COMMUNICATIONS QUARTERLY

THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

Summer 1992

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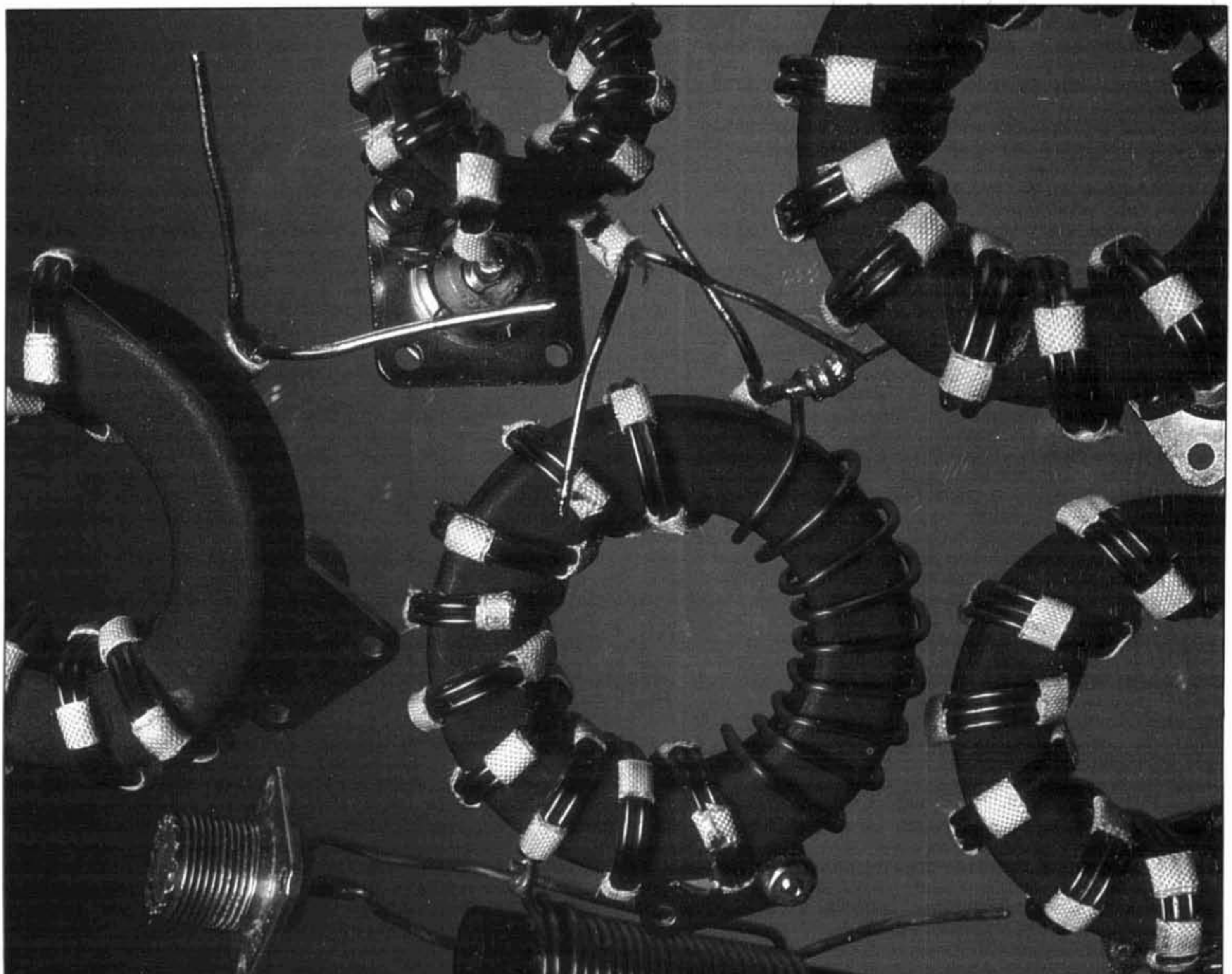

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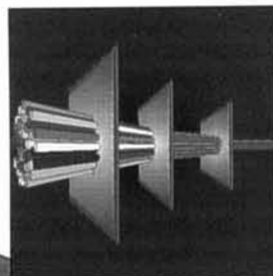

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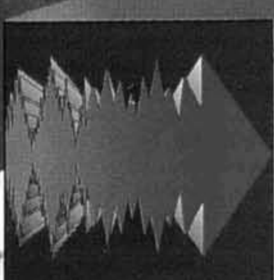

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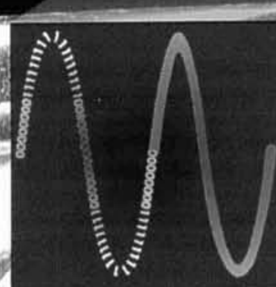

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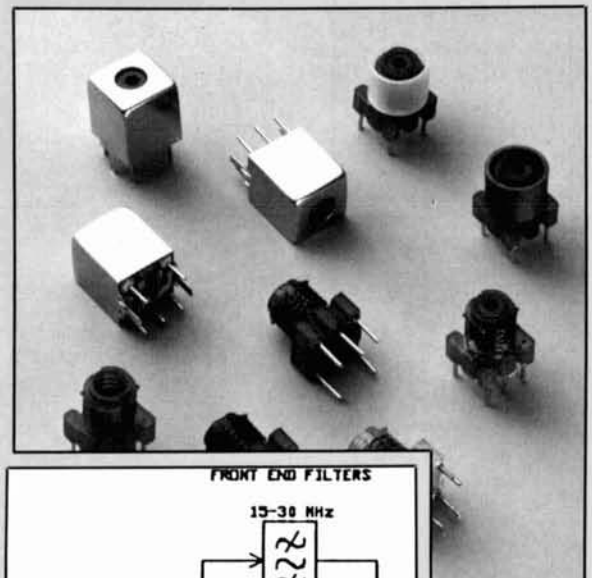
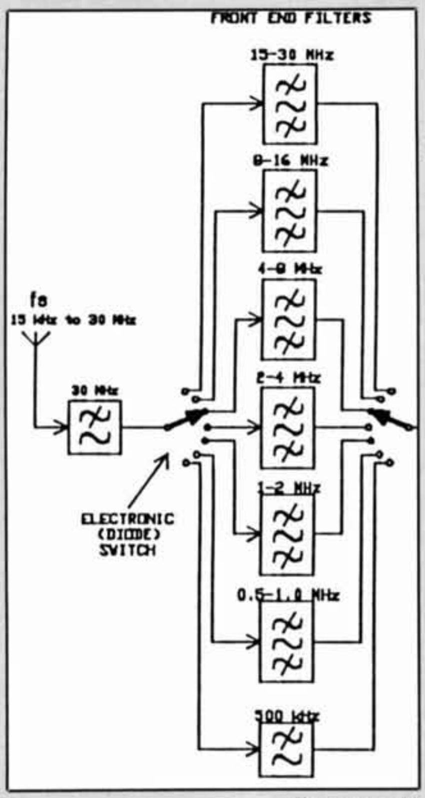


Littlefield, page 38

Cover photo: Baluns and more baluns. See Baluns Revisited, page 13. Photo by Robert S. Le Blanc.



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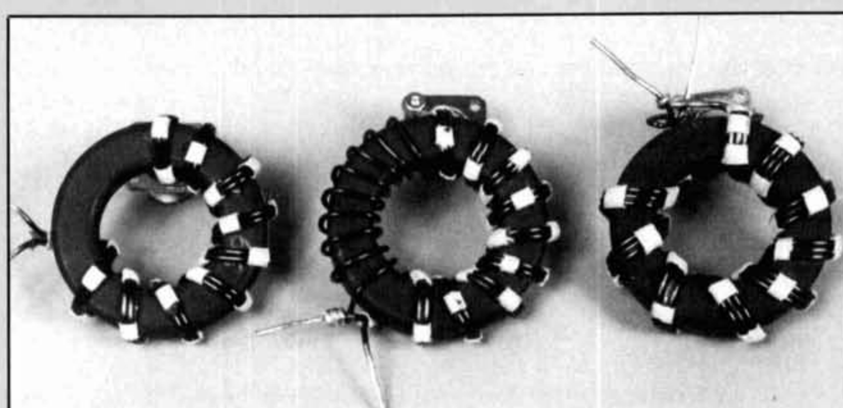

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EDITORIAL

The Code Doesn't Necessarily Make The Ham.

From a numbers standpoint, the Codeless Technician license is a rousing success. According to the April 15, 1992 issue of the *W5YI Report*, the code-free ticket is the entry level of choice for two thirds of all first-time licensees.

Although the code-free licensing option has provided a much-needed boost to our ranks, there are those among us who seem to resent these fledgling hams. Undoubtedly, this is a small segment of the amateur population. But unfortunately, even a small group of naysayers can give the rest of us a bad name.

While many no-code newcomers are warmly received by hams licensed under the old system, some face indifference and even blatant rudeness on the bands. Of course, this shouldn't come as a surprise. When talk resurfaced about a codeless license back in the late 1980s, a vocal contingent of amateurs railed against dropping the code entry requirement. Their rallying cry was: "I had to learn to code to become a ham, and so should everyone else." It's too bad these attitudes have created the need for articles like "Wally and Mike: The Podunk Repeater Club" in the July 1992 issue of *QST*—a piece that underscores the damage that such negativism can cause.

Of course, this snobbery isn't limited to the United States. No-code entrants to the amateur radio service in Canada and Great Britain also face resistance from some old-timers. Letters to the Editor in recent issues of *Practical Wireless* and *The Canadian Amateur Radio Magazine* indicate that attitudes similar to those faced by no-code hams in the U.S. are challenging new Canadian and British code-less amateurs, as well.

It even seems that these attitudes about the new licensing procedures have led to a "jump to conclusions" type of mentality on the bands. A friend of mine obtained his amateur radio license in the traditional way. He first became licensed as a novice, then took his technician exam shortly after, so he could move up through the ranks. He decided to trade his novice call, KA1YGS, in for the tech call he'd earned. A short time after he received his new call (N1KCZ) and went on the air with it, another ham responded in a rather derogatory fashion—asking him if he was "one of those new no-code technicians." Nick was appalled, not because he had been "accused" of being a no-code ham, but by the rudeness of his fellow amatuer. He commented to me at the time that it was a shame some people couldn't welcome all new amateurs equally, but had to play childish one-upmanship games instead.

Since that episode, Nick has upgraded to general, is on the board of directors of his local amateur radio club—and has kept his tech call. When we spoke recently, he noted that knowing the code doesn't make one a "good" amateur radio operator—one only has to listen to 75 meters some evening to get a real earful of racial and gender slurs, foul language, and the like. And let's be realistic, the manners of those competing for rare DX stations aren't much better.

Even though the code requirement has been eliminated as a first step toward getting on the air, it's important to note that the Codeless Technician license doesn't spell the end of code on the ham bands. In order to upgrade, international law requires that these new techs learn the code. No-code licensing simply gives many would-be hams the opportunity to "try things out" before they invest a lot of what has become, for most of us, increasingly precious leisure time in learning the code.

But lest you think all will choose the no-code path, take heart. My twelve-year-old son, Drew, who attended his first Dayton Hamvention this spring, feels that code is the "only way to go." For Drew, learning the code is an important part of the mystique of becoming an amateur radio operator. He plans to start as a Novice, then upgrade to Technician. I only hope that when he upgrades, he won't face the rude treatment Nick encountered.

So remember, the next generation of code-loving hams is waiting in the wings. Those who fear that code will go the way of the dinosaurs need to consider that many continue to flock to the CW bands to find relief from the cacophony present on our phone bands. In the meantime, let's support all who wish to join the amateur radio fraternity—whether they choose to start as Novices or Codeless Technicians. That code-free ham you hear on the radio could be a close friend or a member of your family.

