delay expression

$$T_g = \frac{d\beta}{d\omega} = \left(\frac{d\beta}{d\Omega}\right)\left(\frac{d\Omega}{d\omega}\right)$$

With the partial differential $d\Omega/d\omega = 1/\omega_0$,

$$\omega_0 T_g = \frac{d\beta}{d\Omega}$$

which yields **eq. (22)** in terms of **eq. (19)**.

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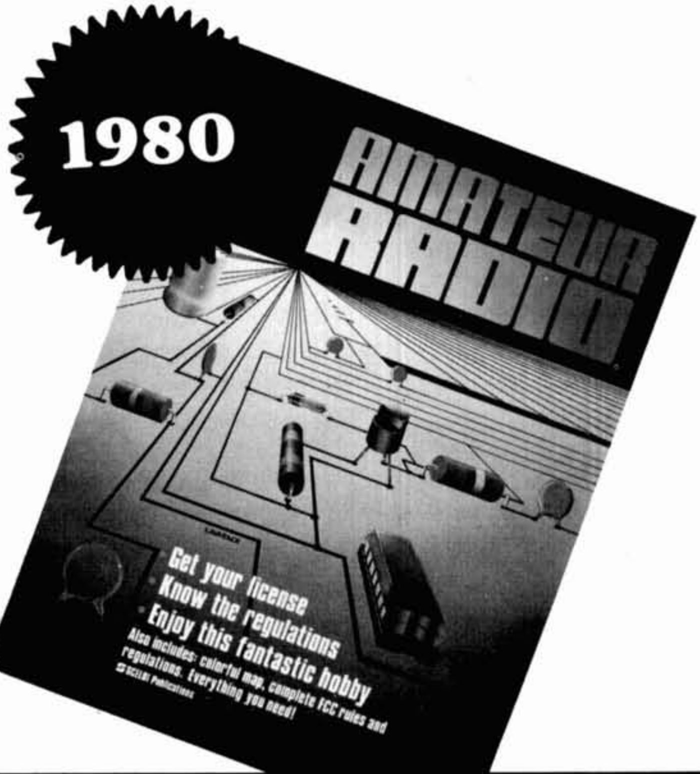

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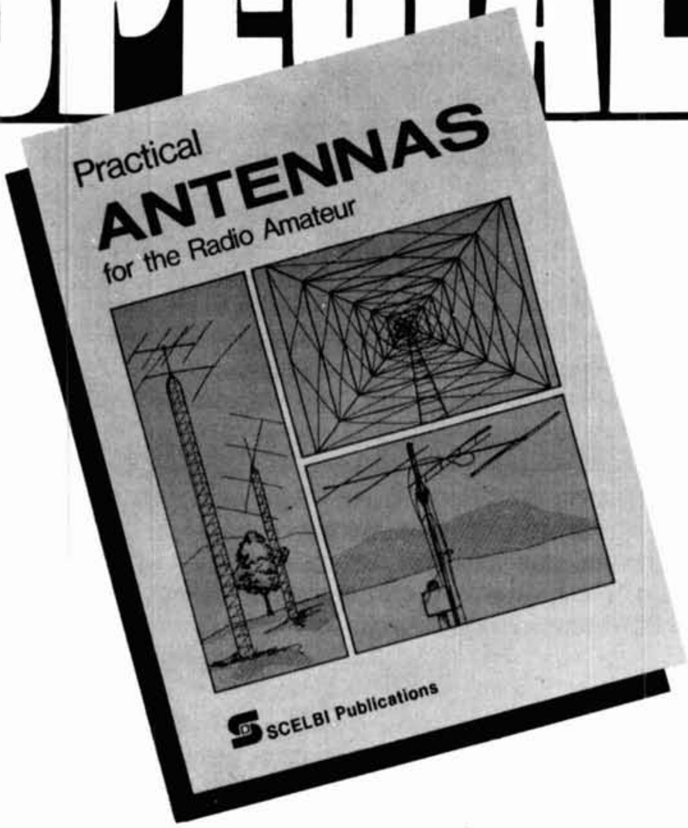

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regulated power supply

for VHF transceivers

Many of today's 2-meter fm transceivers feature 25 watts (or less) of rf power. When used in the home station, these radios generally require 13.8 Vdc at 7-8 amperes. Since I didn't have a power supply of this capacity, the only solution was to build one!

Simplicity is reliability, so I decided that the circuit couldn't be complicated; but good regulation was a requirement. The circuit shown in **fig. 1** features no-load-to-full-load (8.0 amperes) regulation of 0.2 Vdc. Also featured in the circuit is "fold back" current limiting and overvoltage protection.

ton pair. The output voltage (in this case 13.8 Vdc) is set by potentiometer R1. This sampled voltage is applied to U1, a UA723HM voltage regulator, which contains a voltage reference amplifier and an error amplifier. U1 output is applied to Q2 base to adjust the voltage at R1 to its proper value. The low-value resistor, R2, is the current "fold-back limiter." If the power-supply output should exceed 10 amperes (i.e., a short circuit), regulator U1 will bias the transistors to cutoff; thus the output voltage will drop to near zero until the short circuit condition is corrected.

Capacitor C1 and OVP form an over-voltage-protection circuit. The OVP limits the maximum output

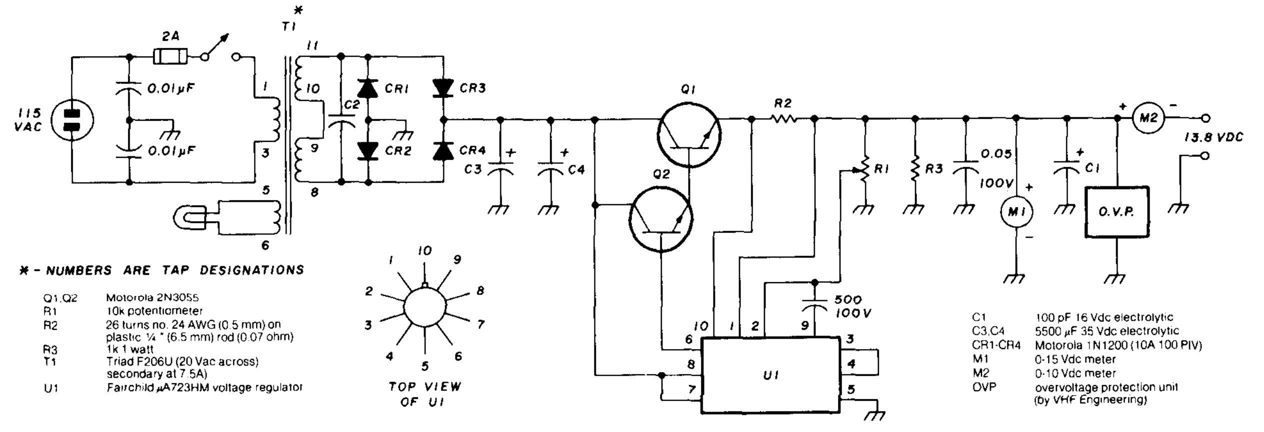

fig. 1. Schematic of the base-station power supply. Current-limiting resistor R2 is hand wound and designed for 10 amperes. If other than 10 amperes is desired, a different value must be used (see text). The overvoltage protector (OVP) is available from VHF Engineering. Q1, Q2 must be mounted on a heat sink, which must be insulated from the chassis.

circuit description

The power supply consists of a full-wave bridge rectifier with capacitor input. Any transformer-capacitor combination that produces 28 volts dc at 7.5 amperes at Q1 collector will work. Voltage at Q1 collector should not be greater than 40 Vdc, otherwise damage to U1 may result. Q1 and Q2 form a Darling-

voltage to + 15 Vdc. (The OVP unit is available from VHF Engineering.) The voltmeter and ammeter are optional, depending on your junk-box supply. The

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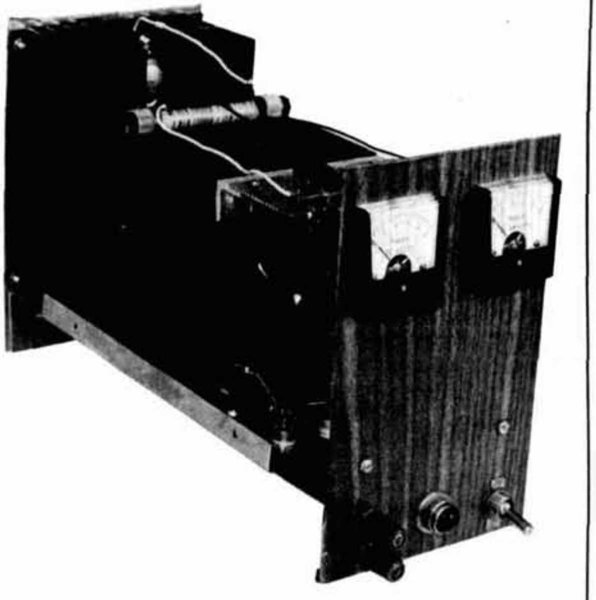


Inside view of the vhf transceiver regulated power supply.

output voltage (determined by R1) is adjustable from about 7-14 Vdc.

construction details

Component layout isn't critical; there's room for wide variation in this regard. Both Q1 and Q2 must be mounted on a suitable heat sink, which must be insulated from the chassis. "Current foldback" resistor, R2, should be wound on a plastic or Teflon rod of about ¼ inch (6.5 mm) diameter. Regulator U1 and potentiometer R1 are mounted on a piece of Vector board. If you wish to have the current limited to other than 10 amperes, a different resistor for R2 will have to be wound. To determine the new resistance:

$$R_{limit} = \frac{0.7}{I_{limit}} \tag{1}$$

Where R_{limit} is the new value resistor (ohms), and I_{limit} is the maximum desired current (amperes).

closing comments

Operating results with the power supply have been excellent. If you have a good junk box, or are a good trader at the Hamvention circuits, this supply should cost less than half that of a similar commercial model. Circuit and component layout aren't critical so you have a weekend of fun in constructing your own base-station power supply.

ham radio

More Details? CHECK—OFF Page 94

counter mixer

for the Kenwood TS-520-SE transceiver

This interface circuit between your transceiver and counter provides high accuracy

Most of the digital readout Amateur equipment available today is accuracy-specified to the nearest 100 Hz. This may mean plus or minus 100 Hz, plus or minus the time-base accuracy, and it may include programmed beat frequency oscillator (BFO) allowances. In this case the BFO output is not actually counted, and accuracy will suffer from tolerances of the BFO crystals. Because of linearity problems with the mechanical dial, the digital system will probably be more accurate across the dial, but just barely.