Terry Littlefield, KA1STC
Editor

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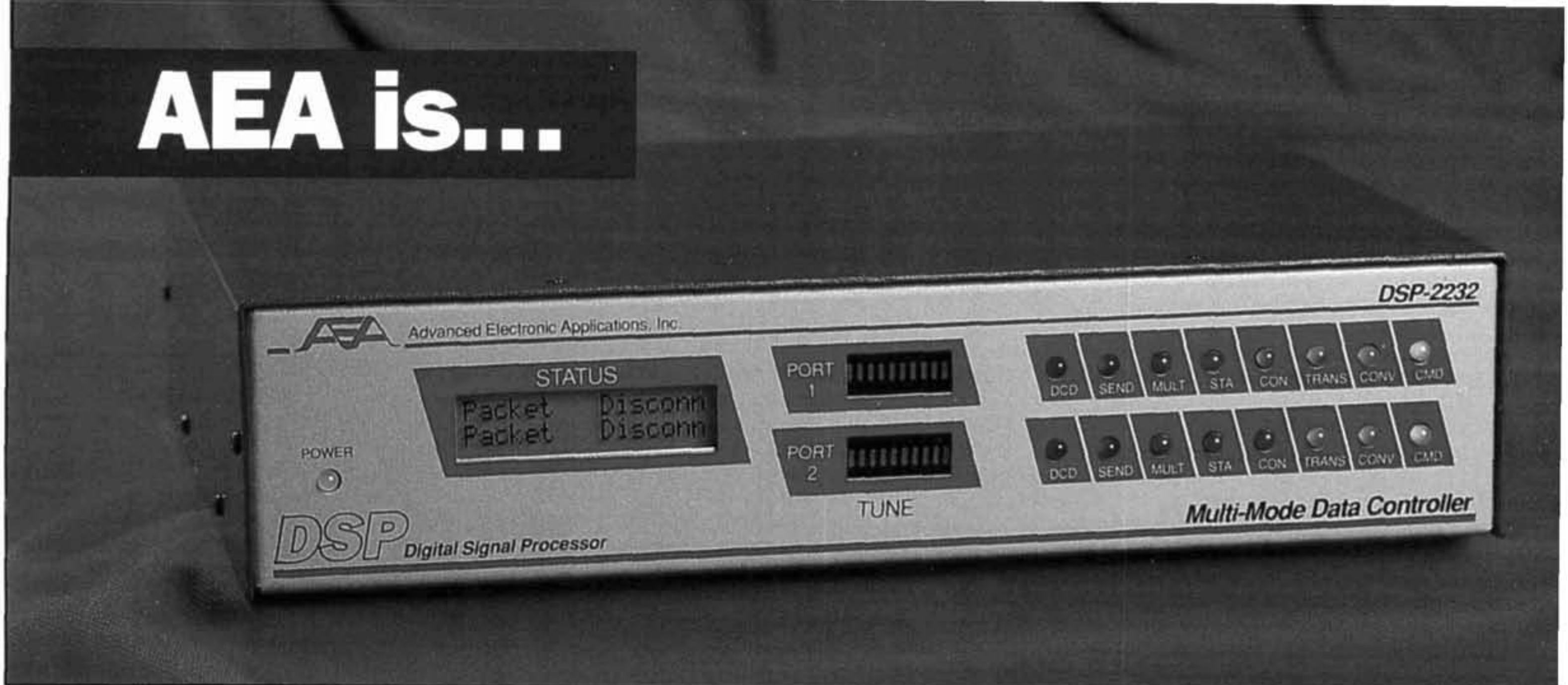

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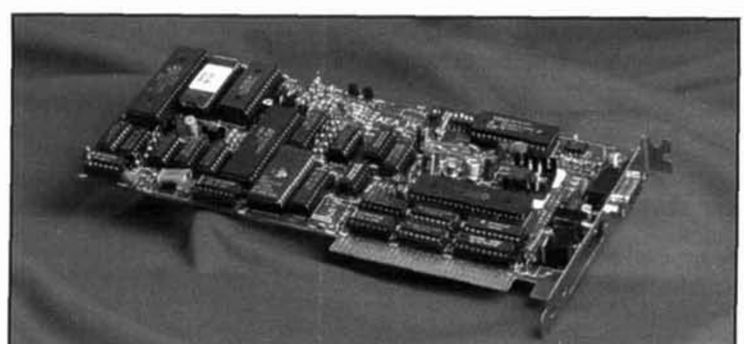

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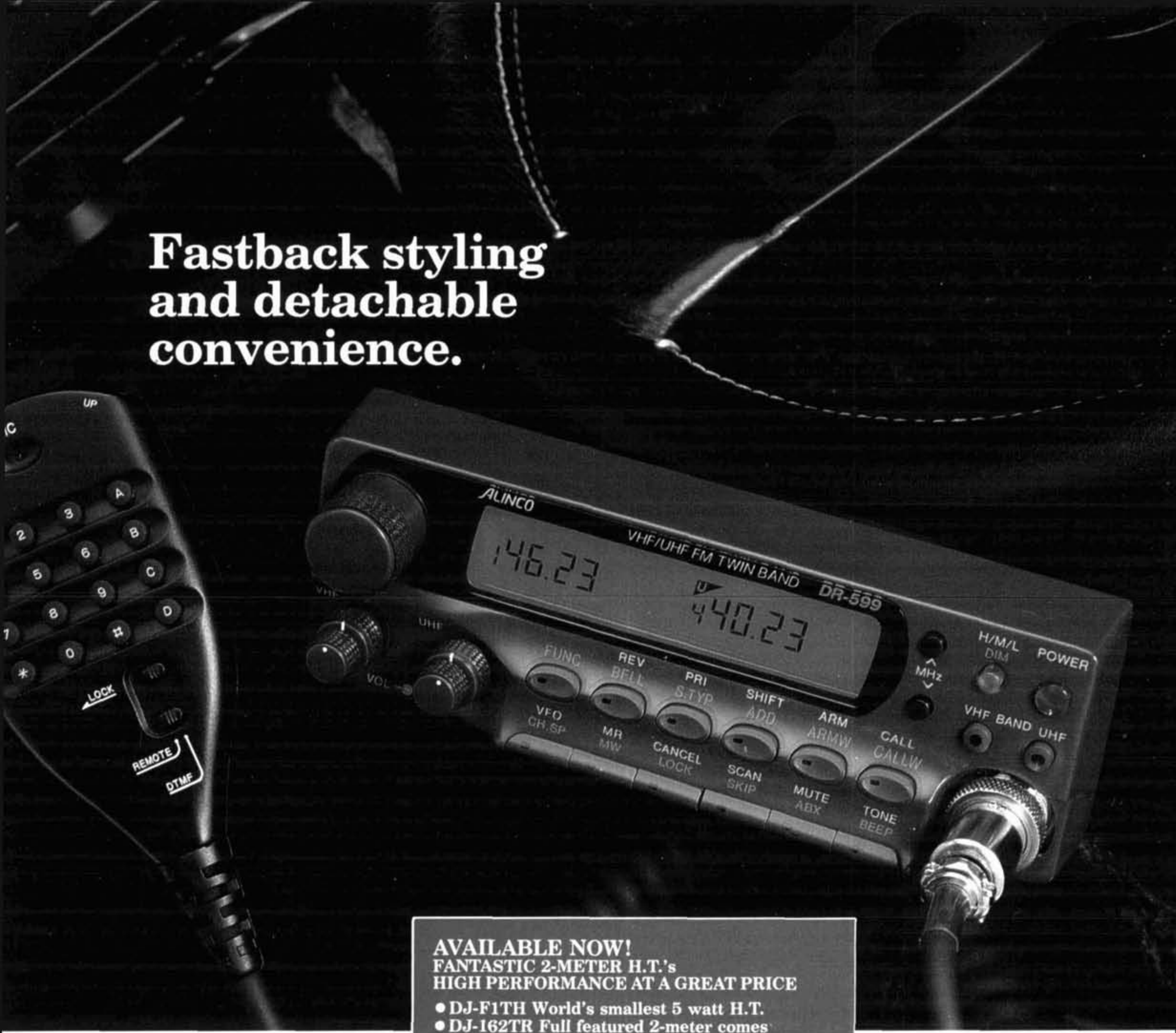

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Space limitations are not a problem for the DR-599T. The detachable control head allows the transceiver body to be hidden from view while the sleek control head can be neatly mounted on a dash or visor. (Optional separation kit required).

This DR-599T has full duplex cross band operation and cross band repeater functions with real-world power and excellent sensitivity. Airband receive is also possible with a simple modification.

The standard features include 3 power output selections, ARM, ABX, Bell, Reverse, Mute, and Auto Dialer functions. Also, with the optional EJ-8U DTMF Decoder board installed, the DR-599T can be remotely operated from another transceiver.

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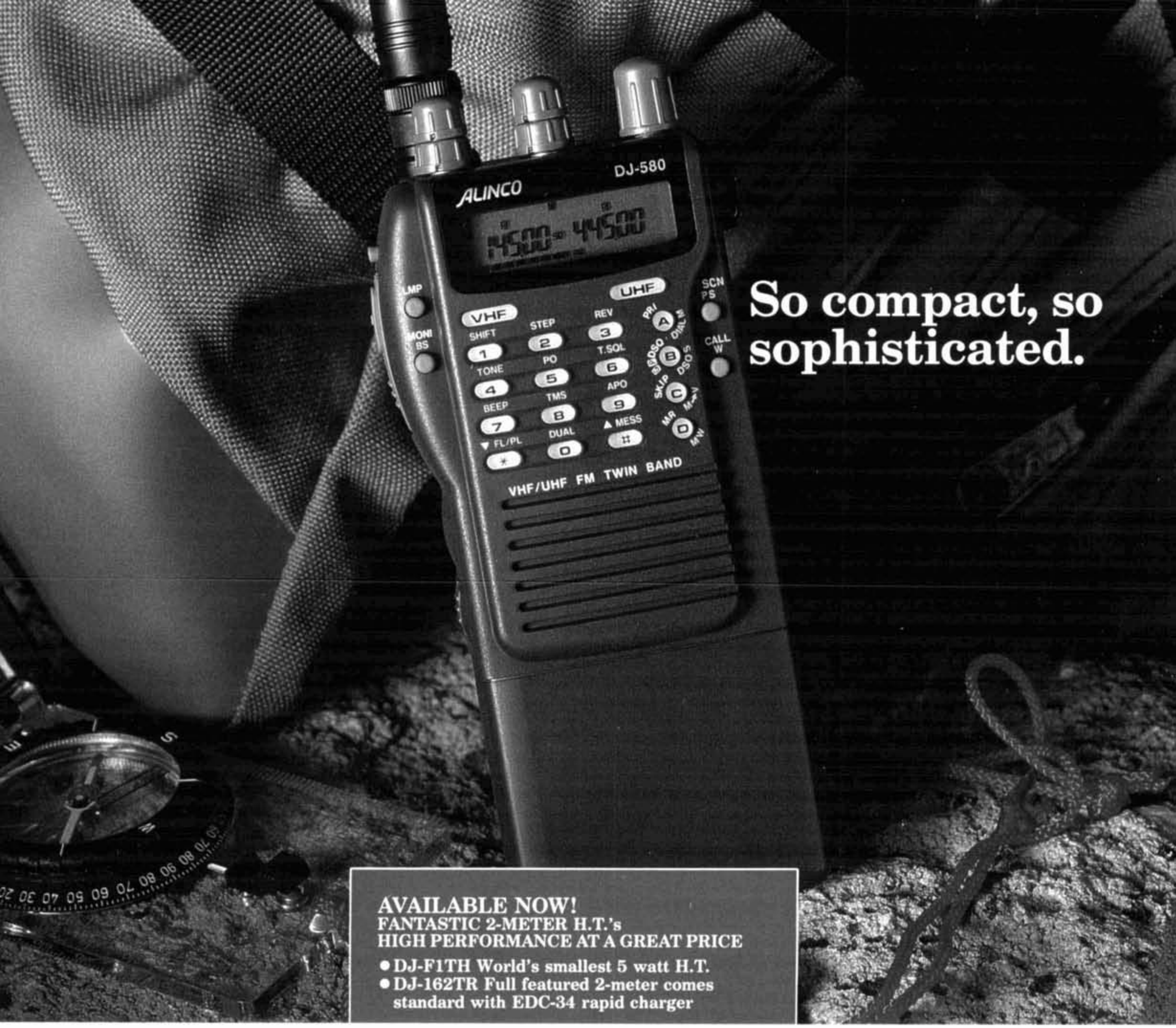

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This unit has built in DSQ for paging, CTCSS encode and decode standard, various scanning functions, 3 power level selections for each band, bell function, and an illuminated keypad.

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Comments from two old radio buffs

I have just received my Spring 1992 copy of the quarterly and, upon looking at the listing of contents, turned to the article on page 93. This article by Joseph Carr, K4IPV, describes long term developments in radio receivers.

I, too, am enamored by reflections on and the sight of old radio apparatus. Joe's article made me think of the good ole days when I, as a short-trousered, down-at-the-heel kid, looked for discarded radios for parts and for dry batteries which still retained some energy.

In Joe's writing on page 98 there appears a photo of a receiver described as a crystal set with audio amplification. I question the accuracy of this description. It appears to me that the receiver is a good old Radiola III, a real mainstay for RCA in the years 1923-24. This receiver did not incorporate a crystal in its circuit.

A minor matter perhaps, but I am sure Joe Carr will agree with me.

Two thoughts come to mind:

1. The receiver pictured on page 98 is not a crystal set.
2. People do read *Communications Quarterly* carefully.

Paul R. Flaugher, W8VEL
Cincinnati, Ohio

It was nice to see the "Radio Receivers of the Past" article in the Spring 1992 issue of *Communications Quarterly*. It should be mentioned that besides the early coherer detector and the galena with its adjustable phospher-bronze catwhisker; other types of detectors, such as the Carborundum, were also commonly used. Most early crystal sets used adjustable inductors to achieve matching and tuning; both sliding contacts along the coil body and tapped coil positions via a "tap switch" were employed.

The TRF receiver shown in **Photo G** was then and is still referred to as a "three-dialer" by radio collectors. Despite their inclusion in various schematics in this article, resistive volume controls were not used in these early sets. Volume reduction (if needed!) was accomplished by rheostats controlling the filament voltages of the RF, detector, and audio tubes—or by simply detuning, or by adjusting the antenna coupling.

One might surmise from the progression shown in the article that the early regenerative was doomed by the new TRF three-dialer sets. However, the Crosley company, started by Colton Crosley in 1921, was still producing regenerative radio models in large numbers well into the mid-20s that were both cost and performance comparable to the TRF designs. The receiver shown in **Photo B** credited with being an early amplified crystal set is an RCA Radiola III and is actually a regenerative design.

Professor L.A. Hazeltine's neutrodyne design was based on his work developing the SE1420 Navy receiver. Hazeltine practiced his engineering first with pencil and paper—unlike the trial and error approach followed by many of his peers—and, as a result, usually produced a working model that worked right the first time (M.A. Molner, "The birthplace of the Neutrodyne, *ARC* March 1992).

(Continued on page 106)

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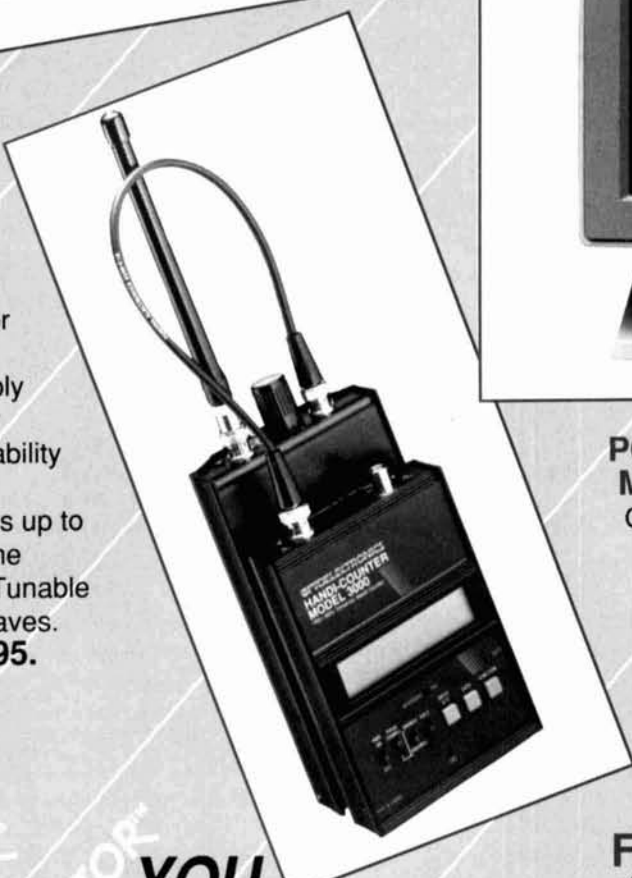

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MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
SL-11A	•	•	7	11	2¾ x 7⅝ x 9¾	11

RS-L SERIES

• POWER SUPPLIES WITH BUILT IN CIGARETTE LIGHTER RECEPTACLE

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-4L	3	4	3½ x 6⅛ x 7¼	6
RS-5L	4	5	3½ x 6⅛ x 7¼	7

RM SERIES

MODEL RM-35M

• 19" RACK MOUNT POWER SUPPLIES

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RM-12A	9	12	5¼ × 19 × 8¼	16
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50
RM-60A	50	55	7 × 19 × 12½	60
• Separate Volt and Amp Meters				
RM-12M	9	12	5¼ × 19 × 8¼	16
RM-35M	25	35	5¼ × 19 × 12½	38
RM-50M	37	50	5¼ × 19 × 12½	50
RM-60M	50	55	7 × 19 × 12½	60

RS-A SERIES

MODEL RS-7A

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-3A		•	2.5	3	3 × 4¾ × 5¾	4
RS-4A	•	•	3	4	3¾ × 6½ × 9	5
RS-5A		•	4	5	3½ × 6⅛ × 7¼	7
RS-7A	•	•	5	7	3¾ × 6½ × 9	9
RS-7B	•	•	5	7	4 × 7½ × 10¾	10
RS-10A	•	•	7.5	10	4 × 7½ × 10¾	11
RS-12A	•	•	9	12	4½ × 8 × 9	13
RS-12B		•	9	12	4 × 7½ × 10¾	13
RS-20A	•	•	16	20	5 × 9 × 10½	18
RS-35A	•	•	25	35	5 × 11 × 11	27
RS-50A	•		37	50	6 × 13¾ × 11	46

RS-M SERIES

MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• Switchable volt and Amp meter				
RS-12M	9	12	4½ × 8 × 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 × 9 × 10½	18
RS-35M	25	35	5 × 11 × 11	27
RS-50M	37	50	6 × 13¾ × 11	46

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• Separate Volt and Amp Meters • Output Voltage adjustable from 2-15 volts • Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps) @13.8VDC	@10VDC	@5VDC	ICS* (Amps) @13.8V	Size (IN) H × W × D	Shipping Wt. (lbs.)
VS-12M	9	5	2	12	4½ × 8 × 9	13
VS-20M	16	9	4	20	5 × 9 × 10½	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13¾ × 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5¼ × 19 × 12½	38
VRM-50M	37	22	10	50	5¼ × 19 × 12½	50

RS-S SERIES

MODEL RS-12S

• Built in speaker

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* Amps	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-7S	•	•	5	7	4 × 7½ × 10¾	10
RS-10S	•	•	7.5	10	4 × 7½ × 10¾	12
RS-12S	•	•	9	12	4½ × 8 × 9	13
RS-20S	•	•	16	20	5 × 9 × 10½	18

Jerry Sevick, *W2FMI*
32 Granville Way
Basking Ridge, New Jersey 07920

BALUNS REVISITED

The first in a series of articles on balun history, theory, application, and design

Recent articles in the amateur radio journals[1-4] have coined new terms for the 1:1 and 4:1 baluns. They are: *current*, *choke*, and *voltage* baluns. New balun designs using coaxial cables either wound around a ferrite core or threaded through ferrite beads have also appeared in these articles. In comparing these "new" baluns with the so-called transformer-type baluns (wire transmission lines wound around a core) or voltage-type baluns (described later), the authors advanced various claims. They include: a) had less loss, b) were not susceptible to unbalanced or mismatched loads, c) were not prone to core saturation, d) exhibited true 1:1 impedance transformations, e) did not exhibit leakage inductance, and f) did not add a reactive component to the input impedance of the antenna.

As I looked further into the amateur literature, I found other claims (which I also question) and information that was overlooked by the authors of the four articles cited above. Therefore, I thought a series of articles giving the history, theory, applications and designs of baluns was needed at this time. My goal is to give readers a better understanding of the claims put forth for these baluns, and the uses and limitations of these highly efficient and broadband transmission line transformers (the 1:1 and 4:1 baluns only being a subset thereof). Many of the views expressed in these articles will be in contradiction to those that have been published before.

This first article presents: a) an explanation of the transmission line transformer, b) the basic building block which is the model for developing the theory for these devices, c) the three basic forms of the 1:1 balun, d) high and low-power versions of my so-called transformer-type balun and finally, e) when to use a 1:1 balun.

The transmission line transformer

The broadband balun belongs to a class of transformers known as *transmission line transformers*.[5] Other popular transformers in this class are "ununs" (unbalanced-to-unbalanced transformers),* hybrids, and combiners. This class of transformers differs completely from the conventional transformer that transmits the energy to the output circuit by flux linkages. The transmission line transformer transmits the energy by an efficient transmission line mode.

Conventional transformers have been constructed to perform over wide bandwidths. Losses on the order of one decibel can exist over a range from a few kilohertz to over 200 MHz. Losses can be as low as 0.2 dB over a considerable portion of this band.

However, transmission line transformers are capable of far wider bandwidths and much greater efficiencies. The stray inductances and interwinding capacitances are generally absorbed into the characteristic impedance of the transmission line. Therefore, they form no resonances that could seriously limit the high-frequency response. Here the response is limited by the deviation of the characteristic impedance from the optimum value; the parasitics not absorbed into the characteristic impedance of the transmission line; and in some designs, the length of the transmission line.

With transmission lines, the flux is effectively canceled out in the core (or beads) and extremely high efficiencies are possible over large portions of the passband. There are losses of only 0.02 to 0.04 dB with certain ferrite materials.

Another major difference between the

*For more information on ununs, read "How to Build a Multi-Tap Unun," by Richard A. Genaille, W4UW, in *CQ*, May 1992, pages 28-32. Ed.

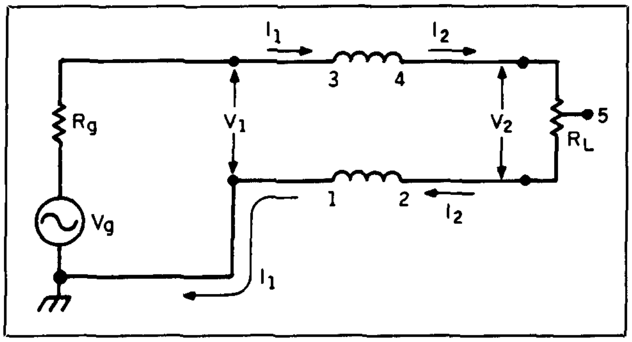

Figure 1. The Guanella 1:1 balun: the basic building block.

two classes of transformers is the loss mechanism. The losses with conventional transformers (eddy-current, wire, and hysteresis losses) are current-dependent. The losses with transmission line transformers are voltage-dependent (a dielectric-type loss). Therefore, higher-impedance transformers or transformers subjected to high VSWRs have greater losses because of the increased voltage drops along the lengths of the transmission lines.

Therefore, the transmission line transformer is in reality a choke (which eliminates the undesirable currents) and a configuration of transmission lines. As a result, true comparisons between baluns can only be made when they are perceived as transmission line transformers and not conventional transformers.

The Guanella Balun: the basic building block

Guanella made the first presentation on broadband baluns in 1944.[6] He proposed coiling a transmission line to form a choke where the input was isolated from the output (an isolation transformer) resulting in the rejection of the undesirable currents (common-mode, conventional transformer, and antenna currents). In other words, only transmission line currents (which are flux canceling in the core) were allowed to flow. In essence, Guanella's classic paper proposed what is known today as the *current* or *choke* balun.