I've used the system described here in various forms and with different gear for the past eight years. It involves mixing the transceiver's three oscillator outputs to produce the operating frequency.[1] To measure the frequency of an incoming carrier, the low i-f is amplified and limited, then substituted for the BFO oscillator output in the mixing scheme. This option is useful for frequency-measuring tests and to calibrate the frequency counter used as a readout by checking WWV.

This particular unit (**fig. 1**) is for use with a Kenwood TS-520-SE. This transceiver has the oscillator outputs as well as a dc-supply connection available on the back panel. To provide an i-f output, another phono jack is installed on the back panel and an emitter follower is used to bring out a tap to the i-f board. See **fig. 2**.

description

Use of this unit with other gear would require providing the proper oscillator, i-f, and power-supply connections. If the rig uses a different mixing scheme, the bandpass circuit between mixers will have to be changed to the new high i-f. Also the input and output coils on the low i-f amplifier and limiter stage will need to be resonated to the different low i-f. The Heath SB-102, for instance, uses the same mixing scheme, and the same coils and capacitors may be used. The six output circuits remain the same, one for each band.

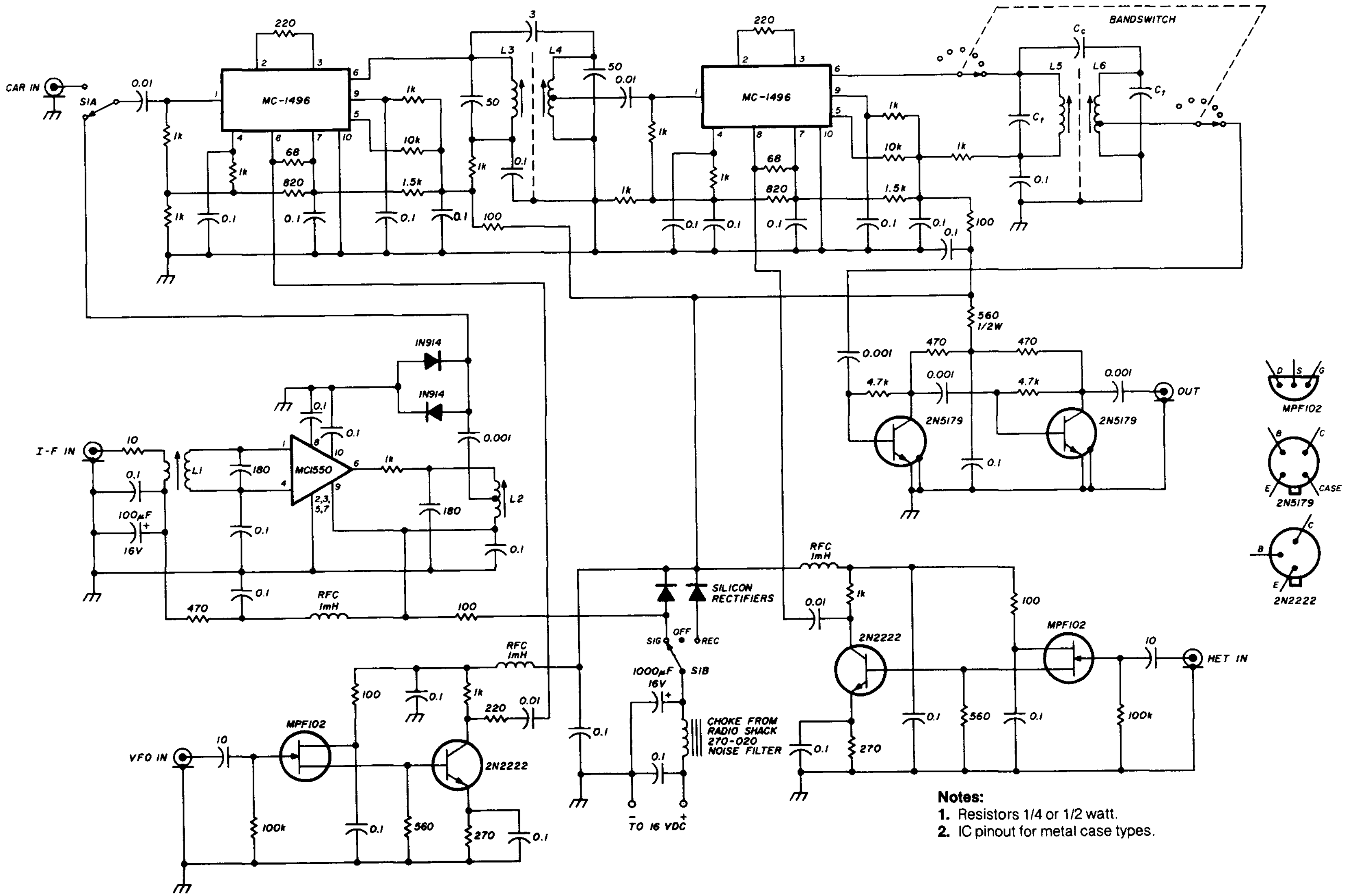
Doubly balanced mixers attenuate unwanted outputs. Separate bandpass circuits cover each frequency range. The 14-15.6-MHz range is covered by one filter as is the entire 10-meter band.

After the bandpass circuits have selected the desired bands, a two-stage broadband amplifier brings up the level to operate a frequency counter.[2] The bandpass circuits are calculated for an *R* of 10,000 ohms and were designed after an article by Anderson.[3] The requirements in this application are not strict. The filter caps were changed to the nearest standard value.

construction

Use good shielding to avoid feedback of the output signal to the receiver input. When used in the signal-measuring mode, feedback is more likely to occur. To operate in this mode, the two isolation amplifiers are necessary to prevent signals from entering the receiver on the heterodyne- and tuning-oscillator cables. Another source of feedback is the frequency counter. The counter should be enclosed in a metal case to prevent radiation of the high-level

By E.R. Lamprecht, W5NPD, Route 3, Box 207, Victoria, Texas 77901



Notes:
1. Resistors 1/4 or 1/2 watt.
2. IC pinout for metal case types.

fig. 1. Schematic of the counter mixer circuit designed for the Kenwood TS-520-SE Amateur-band transceiver. Circuit is easily adaptable to other transceivers.

signals in its input circuits. A short cable between the counter and the mixing unit is helpful, as is a short direct chassis-to-chassis connection between all units.

If you need the unit only as a digital readout of the operating frequency, the isolation amplifiers, i-f amplifier, mode switch, and the i-f connection to the transceiver may be omitted.

operation

When in the operating-frequency mode, the readout changes as the dial is moved. In the signal mode, the readout indicates the frequency of a carrier as long as it is in the passband and of sufficient strength. An S-1 signal will register in the absence of interfering signals. As an example, tuning across WWV results in an unchanging 15 000 000 on the counter. Of course, fading and multipath distortion will alter the reading for some count periods. The counter will count what it sees, and the receiver must provide a countable signal. To help ensure a clean count the receiver should have a CW filter to narrow the passband and separate modulation from the carrier. However, the system will work well with an SSB filter, especially on WWV.

tune up

The following procedure is for use with the Kenwood TS-520-SE transceiver.

1. Set mode switch to RECEIVE.

2. Adjust L3, L4 (**fig. 1**) for maximum signal at pin 1 of the second mixer (MC1496) with the transceiver tuned to about band center.

3. Adjust L5, L6 on each band for proper counter operation. Make small adjustments to allow coverage of the entire band. Observation of the counter is probably the best indication of proper tuning.

4. Set mode switch to SIGNAL.

5. Tune in a steady signal, such as that from the calibrator. Adjust L1, L2 for maximum output at the tap on L2. Reduce signal strength as needed to allow peak tuning.

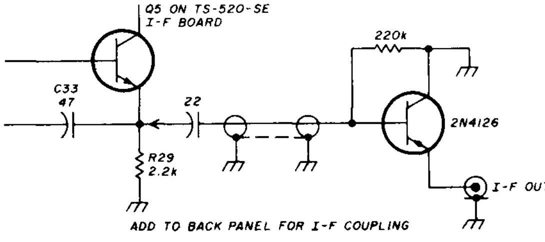


fig. 2. Emitter-follower circuit added to the back panel of the Kenwood TS-520-SE transceiver for i-f signal coupling.

table 1. Coil data for the counter mixer circuit in *fig. 1*

band (MHz)	L5-L6	L6 tap from bottom (turns)	C_t (pF)	C_c (pF)	approximate inductance (μH)
1.8-2.4	no. 32 (0.2 mm) 124 t	10	39	12	109.0
3.5-4.1	no. 32 (0.2 mm) 68 t	6	50	6	32.75
7.0-7.6	no. 32 (0.2 mm) 36 t	3	50	3	8.9
14.0-15.6	no. 30 (0.25 mm) 28 t	3	20	2	5.75
21.0-21.6	no. 30 (0.25 mm) 10 t	1	50	1	1.04
28.0-29.7	no. 30 (0.25 mm) 13 t	1	20	1	1.6

high i-f (MHz)	L3-L4	L4 tap
8.295-8.895	no. 32 (0.2 mm) 30 t	7 turns from bottom

low i-f	L1-L2	L2 tap
3.395	no. 32 (0.2 mm) 42 t	21 turns

Note: All coils are wound on slug-tuned forms 7/32 inch (5.5 mm) diameter with 1/2 inch (12.5 mm) winding space. Coil is confined to 3/8 inch (9.5 mm) nearest terminals. Windings are layer or scramble wound.

6. As a final check, observe the counter and adjust for readout of the weakest possible signal.

performance

To give an example of the capabilities of this system, I made sixteen consecutive readings of the WWV carrier at 15 MHz. A count gate time of 100 seconds was used. The readings showed a slow drift above and below the 15-MHz target frequency, with a maximum error of 0.35 Hz. Most likely, the major portion of the error was the result of the oven control, with some error due to propagation delay. The counter is controlled by a 1-MHz crystal in a proportionately controlled oven, which has been on for over five years except for power failures.

Of course I've not approached the point where I can begin to look to WWV as a source of error. But obtaining consistent readings with an error of *less than one part in fifteen million* is quite satisfying.

references

1. Mac Leish, "A Frequency Counter for the Amateur Station," *QST*, October, 1970.
2. Randall Rhea, WB4KSS, "General-Purpose Wideband rf Amplifier," *ham radio*, April, 1975, pages 58-61.
3. Leonard H. Anderson, "Top-Coupled Bandpass Filter — a Chebyshev Design," *ham radio*, June, 1977, pages 34-40.

ham radio

More Details? CHECK—OFF Page 94

the **weekender**

A simple 40-meter receiver

This article is for those who like to build their own equipment. It is a summary of a solid-state receiver that has performed very well. The receiver is the result of my experience trying to find circuits that work. It's about as simple as you can find. The receiver uses an rf stage, which certainly helps at night when foreign broadcast stations come through on 40 meters.

The most time-consuming part of the project was making the PC boards. I used black PC drafting tape to lay out the boards, which were etched in ferric chloride. Others will probably come up with a better method.

brief circuit description

Many of the receivers shown in the handbooks don't include an rf stage ahead of the mixer. This receiver was first tried using a double-tuned circuit directly into the mixer. However, the circuit was mounted in the chassis and capacitive coupling wasn't satisfactory. At night the shortwave broadcast stations came through. By using this simple rf stage, the selectivity problem was resolved. You can use whatever toroid forms you can find. The T-80-2 forms are probably too large, but these are what I have used. They have a red core and are about 1 inch (25.5 mm) in diameter. Resonance can be checked by holding a grid-dip oscillator to the hot end or by placing a turn or two of wire around the core and checking with the grid-dip oscillator. The input coil was mounted on the chassis topside and the output coil was mounted underneath. **Fig. 1** shows the circuit.

mixer

The mixer output coil was tuned to 5.5 MHz. It is wound on a 3/8-inch (9.5-mm) ceramic slug-tuned coil using a fixed 100 pF cap for tuning. L5 (about 13 turns) on the bottom of the coil feeds directly into the Swan crystal filter.

variable-frequency oscillator

The VFO tunes 12,500-12,800 kHz to cover the 40-meter band. A small cap with two rotary and two fixed plates came out to about 35 pF, which just covers the band. You can pull plates out after the set is going to obtain desired bandspread. Coil L6 was wound on a 3/8-inch (9.5-mm) ceramic slug-tuned coil form. It was wound with silver-plated wire, nylon covered. The two 500-pF caps from the MPF-102 gate to ground are silver micas. A 9-volt zener stabilizes the MPF-102 drain. The two 2N2222 stages are buffers to reduce pulling effect on the VFO and to obtain the 1.5 V rms to feed the mixer.

i-f stage

Only one i-f stage was necessary for this receiver. L7 is another 3/8-inch (9.5-mm) diameter slug-tuned coil adjusted to 5.5 MHz. It's coupled into the product detector with a 0.001 μF cap.

beat-frequency oscillator

The BFO can be varied about 1 kHz with the 0-30 pF variable cap to adjust the SSB tone. The output coil, L8, is another slug-tuned coil. The 10 pF coupling cap to the product detector should be sufficient, but you can try other values if there isn't enough signal injection.

construction

I built the receiver starting from the audio stage and worked backward to the front end. I used 2-inch (50-mm) square PC board. I laid out the circuits using black drafting tape and etched the boards with ferric chloride. Parts were mounted on a 7 × 11-inch (178.5 × 280.5-mm) aluminum chassis.* The VFO was mounted in a partition topside, which was about 3-inches (76.5-mm) square. I used a VTVM and rf probe for tune up.

performance

I'm amazed at the performance of this little receiver. I frequently operate it from a battery supply during park picnics and wonder how I ever got along without it. Don't ask me how to make a transceiver for CW; so far I've not been able to make a mixer to drive a transmitter section.†

*Parts that may be useful for construction are available from Radiokit, Box 411H, Greenville, New Hampshire 03048.

†See John Keith's article, "40-Meter Transceiver for Low-Power Operation," *ham radio*, April, 1980, page 12.

By Ed Marriner, W6XM, 528 Colima Street, La Jolla, California 92037

L1 35T NO.26 (0.3mm) TAPPED 10T BOTTOM ON MICROMETALS T-80-2 FORM (RED CORE)
L2 3T TAPPED FROM GROUND END OF L1
L3 SAME AS L1
L4 25T NO.30 (0.25mm) ON 3/8 IN. (9.5mm) CERAMIC FORM
L5 13T NO.30 (0.25mm) ON BOTTOM OF L4 (FILTER INPUT)

L6 7T NO.26 (0.3mm) ON 3/8 IN. (9.5mm) CERAMIC FORM
L7 25T NO.30 (0.25mm) ON 3/8 IN. (9.5mm) CERAMIC FORM
L8 SAME AS L7

NOTES:
1. ALL RESISTORS ARE 1/2 WATT
2. ALL CAPS 16-25 WVDC

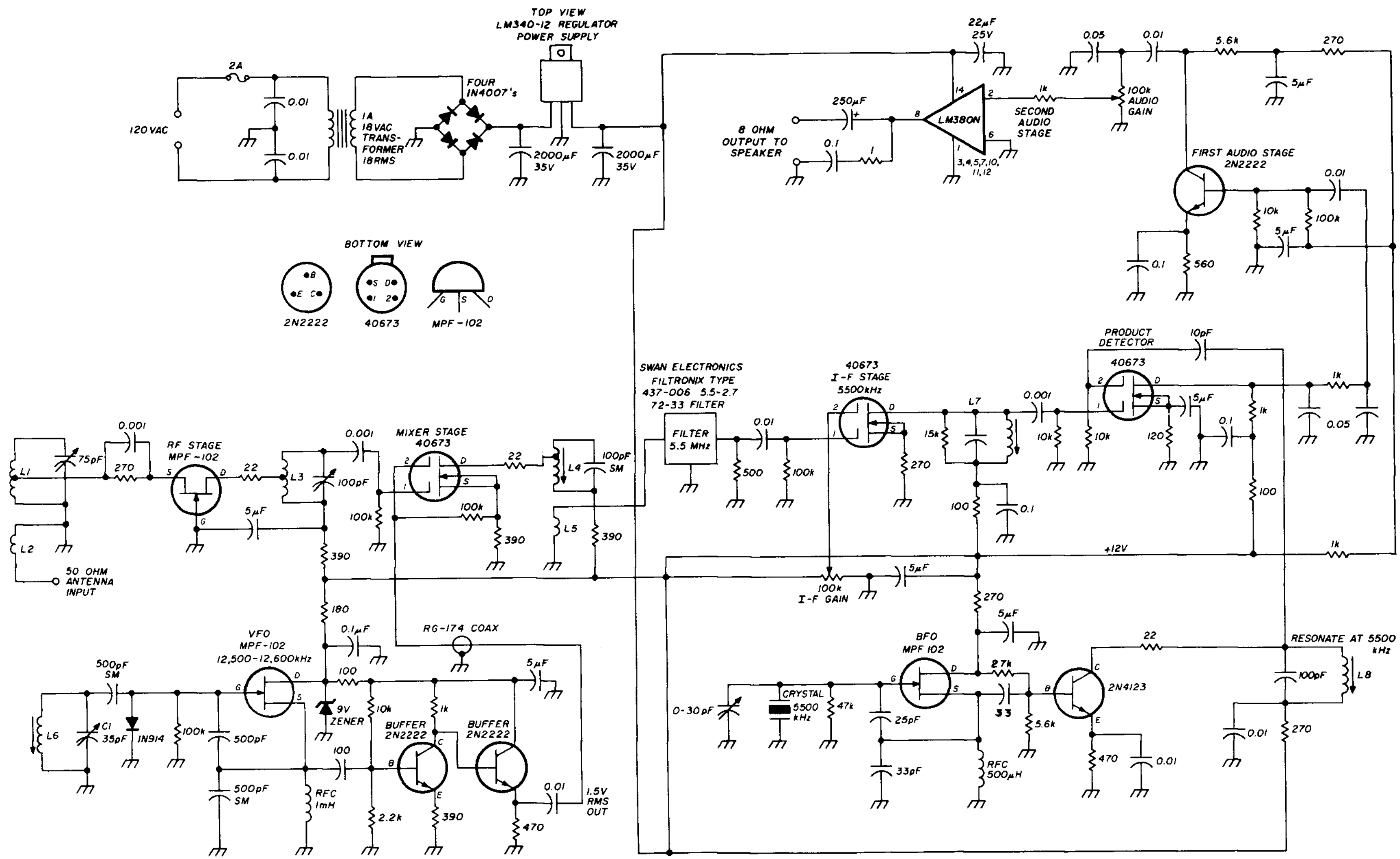


fig. 1. Schematic of the solid-state 40-meter receiver. The MPF-102 rf stage helps selectivity, especially when foreign broadcast stations appear at night. PC boards are used, which were made using the tape-and-etching method. Low-frequency circuits could be mounted on perf boards. The 5.5-MHz filter was picked up at local flea market.

comments

(Continued from Page 6)

the product of baud speed times K, K being a factor depending on the "goodness" of a circuit. The baud speed of Amateur Hellschreiber is 122.5 which multiplied by a K factor of 3 gives a bandwidth of 367.5 Hz. The K factor of 3 comes into the picture because it has long been recognized that a square wave and its third harmonic is perfectly acceptable for normal communications. The bandwidth of a 45.5-baud, 170-Hz shift RTTY computed according to CCIR is 245 Hz.

K6KA is correct in his criticism of the Chinese Hell-Fax signal, lately on 14140 and believed still to be working in the Region 2, 80-meter band. But this is a different system with a baud speed of somewhere in the region of 400, and observedly with little or no attempt at pulse shaping; some channels are even FSK with 800 Hz shift! They are certainly wide band and not to be compared with the Amateur 'hell' in Europe.

Finally, I hold no brief for the Hellschreiber system as such but, as I worked with the system throughout most of its active life and am fully conversant with its advantages and shortcomings, I thought I'd like to put the matter straight.

Stanley A.G. Cook, G5XB
Radio Society of Great Britain
Reading RG4 9BP, England

PCB "threat"

Dear HR:

I noted in "Presstop" in May, 1980, issue your warning regarding the "potentially deadly threat" existing in the form of PCBs or polychlorinated biphenyls, and should like to thank you for bringing the attention of the fraternity to this material.

However, I should like to point out that the PCB hazard has been, like many others, vastly overrated by media exposure. PCB in massive doses fed to lab test animals has been shown to produce malignant tumors, and repeated applications to the skin of mice has indicated some potential as a dermal carcinogen.