Figure 1 shows the schematic of his 1:1 design. If the reactance of the coiled winding is much greater than the load, R_L, then only transmission line currents are allowed to flow, no matter where a ground is connected to the load. With a ground at terminal 5, the center of the load, and the characteristic impedance of the transmission line being equal to the load (a "flat" line), the voltage at terminal 4 is $V_1/2$ and at terminal 2, it is $-V_1/2$. Thus, a balanced output is obtained. If R_L represents the impedance of a dipole, the voltages are still the same because of the virtual ground plane that bisects the center of the antenna.

It's important to note that a negative potential gradient of $-V_1/2$ exists along the length of the transmission line. With sufficient choking so only transmission line currents are allowed to flow, this voltage drop accounts for the loss (a dielectric-type) in all transmission line transformers! With loads greater than the characteristic impedance of the transmission lines, the voltage gradients become larger resulting in higher losses. In other words, with high VSWRs due to larger loads, the losses can become excessive. Additionally, because higher impedance baluns, like the 4:1 (50:200 ohms), have higher voltage gradients, their losses are also greater than those of the 1:1 (50:50 ohm) baluns.

It should also be pointed out that the preceding analysis does not only apply to baluns using coiled-wire transmission lines. It applies equally well to baluns using coaxial cable coiled about a ferrite core or threaded through ferrite beads. They, too, have potential gradients along both the inner conductor and the outside of the outer braid. The one significant difference is that the *threaded-bead coax balun* (like the W2DU balun) requires high-permeability (2500) ferrite beads in order to obtain the necessary choking reactance in the HF band. Accurate measurements have shown[5] that low-permeability ferrites (less than 300) are necessary to achieve the extremely high efficiencies of which these transformers are capable. However, the coiled-type balun can use the lower permeability ferrites (and have higher efficiencies) because of the mutual magnetic coupling between the turns.

Ruthroff, who published the only other classic paper on baluns in 1959,[7] recognized another function that Guanella's circuit provided. By grounding terminal 4, he found that the schematic became a *phase-inverter*. Experiments have shown that the highest choking reactance is now required to prevent a shunting current to ground and hence unequal currents in the windings. Furthermore, with a matched load, the voltage drop along the lengths of the transmission lines now becomes $-V_1$. Thus, the loss in the phase-inverter connection is greater than that of the balun.

With information from the patterns shown by the various forms of baluns and ununs, I was able to add (and complete) the other two functions possible with the basic building block. They are: a) the *delay line* when terminal 2 is grounded and b) the *bootstrap* when terminal 2 is connected to

Photo A. The three basic forms of the 1:1 balun: left, the Guanella (current) balun; center, the Ruthroff balun; and right, the trifilar (voltage) balun.

terminal 3. It can be shown, in the case of the delay line function, that no potential drop exists along the transmission lines—hence ferrite cores or beads are not required. Therefore, the only loss with the delay line function is in the transmission line itself, and this is negligible.

Finally, it can be shown that the bootstrap function plays a very significant role with ununs. By connecting terminal 2 to terminal 3, a positive potential gradient of $+V_1$ appears across the lengths of the windings. Thus terminal 4, with a matched load, has a potential of $2V_1$ to ground. If the bottom of R_L is returned to ground (instead of to terminal 2), the potential across it is also $2V_1$, resulting in Ruthroff's 4:1 unun. By using higher order windings (trifilar, quadrifilar, etc.), I was able to demonstrate ununs with very wideband ratios of less than 4:1.[5]

The three basic forms of the 1:1 balun

The first (and latest) form of the 1:1 balun is that of the two-conductor (coaxial cable or twin-lead) Guanella balun described before. An example is shown on the left in **Photo A**. As was suggested, with sufficient choking, the analysis of this form of balun rests solely upon transmission line theory. But two other basic forms also appeared upon the scene. They are the three-conductor balun by Ruthroff and the three-conductor voltage balun (originator unknown to the author).

A version of the Ruthroff balun is shown in the center of **Photo A**. As you can see, the third winding is on a separate part of the toroid. Unfortunately, the schematic (originally drawn) shown in **Figure 2A** appears to be trifilar in nature. This is far from the truth. A trifilar connection would have a shorted transmission line with windings 3-4 and 5-6. The third winding, as Ruthroff stated, completes the path for the magnetizing current. Thus, the third winding gives his balun a better low-frequency response for the same number of turns, as with the Guanella balun. This is because, at the low-frequency end, the energy is transmitted to the load by both a transmission line mode and an autotransformer mode. In this area, his balun is susceptible to an unbalanced load due to the voltage divider action of the third wire. As the frequency is increased, the Ruthroff balun then takes on

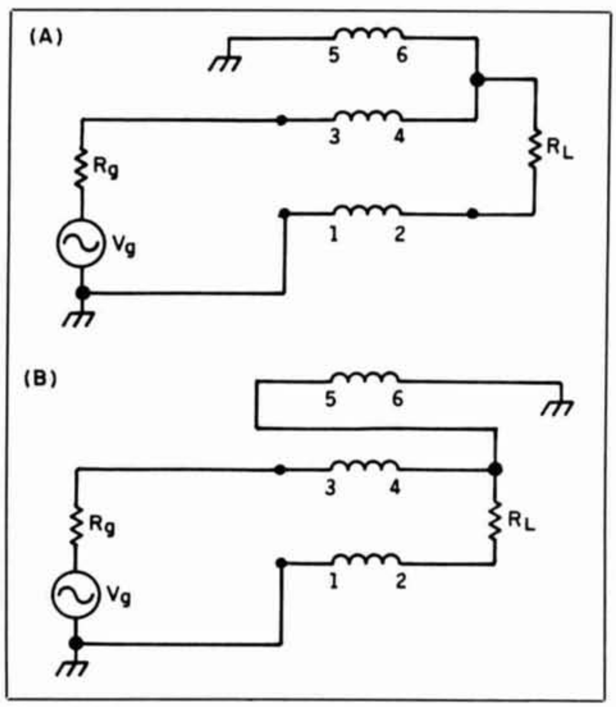

Figure 2. The schematic diagrams of the two basic forms of the three-conductor 1:1 balun. (A) The Ruthroff balun. (B) The voltage balun (originator unknown).

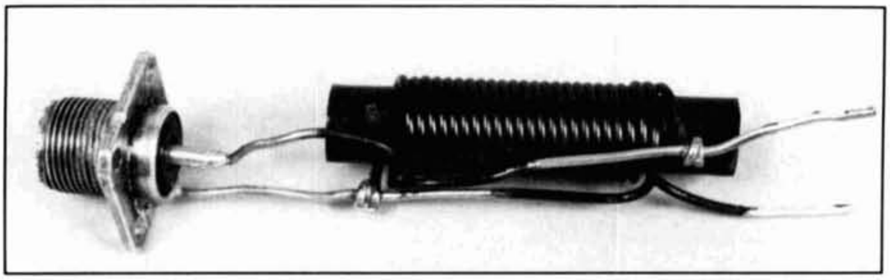

Photo B. A typical 1:1 rod-type voltage balun (W2AU).

the character of a Guanella balun which is insensitive to unbalanced and mismatched loads. However, there is a disadvantage with the third winding. If the choking action is inadequate, it could approach a shorted turn at the low-frequency end with possible damage to the core.

The trifilar voltage balun is shown on the right in **Photo A** and in **Figure 2B**. Unfortunately (again), this form of the 1:1 balun has been the main one used for comparisons with the "new" coaxial cable baluns. Because this configuration results in two transmission lines that are tightly coupled (the center conductor being common), this balun is highly susceptible to unbalanced and mismatched loads throughout its entire frequency range. As can be seen in **Figure 2B**, terminal 4 is at the midpoint of two transmission lines in series. Therefore, if the load is grounded off-center or is not equal to the characteristic impedance of windings 1-2 and 3-4, a reactive component will be seen at the input terminals. Again (as with Ruthroff's balun), if insufficient choking exists, winding 5-6 can approach a short to ground at the low-frequency end. This distinction between the three forms of the 1:1 balun was clearly stated by Turrin in 1969.[8] Apparently, succeeding investigators failed to take advantage of his important paper.

A 1:1 balun design which has been very popular (and available) over the past years is shown in **Photo B**. This is the voltage balun with a low-permeability (125) ferrite rod for a core. The W2AU balun shown in **Photo B** is a typical example. It has several disadvantages. They are: 1) it uses the trifilar design, 2) it uses a rod core with only 8 trifilar turns, which gives insufficient choking reactance at the lower frequencies (80 and 160 meters), and 3) it is susceptible to voltage breakdown. There are much better baluns available today.

High-power and low-power 1:1 baluns

As was shown regarding the three different forms of the 1:1 balun, the voltage balun is *not* the recommended design. The Ruthroff balun, with its third winding on a separate part of the toroid, has a lower-frequency response advantage (for the same number of turns) over the Guanella balun. But this is at the expense of a susceptibility to unbalanced loads at the low-frequency end. Therefore, with sufficient choking reactance, the Guanella design is clearly the 1:1 balun of choice.

The next decision to make is whether to use a beaded-coax balun (like the W2DU balun) or a balun using a transmission line (coaxial cable or twin-lead) wound around a core. The beaded coax-balun is no doubt the easiest to construct. It's simply a matter of threading about 10 inches of coaxial cable through ferrite beads. But this balun has more loss than the coiled-type because high-permeability ferrite beads are required. The balun using coaxial cable wound around a core can result in an efficient design, but at the expense of voltage-breakdown. High voltage-breakdown cable like RG-8 is impossible to use in this structure. Therefore, a coiled-wire design using heavy-coated solid wire with extra layers of Scotch no. 92 polyimide tape to give the desired spacing for a characteristic impedance close to 50 ohms, appears to be the best choice. Its voltage-breakdown capability compares favorably with that of RG-8 cable.

Since 50-ohm twin-lead (of the kind used in these baluns) is not presently available, you must expend some effort to make it yourself. The more difficult part is putting a layer (or two) of Scotch no. 92 tape on the wire. But with a U-frame[5], or by simply using large nails at the proper spacing on a board, the tape can be applied (edgewise—like a window shade) without difficulty. When using no. 12 wire, the 1/2-inch wide tape just adds two needed layers.

Photo C shows high and low-power versions of the coiled-wire (transformer-type) Guanella 1:1 balun.* For operation from 10

*Kits and finished units available from Amidon Associates, Inc., 2216 East Gladwick Street, Dominguez Hills, California 90220.

to 160 meters, the high-power version on the left in **Photo C** has 10 bifilar turns of no. 12 H Thermaleze wire on a 2.4-inch OD ferrite toroid with a permeability of 250. One wire is covered with two layers of Scotch no. 92 polyimide tape, giving the winding a characteristic impedance close to 50 ohms. The wires, in turn, are clamped together about every 1/2-inch with strips of Scotch no. 27 glass tape. If operation is restricted from 10 meters to 80 meters, a ferrite toroid with a permeability of 125 is recommended. This ferrite would yield a slightly higher efficiency.

These two baluns should easily handle 5 kW of continuous power and 10 kW of peak power with good efficiency (98 to 99 percent under matched conditions). They also have sufficient safety margins at their low-frequency ends to handle most mismatched conditions.

The balun shown on the right in **Photo C** is an excellent choice for low-power applications. This balun has 10 bifilar turns of no. 16 H Thermaleze wire on a 1.25-inch OD ferrite toroid with a permeability of 250. One wire is covered with one layer of Scotch no. 92 tape. It is conservatively rated at 150 watts of continuous power and 300 watts of peak power from 10 to 160 meters.

When to use a balun

Baluns have taken on a more significant role in the past few decades with the advent of solid-state transceivers and Class B linear amplifiers which have unbalanced outputs. That is, the voltage on the center conductor of their output chassis connectors varies (plus and minus) with respect to ground. In many cases, coaxial cables are used as the transmission lines from these unbalanced outputs to antennas like dipoles, inverted Vs, and Yagi beams which favor a balanced feed. In essence, they prefer a source of power whose terminals are balanced (voltages being equal and opposite) with respect to actual ground or to the virtual ground plane which bisects the center of the antenna. The question asked frequently is whether a 1:1 balun is really needed.

To illustrate the problem involved and give a basis for my suggestions, I have included **Figure 3**. Here we have, at the feed-point of the dipole, two equal and opposite transmission line currents which have two components each—I_1 and I_2. Also shown is the spacing, *s*, between the center conductor and the outside braid. Theoretically, a balanced antenna with a balanced feed would have a ground (zero potential) plane bisecting this spacing. But since a coax-feed is unbalanced and the outer braid is also connected to ground at some point, an imbalance exists at the feed-point giving rise to two antenna modes. One is with I_1 giving a dipole mode and the other is with I_2 giving an inverted L mode.

If the spacing, *s*, is increased, the imbalance at the feed point becomes greater giving rise to more current on the outer braid and a larger unbalance of currents on the antenna's arms. Several steps can be taken to eliminate or minimize the undesirable inverted L mode (eliminate or reduce I_2). The obvious one is to use a balun which not only provides a balanced feed but also minimizes (by its choking reactance) I_2, if the coaxial cable does not lie in the ground plane which bisects the center of the dipole. The other step is to ground the coaxial cable at a quarter-wave (or odd-multiple thereof)

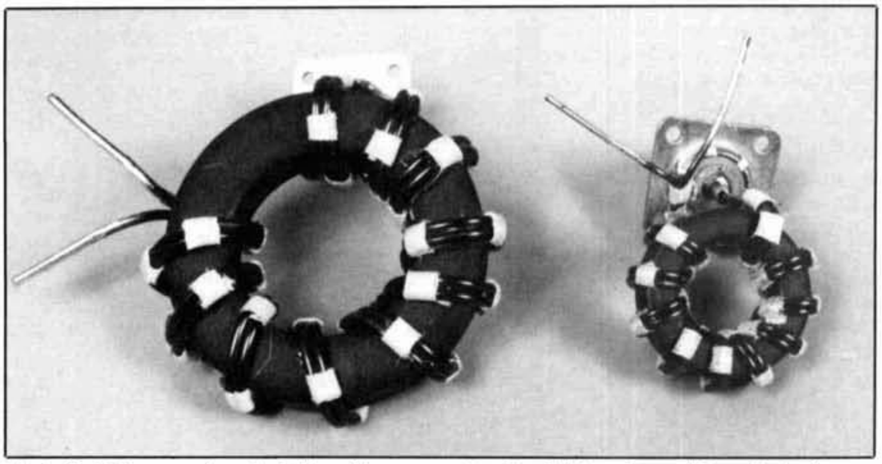

Photo C. High-power (on the left) and low-power (on the right) versions of the coiled-wire (transformer-type) Guanella 1:1 balun. These baluns are also known as current or choke baluns.

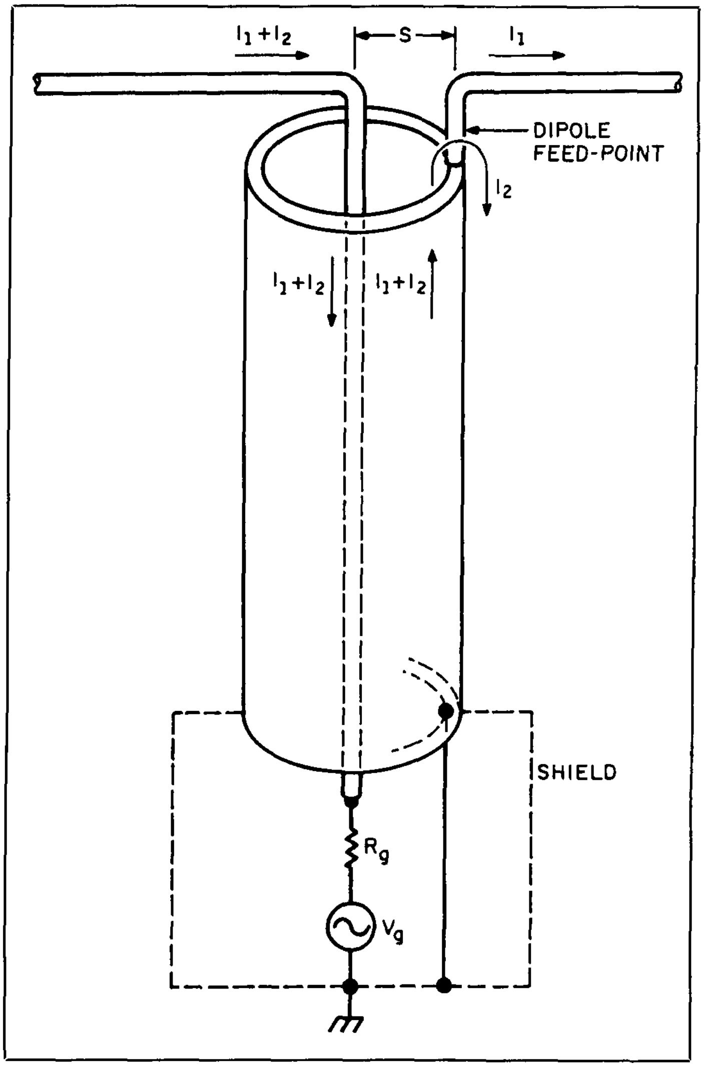

Figure 3. An illustration of the various currents at the feed-point of a dipole. I_1 is the dipole current and I_2, the inverted L current.

from the feed point. This discourages the inverted L mode only when the coaxial cable lies in the ground plane of the antenna.

Experiments with baluns were conducted on a 20-meter dipole at a height of 0.17 wavelengths which gave a resonant impedance of 50 ohms. VSWR curves were compared under various conditions. When the coaxial cable was in the ground plane of the antenna (that is, perpendicular to the axis of the antenna), the VSWR curves were identical with or without a balun no matter where the coaxial cable was grounded. Only when the coaxial cable was out of the ground plane, was a significant difference noted. When the coaxial cable dropped down at a 45 degree angle under the dipole, a large change in the VSWR curve took place. This meant that the inverted L mode was appreciable.

Another interesting experiment involved comparing the VSWR curves with Guanella (current) baluns with bifilar windings having characteristic impedances between 45 and 60 ohms. I did this by using no. 14 wire with various amounts of extra insulation (Scotch no. 92 tape). The result was that all VSWR curves were identical. This was to be expected because the lengths of the transmission lines in the baluns (about 24 inches) were much less than the wavelength on 20 meters.

Feeding a Yagi beam antenna without a balun is a different matter. Since most Yagi designs use shunt-feeding (usually by hairpin matching networks) in order to raise the input impedance close to 50 ohms, the effective spacing, *s*, is greatly increased. Furthermore, the center of the driven element is actually grounded. Thus, connecting the outer braid (which is grounded at some point) to one of the input terminals, creates a large imbalance and hence a real need for a balun. An interesting solution, which would also eliminate the matching network, is to use a step-down balun designed to match 50-ohm cable directly to the lower impedance of the driven element.[5]

In summary, if one considers the theoretical model of **Figure 3**, the experiments performed on 20 meters, and the reports from radio amateurs using dipoles without baluns, it appears that baluns are really needed for: a) Yagi beam antennas and b) dipoles and inverted Vs that have the coaxial cable feed line out of the ground plane of the antennas. That is, if the coaxial cable does not come away at right angles to the axis of a dipole or the plane of an inverted V, a balun is usually required. In general, the need for a balun is not so critical with dipoles and inverted Vs because the diameter of the coaxial cable is much smaller than the wavelengths in the HF band. It should also be pointed out that a well-designed balun is virtually lossless and only adds a foot or two to the coaxial cable feed line. And finally, *never* use a 1:1 balun to match into a *full-wave* dipole or inverted V. That's probably the way most baluns are destroyed. ■

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IMPROVING THE DRAKE TR-7

Reversible modifications enhance performance

Very few amateurs will disagree that today's HF ham gear has come a long way from what it was 10 or 20 years ago. Today's microprocessor-controlled radios offer an incredible number of features in small packages. All this technology has its price, however, as radios have become progressively more expensive, difficult to repair, and contain an increasing amount of surface-mount and proprietary components. At the same time we are offered this "high-tech" equipment, there is a growing interest in the simpler, less "option-intense" HF radios of the past.

The R.L. Drake Company, which held a commanding position in the amateur market in the 60s and 70s, was a trendsetter for the industry. The American-made TR-7, unveiled to the amateur community in 1978, is an excellent example of technological innovation. The TR-7 was one of the first amateur transceivers to offer continuous receiver coverage from 10 kHz to 30 MHz, and continuous transmit coverage from 1.5 to 30 MHz. Drake's use of up-conversion to an IF of 48 MHz and their continuous-duty, 150-watt output solid-state PA were also firsts. The TR-7 definitely established the trend towards the equipment we have today.

However, the TR-7 was not without its problems. Among these was the fact that, while the radio did offer continuous receive coverage down to 10 kHz, sensitivity below 1.5 MHz was poor. For the dyed-in-the-wool CW operator, the radio didn't offer full break-in (QSK) operation and the receive signal-to-noise (S/N) ratio was poor on weak CW signals. Finally, many amateurs found the noise from the PA cooling fan annoying.

Yet, with a few revisions, these problems in the TR-7 can be corrected. The four modifications I designed provide the following features:

- Full receive coverage from 10 kHz to 30 MHz with no reduction in sensitivity across the entire range.
- Smooth, quiet, full break-in (QSK) CW operation at full output power.
- A major improvement in the S/N ratio of weak CW signals.
- A reduction in cooling fan noise, because the fan operates only when it is needed.

One of my primary goals in designing modifications is that they are reversible. Nothing bothers me more than extra holes in front panels, cutouts in chassis, cut PC board traces, etc. All four modifications described in this article are 100 percent reversible and don't require the drilling of any holes. The radio can be returned to its stock configuration at any time, should this ever become necessary.

Theory of the design deficiencies

Receiver

The receiver in the TR-7 was state-of-the-art when it was designed in 1978. Sensitivity, selectivity, and stability are all very good. However, since the TR-7 was primarily designed for use as an HF transceiver, certain compromises were made in its design.