PCB came to the attention of health authorities through two major instances. One was in Japan, where, by error, it was substituted for fish oil in food packaging. The second instance occurred in the U.S., where, in error, it was added in place of vegetable oil, to cattle feed. In both cases severe illness resulted from the consumption of the PCB-contaminated food.

Occasional handling of PCB has shown no deleterious effects on humans. In fact, many Amateurs who are also Industrial Electricians will testify that they have had their hands in it innumerable times, and in big transformer work have literally been immersed in it, with no visible short- or long-term effects.

The properties of PCB, which make it such an excellent electrical insulating fluid, are the qualities that cause the physical and ecological problems. It is heavier than water, non-conductive, and will not break down or decompose at temperatures under 2000°F. In fact it requires the full 3500°F heat of a cement kiln to break it down. Under normal conditions, it is not bio-degradable. This is its biggest hazard. Once spilled, it remains in the ground indefinitely, being propagated by natural ground waters, absorbed unchanged by plants, which are then eaten by animals.

Incidentally, if you have a tube-type television set or refrigerator more than ten years old, fluorescent lights, or a car with brake fluid or hydraulic fluid more than ten years old, you probably have another source of PCB.

Amateurs, building or buying dummy loads without transformer oil, and having gone to their local utility for a gallon of "good, hi-temperature transformer oil" have received a gallon of PCB. All the above is presented to show that PCBs have been around and done a good job for years, and pose no "potentially deadly threat" in the quantities hams use.

PCBs can be differentiated from mineral or vegetable transformer oils by the following means:

1. The smell of PCBs is somewhat similar to that of moth balls. Ordinary vegetable or mineral transformer oils smell like oil.
2. Pure PCB is heavier than water, and a drop dropped into a bottle of water will sink. Ordinary transformer oil will float on water.

If you have a PCB-filled dummy load that has a leak or a filter capacitor filled with PCBs that shows a leak around the bushings these leaks can be easily repaired using "Weldfast 220" or equivalent epoxy. First clean off all PCB seepage with a good solvent; "Xylene" will do fine. Wear rubber gloves to protect you from both the Xylene and the PCBs, and store contaminated wipers in a sealable can. Mix the epoxy, smear over and around the leak, and let it set. Job done.

Clean up any spilled PCBs well with Xylene and rags. Store rags, rubber gloves, and all contaminated materials in a sealed can. The whole object of the game is to keep the PCB from getting directly into your food and from getting into the food chain via the earth and ground water. A call to your public utility will provide a safe method of disposing of your PCB wastes.

Above all, remember PCBs are a hazardous substance, not a "deadly threat." Inspect your capacitors and ensure they are not leaking PCBs. If they are, repair the leaks and clean up the spills properly, or remove the bad component and clean up the spill properly. Put all contaminated materials in a sealed can, wash your hands well, call your public utility and make the necessary disposal arrangements. Don't panic and throw them in the garbage. If you do, you can be sure of getting your share of them back through the food chain.

Tom Ruynon, VE5UK
Saskatoon, Saskatchewan

speed of light

Dear HR:

In his amusing exposé, W7ITB has drawn before our eyes the frightening picture of all our emissions eventually coming back upon us because of the speed of light becoming negative by the 2100th century.

May I draw your attention to a hint at a much earlier date of this reversal? In the May, 1976, issue of *ham radio* on page 31 VK2ZTB cites the speed of light to be 290500 km s^{-1}. Should we all possibly have overlooked this dramatic 3 per cent decrease or could it be a specific development of our fellow Australian hams working towards a "light boomerang?"

Gunter Hoch, KL6WU
Darmstadt-Eberstadt

surplus tubes

Dear HR:

I agree with Bill Orr's suggestion in December *ham radio* that one should test surplus tubes as soon as possible after their receipt. However, I object to his blanket condemnation of mail-order surplus houses, that customers have a "fat chance" of getting a refund or replacement for a defective surplus tube purchased via mail order.

We at Fair Radio Sales have been selling used and unused surplus tubes to Radio Amateurs for many years. As a matter of policy, we replace an unsatisfactory tube or refund its price, provided the customer's claim is made within ten days or so of the tube's receipt.

Like any reputable business, we feel that we have the responsibility to make every reasonable effort to satisfy our customers. Undoubtedly there are other surplus dealers who share this commitment to their customers.

Orr's remark was a disservice to the conscientious surplus mail order companies — many of which advertise in *ham radio*.

George Sellati
Fair Radio Sales
Lima, Ohio

short circuits

Yagi antenna design: performance calculations

The caption for **Table 1** on page 25 of the January, 1980, issue of *ham radio* should read: Element reactance for different wavelength-to-radius ratios, *K*. The caption for **fig. 2** should read: Graph showing the relationship between the wavelength-to-radius ratio, *K* . . . .

coaxial-line transformers

W6TC reports that on page 17 of the February, 1980, issue of *ham radio*, eight lines below the heading "50/200 ohm transformers," the text should read: ". . . two pairs of RG-58 A/U [not RG-59 A/U] cable."

Touch-Tone decoder

The schematic of the *Touch-Tone* decoder that appeared in the February, 1980, issue of *ham radio* (page 37) should show a crystal, not a resistor, at pin 2 of U3. Pins 3 and 5 of U10 are tied to +12 volts, with pin 4 tied to D1 of the sequential control outputs. The price of the complete kit is $140, assembled and tested $160, from James Wyma, WA7DPX, 12952 Osborne St., Arleta, California 91331.

capacitance measurement

Fig. 1 of the *Capacitance Measurement* article that appeared on page 44 of the April, 1980, issue of *ham radio* appeared with the **HI** and **LO** positions of S1B inadvertently reversed. The open contact, which should be marked **HI**, is grounded.

experimental high-gain phased array

In KL7IEH's high-gain phased array article, which appeared on page 44 of the May, 1980, issue of *ham radio*, the reflector elements in **figs. 1** and **5** should be broken in the middle, as should the first director in **fig. 1**. In **fig. 4**, elements D_1 and D_2 are reversed.

More Details? CHECK—OFF Page 94

the ham notebook

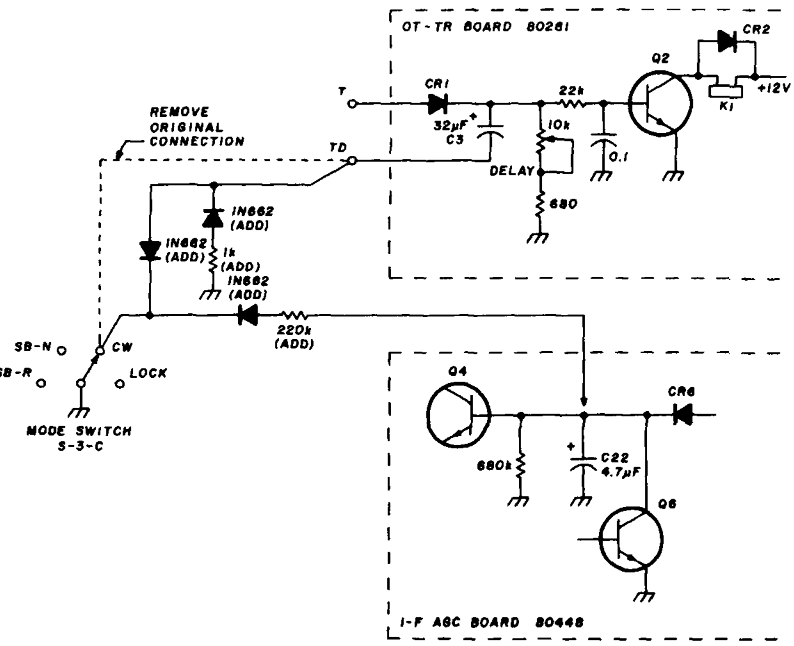
another improvement for the Ten-Tec Omni-D CW agc

The improved CW agc for the Ten-Tec Omni-D transceiver by Doug McDougall (*ham radio*, January, 1980, page 88) adds a needed feature. However, McDougall's approach using a new switch deck didn't appeal to me because it requires a cooling-off delay and material not in my junk box.

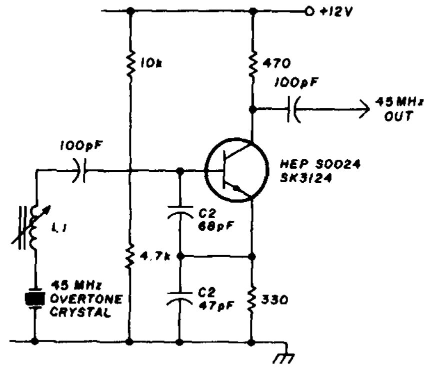
An examination of the circuit suggested that the CW-agc change could be made using diodes instead of a new switch deck. **Fig. 1** shows the circuit installed in my radio. The original mute relay circuit seems to be unaffected by the change, and the improved CW agc works as described by McDougall's modification. The 1-k resistor and diodes allow C3 to charge and discharge only in the CW mode.

My thanks to McDougall and *ham radio* for publishing the article, which led to my circuit.

John Bunting, W4NET



overtone crystal oscillator



I found WB2EGZ's article, "Quartz Crystals — Gems for Frequency Control," *ham radio*, February, 1979, page 37, very interesting. I had a transmitter on my bench with a crystal fault. The transmitter had been modified so that it would operate on 144 MHz; however, the oscillator wouldn't operate in the overtone mode of the crystal. While the article by WB2EGZ was first-class, this is one type of oscillator he didn't cover, although he had a variation of it in his **fig. 4**. So it was a case of finding out what was happening in my transmitter.

I found that the critical part in the circuit (**fig. 2**) was inductance L1 in series with the crystal. With the crystal shorted out L1 will resonate with C1, C2, C3 at or near the crystal overtone frequency when the oscillator is operating as an overtone circuit. When the short circuit across the crystal is removed, the crystal will operate in its overtone mode. If L1 is tuned to a much lower frequency, the crystal will operate in its fundamental mode.

The overtone circuit appears to operate as a crystal-locked Clapp oscillator; with the crystal shorted, it becomes a Clapp oscillator.

B.E.G. Goodger, ZL2RP

impedance of a random-length antenna

Several parameters must be established to provide input for this calculation. There are the wire size and its height above ground, from which the characteristic impedance of the wire as a length of transmission line parallel to the ground is computed. The height above ground must also be expressed in terms of wavelength to establish the radiation resistance of a typical half-wave dipole. This information is found in many texts from Terman, (*Radio Engineers' Handbook*) to publications such as the ARRL *Antenna Book* or *Handbook*.