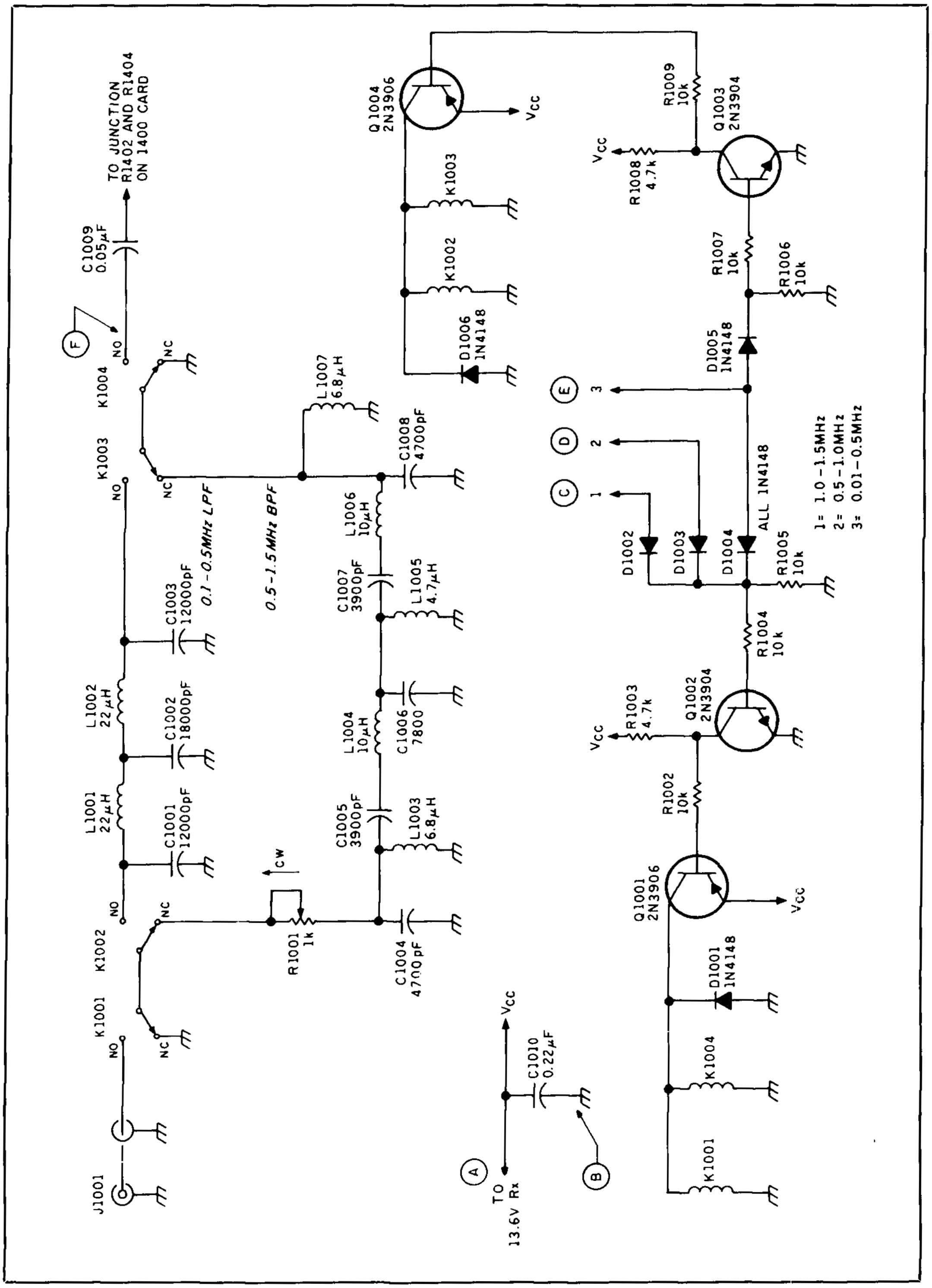

Figure 1. Front-end filter.

One problem is the poor sensitivity below 1.5 MHz. Drake advertised the radio as being capable (with the addition of the AUX-7 module) of receiving continuously from 0.01 to 30 MHz, with reduced performance below 1.5 MHz. However, TR-7 owners with AUX-7s learned that the receiver sensitivity below 1.5 MHz was far worse than the advertising implied. They were further disappointed to learn that the antenna input (marked VLF antenna on the schematic) was inconveniently located on the accessory connector.

The poor performance results from the

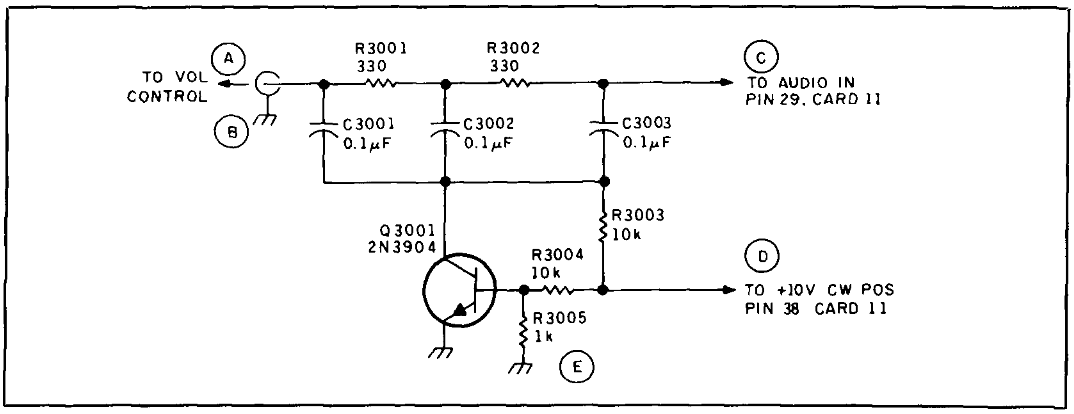

Figure 2. Audio filter circuit.

lack of any low-pass filtering below 1.5 MHz. Instead, Drake inserted a 20-dB fixed attenuator in series with the antenna to protect the mixer against overload in the 0.01 to 1.5-MHz range.

Another minor problem with the receiver is the annoying high-frequency noise present in the audio, which is especially noticeable when operating with narrow CW filters.

Transmitter

The transmitter in the TR-7 is excellent. The transmitted audio is clean, and the PA easily develops a solid 150-watt output across the entire 1.8 to 30-MHz range. Its main drawback is the lack of full break-in (QSK) CW.

The TR-7 manual states that with the CW delay control set fully counter-clockwise, the radio operates in the QSK mode up to about 20 WPM. Those who have tried this have been greatly disappointed. The chatter of the frame-type T/R relay is irritating, and the rapid transition between transmit and receive creates annoying thumps and pops in the speaker.

One other problem with the transmitter is that the original Drake FA-7 fan used to cool the PA operated whenever the set was on, whether it was needed or not. In addition, because it is an AC-powered fan, it wouldn't work at all when the TR-7 was used as a mobile or portable station.

Circuit description

Receiver front end

The installation of one low-pass and one band-pass filter in the front end solves the problem of poor performance below 1.5 MHz. I used a Chebyshev low-pass filter for the 0.01 to 0.5 MHz range and a Chebyshev bandpass filter for the 0.5 to 1.5-MHz range.[1] The bandpass filter used in the 0.5 to 1.5 MHz range protects the mixer from strong shortwave broadcast signals as well as strong VLF signals, like Loran-C. The AUX PROGRAM switch selects the appropriate 500-kHz frequency range and front-end filter (see **Figure 1**). The circuit is very straightforward, using relays for filter/antenna switching instead of diodes. I used relays to avoid the IMD products diode switches can introduce under strong signal conditions (such as those prevalent in the broadcast band).

Two relays on the front-end filter card select the appropriate filter. The normally closed contacts of relays K1002 and K1003 select the 0.5 to 1.5-MHz bandpass filter when the AUX PROGRAM switch is in position 1 or 2. Variable resistor R1001 adjusts the insertion loss of the 0.5 to 1.5-MHz bandpass filter to minimize broadcast-band intermodulation products. Transistor Q1003 is a driver for transistor Q1004, which actuates relays K1002 and K1003 when the AUX PROGRAM switch is placed in position 3. This enables the selection of the 0.01 to 0.5-MHz low-pass filter.[2]

Because the VLF/LF/MF antenna(s) may be left connected to the set at all times, transmitting with the TR-7 above 1.5 MHz could induce enough voltage in the low-frequency antenna to damage the filter or the mixer. Two additional relays on the front-end card prevent this type of damage. The normally closed contacts of relay K1001 disconnect the low-frequency antenna from the filter input, and the normally closed contacts of relay K1004 disconnect the mixer from the filter output.

Whenever the AUX PROGRAM switch is in one of the three positions associated with the 0.01 to 1.5-MHz range, a logic high is applied to the diode OR gate comprised of D1002-D1004. A logic high from this OR gate turns on Q1002, which turns on Q1001 and actuates relays K1001 and K1004. With K1001 and K1004 actuated, the appropriate

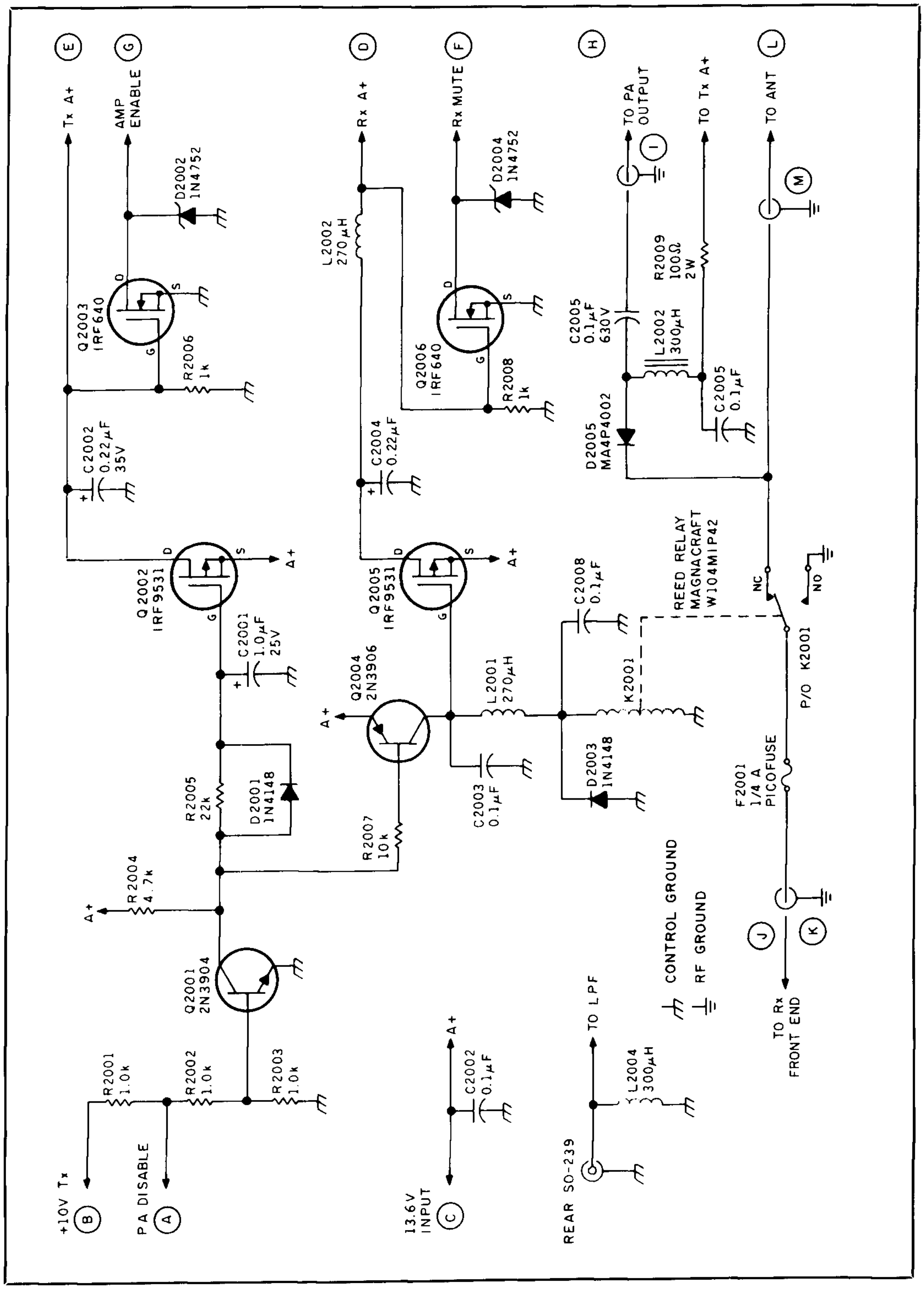

Figure 3. QSK card schematic.

front-end filter is placed in the low-frequency antenna circuit. The OR logic allows the use of this front end in TR-7s where a portion of the AUX-7 card is currently used for WARC bands, MARS frequencies, etc. Therefore, the three low-frequency ranges can be associated with any three positions on the AUX PROGRAM switch.

The inconvenient location of the low-frequency antenna connection on the accessory receptacle makes its relocation desirable. Fortunately, Drake made an unused phono jack available on the rear panel. You can substitute a BNC connector in place of this phono jack for use with the low-frequency antenna.

CW audio filter

The audio filter circuit is quite simple, yet it's very effective in reducing high-frequency noise when using narrow CW filters. The circuit is comprised of a simple RC filter[3] with a cutoff frequency of approximately 1500 Hz, and a simple transistor switch (see **Figure 2**). The filter is inserted into the high-impedance audio path from the AF GAIN control to the input of the audio amplifier.

When the MODE switch is in any position other than CW, transistor Q3001 is turned off, allowing the common for capacitors C3001-3003 to float at approximately 10 k above ground, disabling the low-pass filter. Resistors R3001 and 3002 are of such a low value compared to the input impedance of the amplifier that they introduce no discernable loss. Q3001 is turned on whenever pin 38 on the second IF/Audio card goes to + 10 volts when the MODE switch is in CW. This grounds the common for the RC filter, placing it into the circuit. Resistors R3004 and 3005 form a 10:1 voltage divider for the base of Q3001 to prevent it from turning on with the residual voltage (approximately 1 volt) present on pin 38 when the MODE switch is in any position other than CW.

Transmitter QSK

Modifying the radio to operate QSK was a bit of a challenge, due to Drake's use of a 4PDT relay for T/R switching. The functions of the T/R relay contacts are:

1) Switching A + from transmit to receive.
2) Switching the antenna from the receiver to the transmitter.
3) Protecting the mixers in the internal and external receivers.
4) Providing control for a linear amplifier.
5) Providing a means to mute an external receiver.

By emulating all of the functions of the T/R relay, the modified TR-7 can function in exactly the same transceive configuration as a stock unit.

My QSK circuit employs a high-power PIN diode, a reed relay, and several power MOSFET switching components (see **Figure 3**). Power MOSFETS are the easiest way to switch voltage quickly and quietly as required by QSK operation. Their speed and very low "on" resistance make them a perfect replacement for a mechanical relay. Both Q2002 and Q2005 replace the original RX/TX A + switching contacts. I used P-channel power MOSFETS for A + switching and N-channel devices to emulate the relay contact closures to ground for external control. The N-channel MOSFETS in this circuit were selected for their low "on" resistance (0.18 ohms).

The circuitry on the base of the transistor Q2001 exactly duplicates the existing TR-7 relay-driver circuit.[4] Every time the unit enters the transmit mode, + 10 volts is applied to R2001. Provided the PA Inhibit line isn't low, transistor Q2001 is turned on. When Q2001 turns on, its collector goes low, turning on Q2002 through an RC timing circuit comprised of R2005 and C2001. This delay circuit allows the receiver protection relay to switch prior to the generation of output power from the transmitter.

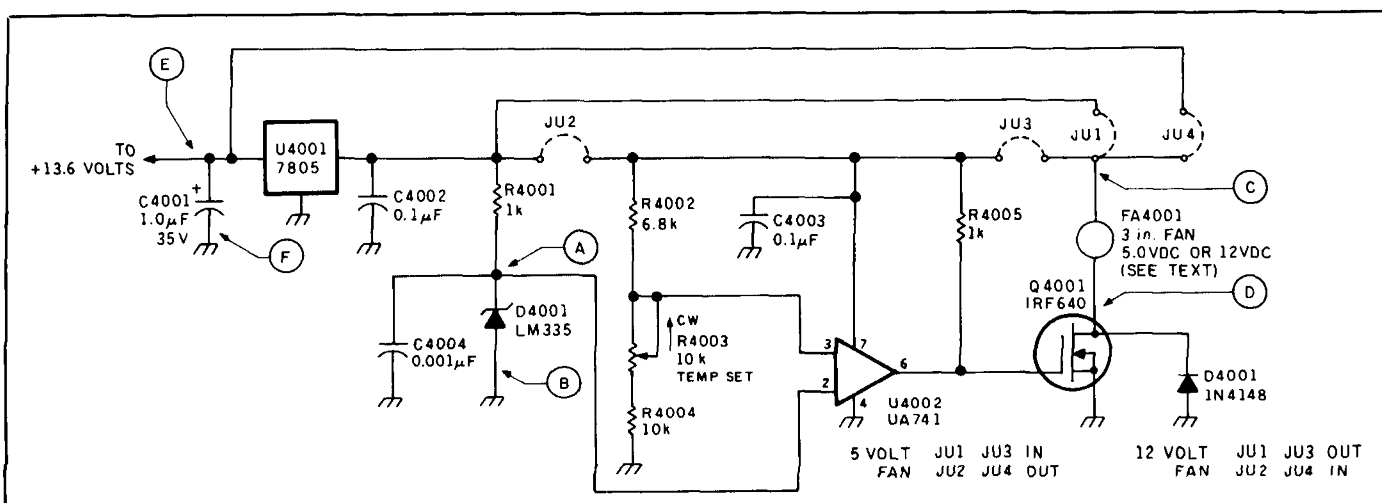

Figure 4. Fan control circuit.

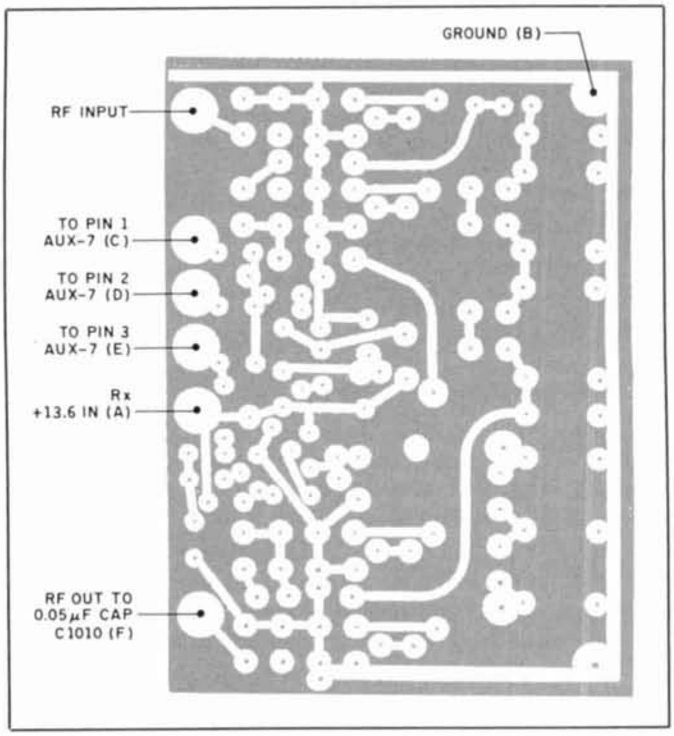

Figure 5. Front-end card foil layout.

When the collector of Q2001 goes low, Q2004 turns on, pulling the gate of Q2005 high and turning off A + to the receiver. Q2004 also actuates reed relay K2001, grounding the receiver input. The gate for MOSFET Q2003 is tied across the TX A + line. When the TX A + line is high, Q2003 turns on, which supplies a ground to control an external linear amplifier. Zener diode D2002 clamps the drain of Q2003 to ground, protecting it from high drain-source voltages.[5]

Transmit/receive antenna switching is accomplished through the use of a high-power PIN diode and a reed relay. Both internal and external receiver front ends are protected by shorting them to ground through relay K2001. I used a reed relay for receiver protection because the broadband design and space limitations of the TR-7 preclude the installation of the necessary quarter-wave transformer for a shunt PIN switching diode.

The high-power PIN diode, D2005, is a low "on" resistance, long carrier-life device made by Microwave Associates (M/A-

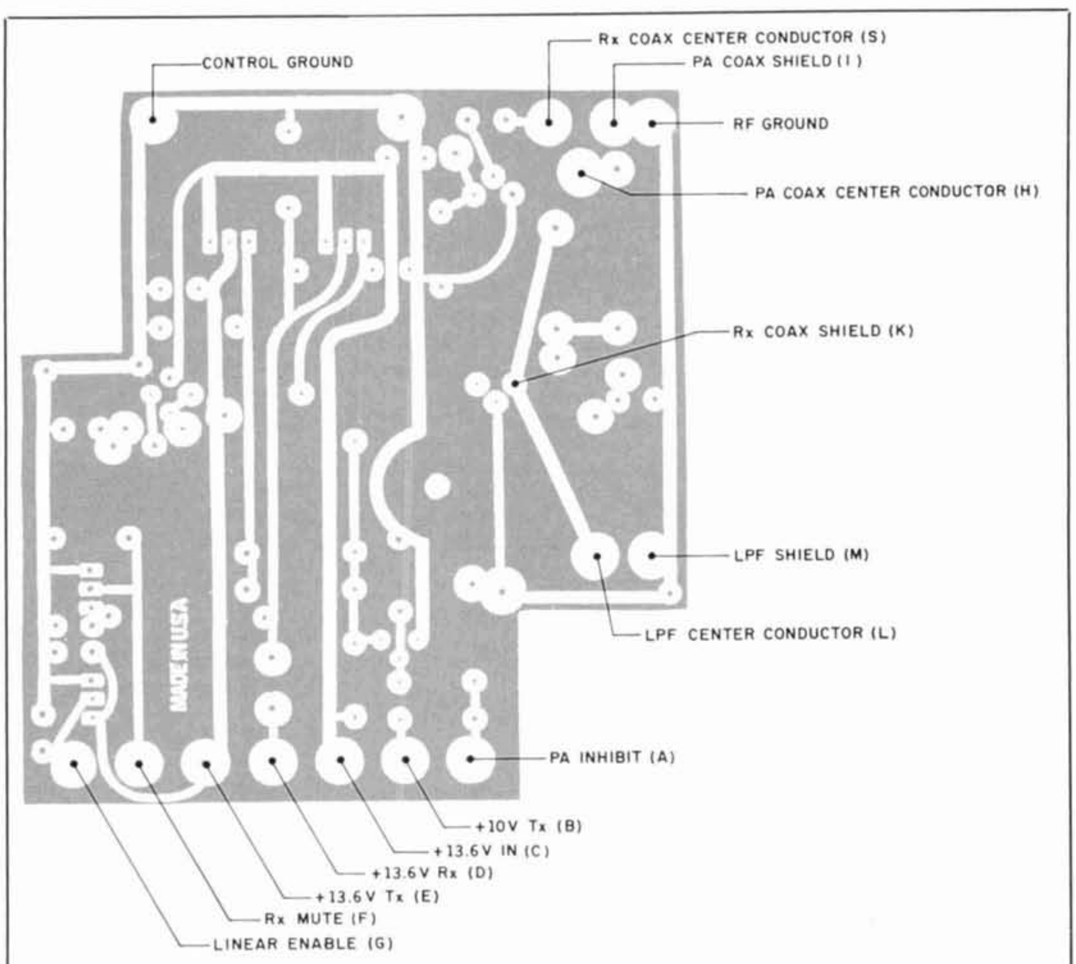

Figure 6. QSK card foil layout.

COM). It provides a high degree of isolation between the PA and the antenna circuit except when forward biased. Capacitor C2005 provides DC isolation for the forward bias voltage injected through RFC L2003. Resistor R2009 limits the forward bias current in D2005 to 135 mA. A second RFC is tied across the antenna jack, providing a low resistance DC ground return for D2005.[6]

During receive, transistor Q2001 is turned off. The collector of Q2001 goes high, turning off Q2002. Diode D2001 speeds up the turn off of Q2002 by bypassing R2005 during receive. With Q2001 turned off, Q2004 turns off, turning on MOSFET Q2005. This turns on A+ to the receiver. At the same time, K2001 is no longer actuated, connecting the antenna low-pass filter to the input of the receiver. When the RX A+ line is high, MOSFET Q2006 turns on, which in turn supplies a ground to the RX Mute line, enabling an external receiver. Zener diode D2004 acts as a voltage "clamp" to protect Q2006 from high drain-source voltages.

One final modification speeds up the transfer time from transmit back to receive. A stock TR-7 with the CW DELAY control fully counter-clockwise switches from transmit to receive in about 50 ms. Changing R310 on the TX Exciter card from 47 k to 22 k shortens this transition time to about 25 ms.

I made no modifications to the receiver AGC time constants. In the FAST AGC position in the CW mode, the AGC time constant is only 40 ms. This should be fast enough for most break-in operation.