A change from one band to another will have a major effect on this last parameter. So as a starting point, to illustrate the method, I decided to consider a No. 12 (2.1-mm) wire, 37.5 feet (11.4 meters) above ground, which yields the convenient characteristic impedance of 600 ohms. The 20-meter band was chosen, where the height of 0.57 wavelength results in a radiation resistance to a half-wave dipole of 68 ohms. This number was divided by two, assigning 34 ohms to each quarter wavelength.

After all these preliminaries we have two numbers: 600 ohms and 34 ohms as the wire characteristic impedance; and we have the radiation resistance of a quarter wavelength (68 ohms). We now shift our attention to the Smith chart (**fig. 3**). If the length of our wire is zero, its impedance must be infinite. This is plotted on the Smith chart as point 0. The 34-ohms radiation resistance of a quarter wavelength, when normalized to 600 ohms, is 0.05666, which was rounded off to 0.057 and plotted on the real axis of the Smith chart as point 1. Points between zero and a quarter wavelength lie on a spiral connecting these two points, which, for simplicity, was approximated by a semicircle centered on the real axis and passing through those two points lying in the left-hand, or capacitive, side of the chart.

The VSWR of point 1 is the inverse of 0.05666, or 17.65. From this number, the reflection coefficient, Γ, is computed as 0.893.*

The wire length is now increased to two quarter wavelengths. The reflection coefficient is now the second power of 0.893 or 0.797, corresponding to a VSWR of 8.85. This is plotted as point 2, and is connected to point 1 by a semicircle, centered on the real axis as before, but this time lying in the right-hand half of the chart because it is inductive.

In a like manner, successive values for Γ are computed, as the length of wire is increased by successive quarter wavelength additions, and connected by semicircles, as before.

Although this method is only an approximation, it does afford considerable insight into the characteristics of a long- or random-length wire antenna. For example, suppose you're considering erecting a full-wave antenna fed a quarter wavelength from one end. The quarter-wave end section will have a radiation resistance of about 34 ohms, while the three-quarter-wave end section will present a radiation resistance of about 100 ohms. Your chances of balancing your feed system to prevent feedline radiation have just gone out the window! It will still radiate effectively, but the opportunity for complications is enhanced.

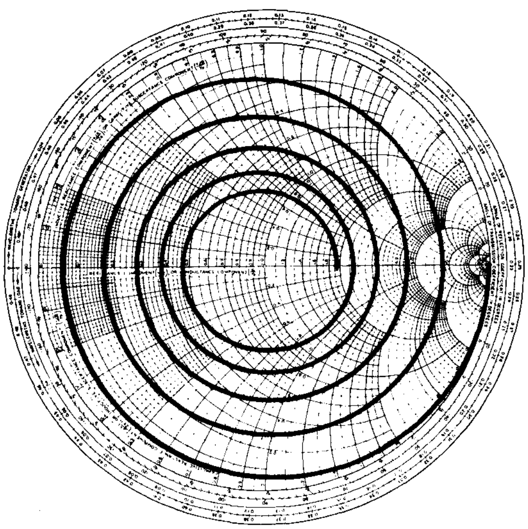
Using a chart of this type is simple enough. You might become confused with the markings of wavelength on the circumference of the Smith chart if you're not careful. Suppose, for example that you want to make an educated guess about this wire at a length of, say, 1.2 wavelengths. This point would lie between four and five quarter wavelengths and would be located by a radius from the center of the chart to where the inner scale reads 0.05 wavelength. Unfortunately, our starting point (0) is marked 0.25 wavelength, rather than 0, and you must be aware of the possible foul-up. A straight edge marking out this radius intersects the spiral at about 0.296-j0.296. Multiplying these values by 600 ohms gives 177.6-j177.6 as our estimate of what we would have to match to load such an antenna. All of this may seem merely academic, but it should put us in the ball park when it comes to designing a matching network.

Henry S. Keen, W5TRS

* $$\Gamma = \frac{VSWR - 1}{VSWR + 1} \qquad (1)$$

$$Z_0 = 138 \log \frac{4h}{d}$$

where h = height above ground
d = diameter of wire



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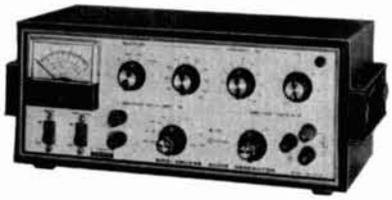
Sine-square wave Audio Generator



$185.00

- 1 Hz-100 kHz frequency range
- 0.003-10 Vrms sine wave output (10k Ω load)
- 0.003-1 V sine wave output (600 Ω load)
- Meter calibrated in volts and dB
- −62 to +22 dB ranges
- 0.1-10 V square wave output (2000 Ω load)
- 50 nanosecond risetime
- 5.13" H x 13.25" W x 7.0" D

SG-5218 .... 185.00
($2.85 shipping & handling)

Dual-trace DC-10 MHz Oscilloscope

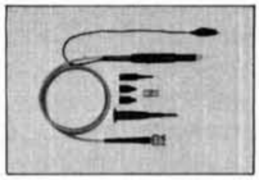
$650.00

- Two vertical input channels with 10 mV/cm sensitivity
- 11-step attenuator for 10mV/cm to 20V/cm deflection factors
- 19-step horizontal time base from 0.2 sec/cm to 0.2 usec/cm
- Vertical accuracy within 3%
- X5 horizontal expansion
- Calibrated 1V peak-to-peak square wave signal
- 35 ns vertical rise time
- Automatic triggering
- 120/240 VAC, switch-selectable
- 6.9" H x 12.9" W x 19.3" D

SO-4550 .... 650.00
($5.50 shipping & handling)

Combination x1, x10 Scope Probe

$29.95



- Switch-selectable x1 and x10 attenuation at probe tip
- Center (ground) switch position allows quick zero level check
- DC to 15 MHz (x1) and DC to 80 MHz (x10) bandwidths
- 4.0 nS (x10) rise time
- Insulating tip, BNC tip adapter, IC tip, insulated compensation capacitor adjustment tool, vinyl case

PKW-105 .... 29.95
($1.60 shipping & handling)

Order TOLL-FREE: 800-253-0570

8:00 AM to 8:00 PM Eastern Time M-F. Sorry, toll-free service not available in Alaska, Hawaii or Michigan. Call 616-982-3411, 24 hours a day, seven days a week. TLX: 72-9421

HEATH Zenith Instruments

For information on other Heath/Zenith Instruments write: Dept. 122-698, Benton Harbor, MI 49022

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NEW products

RTTY Reader

A new radioteletype code reader has been introduced by Microcraft Corporation for SWLs, Novices, and veteran radio operators. It is completely self-contained, featuring an eight-character moving LED display, separate, active, mark and space filters, and tuning LEDs. All text characters — letters, numbers, and punctuation, are shown sequentially on the display. It features an extremely versatile decoding system capable of handling 170, 425, and 850 Hz FSK with RTTY speeds of 60, 67, 75, and 100 WPM Baudot and 100 WPM ASCII. All that is required for operation is to connect it to the loudspeaker of a communications receiver — no CRT is needed. It is compact, measuring 7.375 x 5.75 x 3.375 inches (18.73 x 14.6 x 8.57 cm), and weighs 4 pounds (1.8 kg). The kit version, RRK, recommended only for intermediate to advanced builders, costs \$189.95. The wired and tested version, RRF, is \$269.95. An optional 220 Vac, 50/60 Hz transformer is available for an additional \$6.00. Shipping and handling in the Continental United States is an additional \$5.00. Shipments are made worldwide and requests for quotes are invited. Contact Microcraft Corporation, P.O. Box 513, Thiensville, Wisconsin 53092.

Spectrum Communications vhf transceivers

Spectrum Communications Corporation has introduced its Professional Communications Line of base-station and mobile transceivers.

The PCL250 mobile unit and the PCL300 base station are 30-watt (nominal) transceivers covering the 136-174 MHz range. Features include: excellent receiver sensitivity (0.3 μV), very wide receiver dynamic range for superior intermodulation rejection, eight-pole crystal filters and four-pole ceramic filters, "super-rugged" housing (1/8" aircraft aluminum), six channels, and high quality design, components, and workmanship throughout.

The PCL300 base station has a built-in ac power supply with optional auto-switchover to dc battery power, front-panel status indicator lights, optional receiver scanning function, pre-amp desk or hand-held microphone, and wood grain housing.

For further information, write Spectrum Communications Corp., 1055 W. Germantown Pike, Norristown, Pennsylvania 19401.

heavy-duty line filters

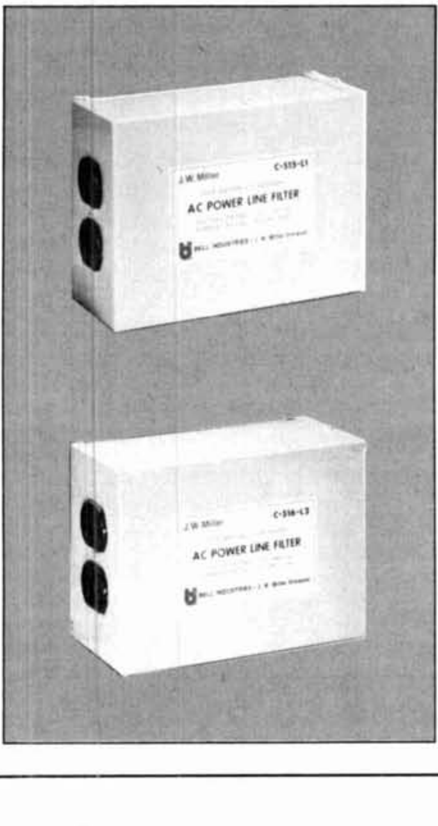
New heavy-duty, ac power-line filters that handle up to 15 amperes have been introduced by the J.W. Miller Division of Bell Industries in Compton, California.

Sensitive equipment can be protected from interference by these filters. They are ideal for preventing power-line interference from virtually all sources such as copying machines, small computers, and appliances.

Model C-515-L1 (110-120 Vac) and Model C-516-L2 (220-240 Vac) five-section LC-network filters provide 50 dB attenuation or better for 500 kHz to 300 MHz.

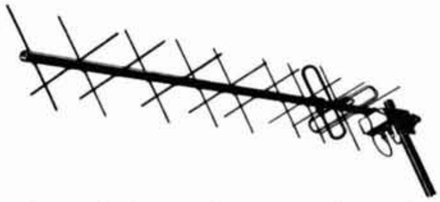
In use, the equipment causing the interference is plugged into the filter, and the filter is plugged into an ac outlet. Also, the filter can be plugged in between equipment to be protected and the ac power.

Additional information may be obtained from Jerry Hall, J.W. Miller Division, Bell Industries, 19070 Reyes Avenue, Compton, California 90221.



KLM circularly polarized antenna

KLM's new 420-450-18C antenna will bring all the advantages of circular polarity to the uhf bands. Flutter and multipath fading can now be reduced for optimized satellite and terrestrial communications. For efficient use of the new Phase III-B OSCAR satellite, the 18C will be a virtual necessity. Its broadband character, in the KLM tradition, also permits the 18C to meet the critical needs of DXers, ATVers, and the 440-450 FM group. According to KLM, gain is conservatively rated at 12 dB.



New design and construction techniques have been used to meet the strict electrical requirements for good circularity at uhf. The reflector and director elements of the 420-450-18C pass through the center of the boom and are secured with integral polyethylene insulators and locking snap-rings. Folded dipoles are used for the driven elements. They provide excellent bandwidth and maintain perfect element symmetry. Circularity is held within 1 dB, 430-440 MHz, and 3 dB, 420-450 MHz.

Electrically, the 18C has nine elements in the vertical plane and nine, offset by 1/4 wave, in the horizontal plane. Feed impedance of each section is 50 ohms balanced. Two coaxial baluns are supplied.

The optional CS-2 Circularity Switcher, mounted on the antenna, features fingertip control of circularity (RHC-LHC) in the shack and a built-in power divider, for single-feedline convenience.

Price of the 420-450-18C is $59.95; the CS-2 is $49.95. For more information on the 18C and other Phase III related products, contact KLM Electronics, P.O. Box 816, Morgan Hill, California 95037.

More Details? CHECK—OFF Page 94

microprocessor controlled repeater

A line of repeaters covering the 144, 220, and 450 MHz bands has just been introduced by Micro Control Specialties. The new Mark 3CR repeaters combine all the features of the popular Mark 3C repeater controller plus transmitter, receiver, and power supply in a rack-mount cabinet, ready for immediate service. The microprocessor-based repeater provides 39 tone-accessible functions, including autopatch, autodial, redial, reverse patch, external outputs, and secure control-operator commands. Crystal-controlled digital tone-decoding assures stable and reliable function access. To keep users informed of its status, the repeater generates thirteen different Morse messages, several of which are custom programmed to user specifications. Basic repeater operations such as timeout, tail, and ID timing are also directed by the microprocessor so the repeater can discriminate intelligently against noise and "kerchunkers." Several operations can be modified remotely by command functions.

The repeater receiver uses dual-gate MOSFETS in both rf amplifier and mixer stages for high sensitivity (20 dB quieting with only 0.25 μV of input signal) and freedom from overload in the presence of 0.5-volt signals. Crystal filtering and double conversion are both used to obtain 65 dB rejection of off-frequency signals.

Transmitter output is 2 watts, but optional amplifiers are available to increase the power output to any desired level. Transmitter and receiver oscillators are temperature compensated to meet commercial frequency-stability requirements. The audio circuits combine generous amounts of feedback with symmetrical clipping for virtually transparent audio quality.

For further information, write Micro Control Specialties, 23 Elm Park, Groveland, Massachusetts 08134.

More Details? CHECK—OFF Page 94

MHz electronics

MOTOROLA Semiconductor The RF Line

MRF458 $20.68

NPN SILICON RF POWER TRANSISTOR

. . . designed for power amplifier applications in industrial, commerical and amateur radio equipment to 30 MHz.

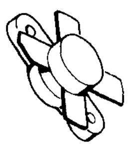
- Specified 12.5 Volt, 30 MHz Characteristics –
  - Output Power = 80 Watts
  - Minimum Gain = 12 dB
  - Efficiency = 50%
- Capable of Withstanding 30:1 Load VSWR @ Rated P_{out} and V_{CC}

MRF454

$21.83

NPN SILICON RF POWER TRANSISTORS

. . . designed for power amplifier applications in industrial, commercial and amateur radio equipment to 30 MHz.



- Specified 12.5 Volt, 30 MHz Characteristics –
  - Output Power = 80 Watts
  - Minimum Gain = 12 dB
  - Efficiency = 50%

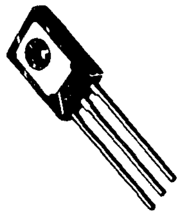
MRF472

$2.50

NPN SILICON RF POWER TRANSISTOR

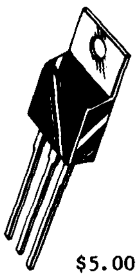
. . . designed primarily for use in large-signal output amplifier stages. Intended for use in Citizen-Band communications equipment operating at 27 MHz. High breakdown voltages allow a high percentage of up-modulation in AM circuits.

- Specified 12.5 V, 27 MHz Characteristics –
  - Power Output = 4.0 Watts
  - Power Gain = 10 dB Minimum
  - Efficiency = 65% Typical



MRF475

NPN SILICON RF POWER TRANSISTOR



$5.00

. . . designed primarily for use in single sideband linear amplifier output applications in citizens band and other communications equipment operating to 30 MHz.

- Characterized for Single Sideband and Large-Signal Amplifier Applications Utilizing Low-Level Modulation.
- Specified 13.6 V, 30 MHz Characteristics –
  - Output Power = 12 W (PEP)
  - Minimum Efficiency = 40% (SSB)
  - Output Power = 4.0 W (CW)
  - Minimum Efficiency = 50% (CW)
  - Minimum Power Gain = 10 dB (PEP & CW)
- Common Collector Characterization

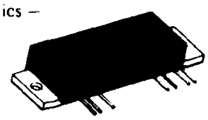
MHW710 - 2

$46.45
440 to 470MC

UHF POWER AMPLIFIER MODULE

. . . designed for 12.5 volt UHF power amplifier applications in industrial and commercial FM equipment operating from 400 to 512 MHz.

- Specified 12.5 Volt, UHF Characteristics –
  - Output Power = 13 Watts
  - Minimum Gain = 19.4 dB
  - Harmonics = 40 dB



- 50 Ω Input/Output Impedance
- Guaranteed Stability and Ruggedness
- Gain Control Pin for Manual or Automatic Output Level Control
- Thin Film Hybrid Construction Gives Consistent Performance and Reliability

Tektronix Test Equipment

B	Wideband High Gain Plug In	$ 51.00
CA	Dual Trace Plug In	150.00
K	Fast Rise DC Plug In	63.00
N	Sampling Plug In	200.00
R	Transistor Risetime Plug In	116.00
W	High Gain Differential Comparator Plug In	283.00
TU-2	Test Load Plug In for 530/540/550 Main Frames	50.00
1A2	Wideband Dual Trace Plug In	216.00
1S1	Sampling Unit With 350PS Risetime DC to 1GHZ	730.00
2A61	AC Differential Plug In	133.00
3S3	Dual Trace Sampling DC to 1GHZ Plug In	250.00
3S76	Dual Trace Sampling DC to 875MHZ Plug IN	250.00
3T77A	Sampling Sweep Plug In	250.00
3L10	Spectrum Analyzer 1 to 36MHZ Plug IN	1000.00
50	Amplifier Plug In	50.00
51	Sweep Plug In	50.00
53B	Wideband High Gain Plug In	25.00
53/54B	Wideband High Gain Plug In	45.00
53/54C	Dual Trace Plug In	112.50
53/54D	High Gain DC Differential Plug In	38.00
53/54G	Wideband DC Differential Plug In	68.00
53/54L	Fast Rise High Gain Plug In	68.00
84	Test Plug In For 580/581 Main Frames	75.00
107	Square Wave Generator .4 to 1MHZ	48.00
RM122	Preamplifier 2Hz to 40KHZ	63.00
123	AC Coupled Preamplifier	25.00
127	Power Supply For 2 Plug In's	148.00
131	Current Probe Amplifier	50.00
184	Time Mark Generator	363.00
R240	Program Control Unit	150.00
280	Trigger Countdown Unit	84.00
455	Portable Dual Trace 50MHZ Scope	2000.00
465	Portable Dual Trace 100MHZ Scope	2500.00
503	DC to 450KHZ Scope Rack Mount	250.00
535A	DC to 15MHZ Scope Rack Mount	263.00
543	DC to 33MHZ Scope	300.00
561	DC to 10MHZ Scope Rack Mount	150.00
561A	DC to 10MHZ Scope Rack Mount	200.00

Scopes with Plug-in's

567	Digital Readout Scope with a 6R1A Digital Unit and a 3S3 Dual Trace DC to 1GHZ Sampling Plug In and a 3T77A Sweep Plug In.	750.00
561A	DC to 10MHZ Scope with a 3S76 Dual Trace DC to 875MHZ Sampling Plug In and a 3T77A Sweep Plug In. Rack Mount	600.00
565	DC to 10MHZ Dual Beam Scope with a 2A63 Diff. and a 2A61 Diff. Plug In's	900.00
581	DC to 80MHZ Scope with a 82 Dual Trace High Gain Plug In	650.00
661	Sampling Scope with a 5T3 Timing Plug In and a 4S2 Dual Trace DC to 3.5GHZ Sampling Plug In.	575.00

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2E26	$ 5.00	4CX350FJ	$116.00	6146W	12.00
3-500Z	102.00	4CX1000A	300.00	6159	10.60
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4CX250K	113.00	5894/A	42.00	8295/PL172	328.00
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4CX300A	147.00	6146A	6.00	8560A/AS	50.00
4CX350A	107.00	6146B/8298A	7.00	8908	9.00
				8950	9.00

More Details? CHECK—OFF Page 94

MICROWAVE COMPONENTS

ARRA

2416	Variable Attenuator	$ 50.00
3614-60	Variable Attenuator 0 to 60dB	75.00
KU520A	Variable Attenuator 18 to 26.5 GHz	100.00
4684-20C	Variable Attenuator 0 to 180dB	100.00
6684-20F	Variable Attenuator 0 to 180dB	100.00

General Microwave

Directional Coupler 2 to 4GHz 20dB Type N	75.00

Hewlett Packard

H487B	100 ohms Neg. Thermistor Mount (NEW)	150.00
H487B	100 ohms Neg. Thermistor Mount (USED)	100.00
477B	200 ohms Neg. Thermistor Mount (USED)	100.00
X487A	100 ohms Neg. Thermistor Mount (USED)	100.00
X487B	100 ohms Neg. Thermistor Mount (USED)	125.00
J468A	100 ohms Neg. Thermistor Mount (USED)	150.00
478A	200 ohms Neg. Thermistor Mount (USED)	150.00
8478A	200 ohms Balanced Neg. Thermistor Mount (USED)	175.00
J382	5.85 to 8.2 GHz Variable Attenuator 0 to 50dB	250.00
X382A	8.2 to 12.4 GHz Variable Attenuator 0 to 50dB	250.00
X885A	8.2 to 12.4 GHz Phase Shifter +/- 360°	250.00
394A	1 to 2 GHz Variable Attenuator 6 to 120dB	250.00
NK292A	Waveguide Adapter	65.00
K422A	18 to 26.5 GHz Crystal Detector	250.00
K375A	18 to 26.5 GHz Variable Attenuator	300.00
8436A	Bandpass Filter 8 to 12.4 GHz	75.00
8439A	2 GHz Notch Filter	75.00
8471A	RF Detector	50.00
342A	VHF Noise Source	100.00
X347A	8.2 to 12.4 GHz Noise Source	250.00
H532A	7.05 to 10 GHz Frequency Meter	300.00
G532A	3.95 to 5.85 GHz Frequency Meter	300.00
J532A	5.85 to 8.2 GHz Frequency Meter	300.00
809A	Carriage with a 444A Slotted Line Untuned Detector Probe and 809B Coaxial Slotted Section 2.6 to 18 GHz	175.00
809B	Carriage with a 442B Broadband Probe 2.6 to 12.4 GHz and a X810B Slotted Section	200.00
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Merrimac

AU-25A/	801115 Variable Attenuator	100.00
AU-26A/	801162 Variable Attenuator	100.00

Microlab/FXR

Y410A	Frequency Meter 12400 - 18000 MC	250.00
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Y610D	Coupler	75.00

Narda

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3042-20	Directional Coupler 950 to 2 GHz 20dB Type N	125.00
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22574	Directional Coupler 2 to 4 GHz 10dB Type N	125.00
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784/	22380 Variable Attenuator 1 to 90dB 2 to 2.5 GHz Type SMA	550.00
22377	Waveguide to Type N Adapter	35.00
720-6	Fixed Attenuator 8.2 to 14.4 GHz 6 dB	50.00
3503	Waveguide	25.00

PRD

U101	12.4 to 18 GHz Variable Attenuator 0 to 60dB	300.00
X101	8.2 to 12.4 GHz Variable Attenuator 0 to 60dB	200.00
C101	Variable Attenuator 0 to 60dB	200.00
205A/367	Slotted Line with Type N Adapter	100.00
195B	8.2 to 12.4 GHz Variable Attenuator 0 to 50dB	100.00
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196C	8.2 to 12.4 GHz Variable Attenuator 0 to 45dB	100.00
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109J,I	Fixed Attenuators	25.00
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COMPUTER I.C. SPECIALS

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2708	1K x 8 EPROM	$ 7.99
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2114/9114	1K x 4 Static RAM 450ns	6.99
2114L2	1K x 4 Static RAM 250ns	8.99
2114L3	1K x 4 Static RAM 350ns	7.99
4027	4K x 1 Dynamic RAM	3.99
4060/2107	4K x 1 Dynamic RAM	3.99
4050/9050	4K x 1 Dynamic RAM	3.99
2111A-2/8111	256 x 4 Static RAM	3.99
2112A-2	256 x 4 Static RAM	3.99
2115AL-2	1K x 1 Static RAM 55ns	4.99
6104-3/4104	4K x 1 Static RAM 320ns	14.99
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MC6821P	PIA	8.99
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MC6852L	SSDA	11.99
MC6854P	ADLC	22.00
MC6860CJCS	0-600 BPS Modem	29.00
MC6862L	2400 BPS Modem	14.99
MK3850N-3	F8 Microprocessor	9.99
MK3852P	F8 Memory Interface	16.99
MK3852N	F8 Memory Interface	9.99
MK3854N	F8 Direct Memory Access	9.99
8008-1	Microprocessor	4.99
8080A	Microprocessor	8.99
Z80CPU	Microprocessor	14.99
6520	PIA	7.99
6530	Support For 6500 series	15.99
2650	Microprocessor	10.99
TMS1000NL	Four Bit Microprocessor	9.99
TMS4024NC	9 x 64 Digital Storage Buffer (FIFO)	9.99
TMS6011NC	UART	9.99
MC14411	Bit Rate Generator	11.99
AY5-4007D	Four Digit Counter/Display Drivers	8.99
AY5-9200	Repertory Dialler	9.99
AY5-9100	Push Button Telephone Diallers	7.99
AY5-2376	Keyboard Encoder	19.99
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TR1402A	UART	9.99
PR1472B	UART	9.99
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8257	DMA Controller	9.99
8251	Communication Interface	9.99
8228	System Controller & Bus Driver	5.00
8212	8 Bit Input/Output Port	5.00
MC14410CP	2 of 8 Tone Encoder	9.99
MC14412	Low Speed Modem	14.99
MC14408	Binary to Phone Pulse Converter	12.99
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MC1489L	RS232 Receiver	1.00
MC1405L	A/D Converter Subsystem	9.00
MC1406L	6 Bit D/A Converter	7.50
MC1408/6/7/8	8 Bit D/A Converter	4.50
MC1330P	Low Level Video Detector	1.50
MC1349/50	Video IF Amplifier	1.17
MC1733L	LM733 OP Amplifier	2.40
LM565	Phase Lock Loop	2.50

MHz electronics

(602) 242-3037
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Phoenix, Arizona 85015

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WANTED: Millen ant. bridge model 90672, complete. For sale: Ten-Tec Omni D "digital" series "B," filters, N.B., 20 amp supply, manual & cartons. $1135.00. Jim Gysan, W1VYB (617) 922-3850 after 5:30 PM.

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XTAL FILTERS: C-F Networks 9 MHz SSB filter — 2.1 kHz BW, similar to KVG XF9-B, $25 postpaid. K8DRN, J. Wiggenhorn, 1678 NW 84th Dr., Coral Springs, FL 33065. 305-752-7444.

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SPECIAL SALE: Alliance HD-73 Heavy Duty Rotor $99.99 plus $3.00 shipping Continental USA. MC and Visa accepted. Scanner World, USA., 10-H New Scotland Ave., Albany, NY 12208. 518-436-9606.

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RECONDITIONED TEST EQUIPMENT for sale. Catalog $.50. Walter, 2697 Nickel, San Pablo, CA 94806.

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HP 416A Ratio Meter $10. HP 415A SWR Meter $22. HP 430A Power Meter, defective $15. Transformer 3 KV CT, 35 lbs. Kurt Bittmann, WB2YVY, 147 McGaw Ave., Centereach, NY 11720. (516) 585-9775.

WANTED: Microwave Dish 6 feet or larger. Willis, W9FGJ, 402 E. Cole St., DuQuoin, Il. 62832. (618) 542-2274.

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TRANSMITTER TECHNICIANS: Voice of America has opportunities for qualified technicians at VOA stations near Delano, California, and Greenville, North Carolina. Duties include operations/maintenance of high power shortwave transmitters and related facilities on shift basis. Minimum qualifications: 3-years chief broadcast engineer 5 to 50 KW, or 3-years supervisor of operations/maintenance high power military transmitting plant, or equivalent. U.S. citizenship required. Starting salary $18,760. Submit standard Federal application form, SF-171, to International Communication Agency, MGT/PDE (1-78) Washington, D.C. 20547. An Equal Opportunity Employer.

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Coming Events

NEW JERSEY: South Jersey Radio Association's 32nd annual Hamfest on September 7th at the Pennsauken Senior High School, Pennsauken, NJ. Flea market, prize drawings, contests, food and drink available. Indoor facilities if inclement weather. Admission: $3.00. Starts: 9:00 AM. Talk-in on 146.52 simplex, 146.22/.82 Repeater. More info: Edwin T. Kephart, 4309 Willis Ave., Pennsauken, NJ 08109.

NEW YORK: HAM-O-RAMA '80 on September 12th and 13th at the Erie County Fairgrounds in Hamburg. Exhibits, tech programs, prizes, flea market. Plenty of free parking, free RV hookups. Advance tickets $3 — contact Ron Brodowski, KC2P, 260 Hilltop Drive, Elma, NY 14059. (716) 652-6754.

NEW HAMPSHIRE: C.V.F.M. Hamfest on September 28 at the King Ridge Ski Area, Sutton, NH (follow signs off of I-93). Admission: $3.00. Giant flea market, florist exhibit for the gals, dealer's exhibits, food, overnight camping available, door prizes and much more. More info: C.A. Breuning, 54 Myrtle St., Newport, NH 03773.

More Details? CHECK—OFF Page 94

MICHIGAN: Five County Swap-N-Shop on September 21 at the Southwestern High School, 1420 W. 12th St., Flint, Michigan. Food, free parking, prizes and more. Tickets: $2.00 per person advance and $3.00 at door. More info: Bob Ross, P.O. Box 7671, Flint, Michigan 48507. (313) 239-0397.

VIRGINIA STATE ARRL CONVENTION: The Fifth Annual Tidewater Hamfest and ARRL Virginia State Convention will be in the great new Virginia Beach, Virginia Arts and Conference Center, October 4 and 5, 1980. ARRL, Traffic, DX Forums, XYL free bingo and lounge. Admission $3.50. Advance admission ticket drawing for Kenwood FM transceiver. Flea market spaces $3.00 day. Ticket and information - TRC, P.O. Box 7101, Portsmouth, Virginia 23707 SASE.

INDIANA: Porter County Amateur Radio Club's annual Hamfest at the Porter County Fairgrounds, Valparaiso, Indiana, on September 14. Flea market, door prizes, dealers, plus more. Talk-in on 147.96/.36 and 146.52 MHz. Advanced tickets: $1.50. At gate: $2.00. More info or tickets: Charles Baker, P.O. Box 251, Portage, Indiana 46368.

GEORGIA: Lanierland Amateur Radio Club's HAMNIC at Lake Lanier Islands on September 28th. Large covered pavilion and parking area for Swap Shop and exhibits. Food available. No entry free for HAMNIC. Lanier Islands charge $2.50 entry fee per car. Picnic, hiking and swimming for kids. Trailer hook-ups and camping available on site. Many prizes. Talk-in on .07/.67. More info: Fred Runkle, 25 Stonehedge Dr., Buford, GA 30518.

FLORIDA: 15th annual Hamfest of the Platinum Coast Amateur Radio Society on September 6 and 7 at the Melbourne Civic Center. Swap tables, meetings, forums, and more. For info, swap tables, or reservations: PCARS, P.O. Box 1004, Melbourne, FL 32901.

PENNSYLVANIA: Skyview Radio annual Swap and Shop on September 21, 12:00 to 4:00 PM at Sokol Camp, Lower Burrell, Pennsylvania. Registration: $1.00, XYL's, YL's and children free. More info: Jim Jackson, RD #1, Box 7A, Apollo, PA 15613.

PENNSYLVANIA QSO PARTY sponsored by NARC on September 13 and 14. One of the "Grandaddys" of the state contests. For more info: Walt Supina, 525 W. Ridge Ave., State College, Pennsylvania 16801.

PENNSYLVANIA: Fort Venango Mike and Key Club will sponsor an expedition to Venus, Pennsylvania, on September 6 and 7. Operations on Novice and General portions of the band. All modes, all bands including 2 meter FM on 147.12-147.72. Operation will be in Clarion county. Certificate available S.A.S.E. required. More info: Joseph Szabat, 228 Plummer St., Oil City, PA 16301.

NEW YORK: Elmira Amateur Radio Association's 5th annual International Hamfest at Chemung Co. Fairgrounds on September 27. Starts at 8:00 AM. FCC and ARRL forums, flea market, many door prizes and food available. Dealer displays also. Contact: John Breese, 340 West Ave., Horseheads, New York 14845.

NEW YORK: Tu-boro A.R.C.'s auction on September 18 at the Odd Fellows Hall, 149-14 14th Ave., Whitestone, NY. For info call Walt, WB2PFO at (212) 539-5732. Talk-in on 146.52.

YLRL's HOWDY DAYS on September 10 at 1800 UTC through September 12 at 1800 UTC. Start the fall season by chatting with old friends and finding new ones. Extend invitation to join YLRL. For more info: Ione O'Donnell, WA2DMK, YLRL Vice-President, Newcomb, NY 12852.

DELTA QSO PARTY sponsored by the Delta Division of the ARRL on Sept. 27 (from 1800Z) to Sept. 28 (2400Z). No time or power restrictions. Amateurs outside the Delta Division will try and contact as many amateurs inside of the Delta Division (Ark-La-Miss-Tenn) as possible. Delta amateurs try for as many inside and outside Delta Division. For more info: Malcolm P. Keown, W5XX, 213 Moonmist, Vicksburg, MS 39180.

EX-KZ5 REUNION for all former KZ5s for 48 hours beginning 0001Z, September 27. Listen within the lowest 25 kHz of the CW and phone segments of the U.S. General Class portion on each band. More info: John B. Barham, PSC Box 4481, APO Miami, FL 34001.

DXPO '80 on September 27 and 28 at the Ramada Inn, Tysons Corner, Virginia. Sponsored by the National Capital DX Association. DXPO program, Attitude Adjustment Party (with ALC), DXPO International Banquet and prizes. More info: Dick Vincent, Rte. 1 Box 230, Bryantown, MD 20617.

VIRGINIA: Lynchburg's Kaleidescope Festival from September 22-28. CW operation will be 50 kcs up from the bottom of each band and SSB 10 kcs inside the General portion of the phone band. Obtain certificate QSL by sending 28¢ postage or 2 IRC's to K4HEX Manager, 212 Sandown Circle, Lynchburg, VA 24503.

NEW YORK: Yonkers Amateur Radio Club's annual flea market and Hamfest at Redmond Field in Yonkers, New York, on October 5, rain date: October 12. 9:00 A.M. - 5:00 P.M. Door prizes every hour and a giant final door prize. Giant auction starting at 3:00 P.M. Free parking, refreshments, picnic tables and more. Bring the family. For all Hams, CB'ers, SWL'ers, Hi-Fi'ers and Comp-Buffs. Admission: $1.50 per person. Under 12 free. Sellers: $3.00 per parking space. Talk-in on 146.865/.52 simplex. CB'ers, Channel 4 at 8:00 A.M. Advanced registration and more info: (914) 969-1053 (after 3:00). Ask for Otto.

IOWA: Cedar Valley Amateur Radio Club's hamfest on October 5th at the Hawkeye Downs Exhibition Hall in Cedar Rapids. Technical talks, large flea market, manufacturers and dealers. Tickets: $2.00 advance, $3.00 at door. Write CVARC Hamfest, Box 994, Cedar Rapids, IA 52406.

ALABAMA: Calhoun County Amateur Radio Association's hamfest on September 27 and 28 at the Municipal Auditorium, 1128 Gurnee Ave., Anniston, Alabama. 9 AM - 5 PM on Saturday and 9 AM - 3 PM on Sunday. Free admission, parking, bingo, and overnight self-contained RV parking. Over 8,000 sq. ft. air-conditioned exhibit area. Hourly drawings and final drawing on Sunday for great grand prizes. Donations for door prizes: $1 or 6 for $5. Talk-in on 69/09. Tables: $3.00 for one day, $5.00 for both. Contact: CCARA, P.O. Box 1624, Anniston, AL 36202. (205) 820-3619.

TEXAS: Houston Con-Vention 80 on October 3-5 at the Marriott Brookhollow Hotel, Houston, Texas. Commercial exhibitors, technical sessions, DX and contest activities, covered flea market, banquet, transmitter hunt, and much more. Host: Texas DX Society. Banquet speaker: Roy Neal, K6NUE, science editor for NBC news. Special hotel rate of $30 per day per room (four people) for convention attendees. Plenty of parking. More info: HHC, P.O. Box 79252, Houston, TX 77024.

LOUISIANA: New Orleans Hamfest-Computerfest (AMACOM '80) on October 10-12 at the Airport Hilton Inn, Kenner, Louisiana, across the street from the New Orleans International Airport. One of the largest gatherings of electronics hobbyists in the Deep South.

ILLINOIS: Sangamon Valley Radio Club's hamfest on September 28th at the Sangamon County Fairgrounds, New Berlin, 12 miles west of Springfield on Rt. 36. Indoor display and covered pavilion. Randy Rowe, NØTG talks on the Navassa DXpedition! Exhibits, kids activities and food available. Overnight camping. First prize: Kenwood TR-2400 H/T. Tickets: $1.50 advance, $2.00 gate. Info: Joe Suarez, WB9RFC, S.V.R.C., 1025 S. 6th, Springfield, IL 62703.

SOUTH CAROLINA: York County Amateur Radio Society's hamfest on October 5, at Joslin Park in Rock Hill. B-B-Q dinner and snack bar in park. For registration and prize info: Y.C.A.R.S., P.O. Box 414CRS, Rock Hill, SC 29730.

COLORADO: Boulder Amateur Radio Club's BARCFEST '80 on September 28th beginning at 9:00 A.M. at the Boulder National Guard Armory on North Broadway at the city limits. Auction, snack bar. $2.00 admission per family includes door prizes and swap space. Talk-in on 146.10/70 and 52/52. More info: Mark Call, NØMC, 4297 Redwood Ct., Boulder, CO 80301. (303) 442-2616.

GREATER DELAWARE VALLEY — 80 HAMFEST will be held October 19, 1980, in Pennsauken, New Jersey, at the Nashville East Cotillion Ballroom on Rt. 73 from 8 AM to 5 PM. Over 19,000 square feet of exhibit space (no hallways). Seminars, YL/XYL activities, and films. Door prizes hourly until 3:30. Talk-in 146.22/82. Tailgating is $3.00/10' space, indoor tables are $5.00. Tickets are $2.50 at the gate and $2.00 in advance. For reservations, maps or tickets write GDV-80, 15 East Camden Avenue, Moorestown, New Jersey 08057 or call 609-234-3926.

RADIO EXPO "80" Lake County fair grounds, Rt. 45 & 120. Sept. 6 & 7 — advanced tickets $2.00, $3.00 at gate. Write: Radio Expo Tickets, P.O. Box 1532, Evanston, IL 60204. Exhibitor information call (312) BST-EXPO.

MICHIGAN: Adrian Amateur Radio Club's 8th annual Hamfest on September 28 at the Lenawee County Fairgrounds, Adrian, Michigan. Tickets, tables, info: Adrian Amateur Radio Club, Inc., P.O. Box 26, Adrian, Michigan 49221.

FINDLAY HAMFEST: The 38th Annual Findlay Hamfest greets you on Sept. 7th with a fine new indoor/outdoor location, The Hancock Recreational Center, just east of I-75 exit 161, on the north edge of Findlay, 40 miles south of Toledo. Main Prizes: a TS-120s W/supply, two TR-2400's, and an AT-120 matcher. Tickets $2.00 advance and $2.50 at the door. Reserve your tables early: $2.50 per ½. Open Saturday 17:00 till 22:00 for forums and set-up, Sunday at 05:00. Join the over 6000 people attending Findlay Hamfest this year and spend your bucks on the best! For tickets, info, and reservations send S.A.S.E. to P.O. Box 587, Findlay, Ohio 45840.

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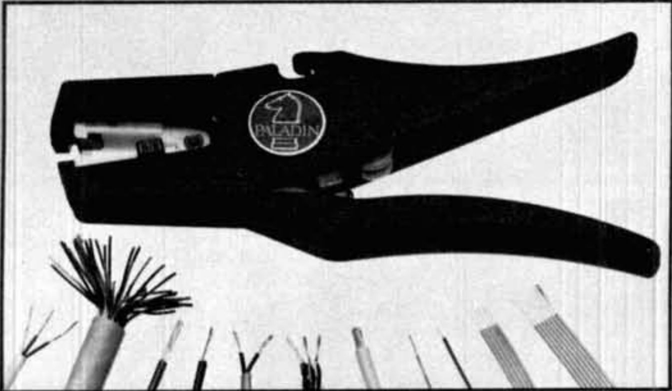

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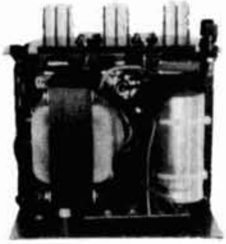
RS-4A	3	4	3¾ X 6½ X 9	5	$39.95



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