Fan control circuit

The fan control circuit is comprised of a temperature-sensitive zener diode and a simple op-amp voltage comparator (see **Figure 4**). The output voltage from zener diode D4001 varies at a constant 10 mV/degree Kelvin. At 25 degrees Celsius, the voltage from D4001 is approximately 3 volts.[7] Resistors R4002 through 4004 serve as a variable voltage divider, with a range from approximately 3.0 to 3.7 volts. This voltage range corresponds to a temperature range of approximately 25 to 97 degrees Celsius. Voltage from the temperature sensor and the voltage divider are fed into pins 2 and 3 (respectively) of the UA741 op amp. Whenever the voltage on pin 2 (from the sensor) goes above the voltage present on pin 3, pin 6 of U4002 goes high, turning on power MOSFET Q4001 and turning on the cooling fan. I used a 5-volt fan in this application as they are readily available at very reasonable

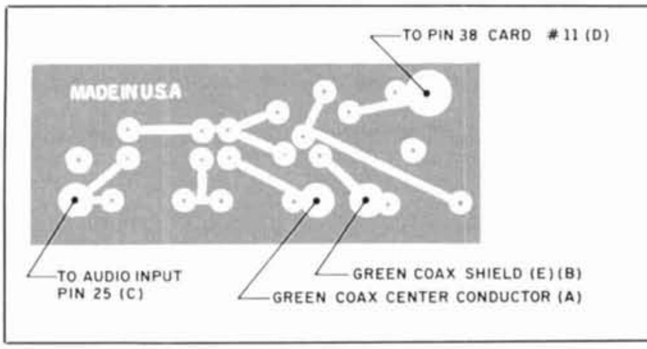

Figure 7. Audio filter foil layout.

prices on the surplus market. However, if you can't find a 5-volt fan, a simple wiring change lets you substitute a 12-volt fan. A DC fan will operate during mobile or portable operation—a feature that wasn't possible with the stock FA-7.

Construction notes

Before you start, choose the modifications you wish to make. *Each of the modifications are independent, so I highly recommend that you put them in and test them one at a time.*

I've made every effort to be as thorough as possible in my construction and installation procedures, but I couldn't cover every aspect of this project. Be careful, take your time, and remember that you install these modifications at your own risk.

If you decide to make any of the modifications described here, you must construct the appropriate circuit boards shown in **Figures 5, 6, 7,** and **8**, and **Photos A, B, C,** and **D**. You may use the printed circuit board artwork from this article, or create

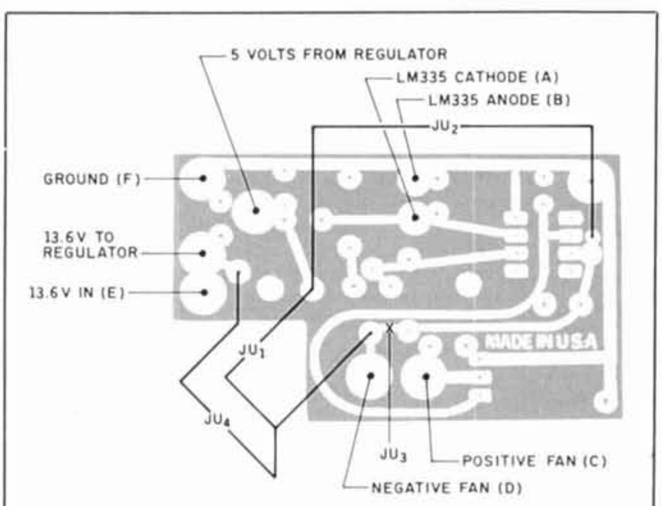

Figure 8. Fan control foil layout.

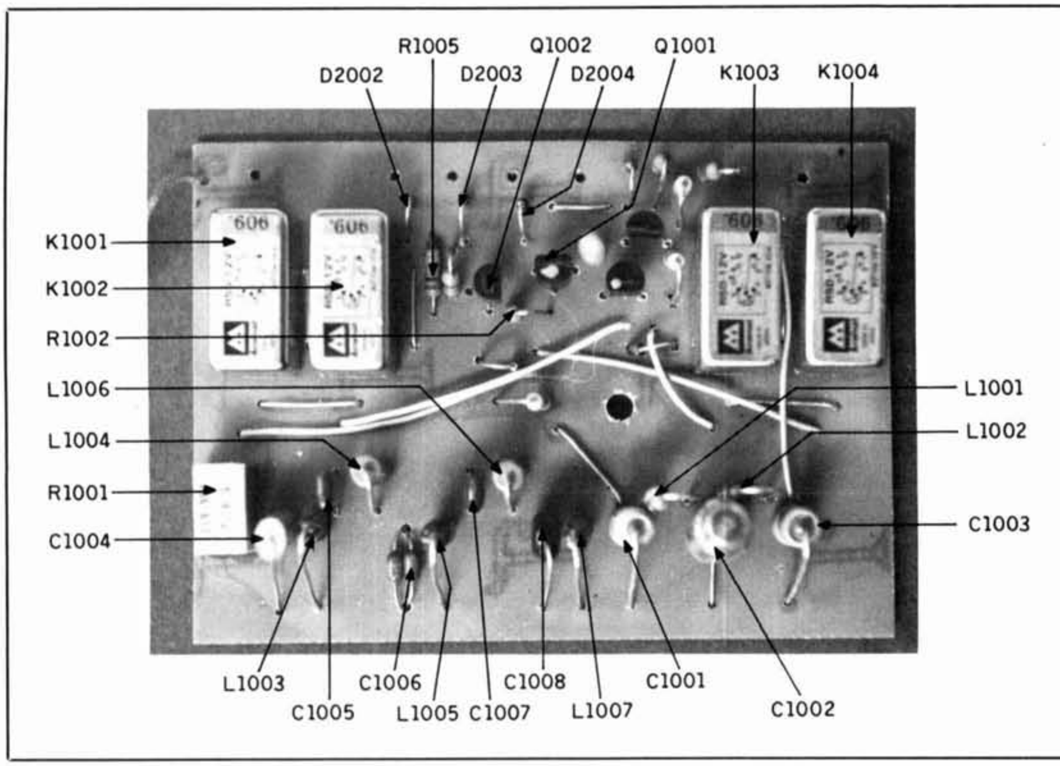

Photo A. Front-end card layout.

your own. I made all of my boards using the direct-etch method. If you decide to use this method as well, try to place all components exactly as they are shown on the artwork. Space is tight in the TR-7, so the board sizes and layouts are rather critical. This is especially true for the front-end and QSK cards, because the bandswitch passes over the center of each. If you must make parts substitutions, do so with some care; poor performance, inadequate space, or damage to the radio may result.

Make sure that your ground for the receiver coax duplicates the grounds shown in **Photos E** and **F**. It's important to keep the RF ground separate from the control ground on the QSK card. Ground the board to the chassis only to the two points mentioned in the text. Because of the power levels involved, ground loops may cause parasitic oscillations and/or receiver damage if you don't heed this advice. If you design your own QSK card layout, I strongly recommend that you test the circuit with a spectrum analyzer.

Don't attempt to use a higher permeability core with fewer turns for L2003 and L2004. At the power levels present in the TR-7, a higher permeability core may get so hot that the PA will shut down due to a high VSWR condition.

Finally, don't forget the heat sink for D2005. I used a 3/16-inch hex, 3/8-inch long, 4-40 standoff to secure the mounting stud for D2005. When mounted on this stud, the specified heat sink is sufficient to permit key-down operation at full output power for a minimum of ten minutes at a time. RFC L2003 is secured to the circuit board with a small wire tie through the board, and L2004 is secured to the ground strap for the LPF in the right rear corner of the LPF compartment (as viewed from the rear).

In order to have receive coverage below 1.5 MHz, you must have the DR-7 digital display and the AUX-7 card with the low-frequency PROM installed in your TR-7. If you don't have the AUX-7, or your AUX-7 doesn't have the low-frequency PROM, you must install the programming diodes described in the Drake service bulletin "RTM/RRM7 Replacement." These diodes are simple to install, and can be used to add low-frequency coverage to any TR-7, whether it has the AUX-7 or not. The RTM/RRM7 bulletin is available by sending an SASE with a note requesting the same to:

R.L. Drake Company
P.O. Box 3006
Miamisburg, Ohio 45342
Attn: Bill Frost

Install the diode modification as de-

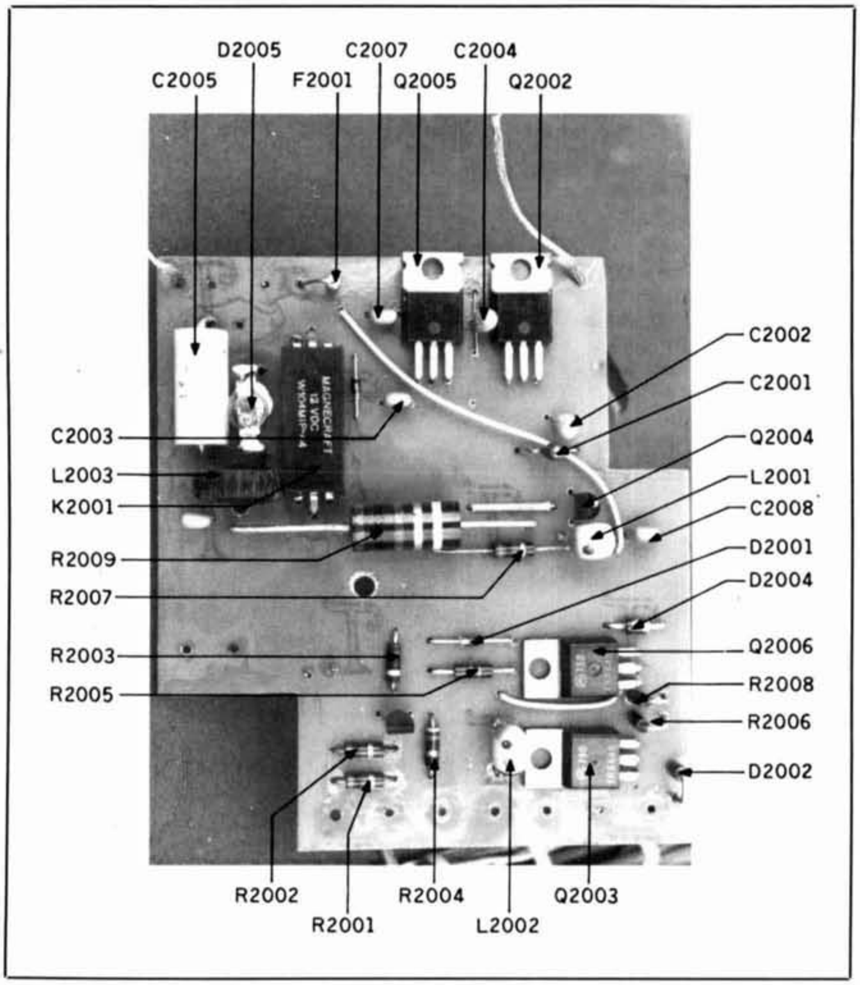

Photo B. QSK card layout.

scribed by Drake and test its operation according to their procedure. You are now ready to install the front-end card.

Installing the new front-end card

Begin by disconnecting the violet coaxial cable from the front low-pass filter switch card (Drake assembly #1400).[8] This coax is in the upper right-hand corner of the 1400 card as viewed from the front. Pull the coax up and out of the way; you'll need it for the new front-end card. Next, cut five, 12-inch long pieces of hookup wire of the following colors: red, black, yellow, blue, and white. Refer to **Figure 1**, and connect the five wires as follows:

1) Solder the red wire to point A.
2) Solder the black wire to point B.
3) Solder the yellow wire to point C.
4) Solder the blue wire to point D.
5) Solder the white wire to point E.
6) Solder one end of a 0.05-μF/100-volt mylar capacitor to point F.

The front-end card mounts inside the TR-7 high-pass filter assembly (see **Photo F**). To install the front-end card, remove the bandswitch knob, then remove the two screws holding the bandswitch detent unit onto the back of the radio. Mark the top side of the detent unit so you'll know how to reassemble the switch. With the detent loose, carefully pull the bandswitch rod out of the radio. Set it aside, along with its hardware and knob, for later reassembly. Next, unsolder the hot and ground wire from the speaker. Remove the four flat-head screws that retain the speaker and carefully remove it from the radio.

Mount the front-end card using a single 1/4-inch hex, 1/2-inch long brass standoff with no more than 1/4 inch of 4-40 thread. Screw the standoff into the short bushing pressed into the shield between the high-pass filter assembly and the main circuit area. Using an internal-tooth lockwasher,

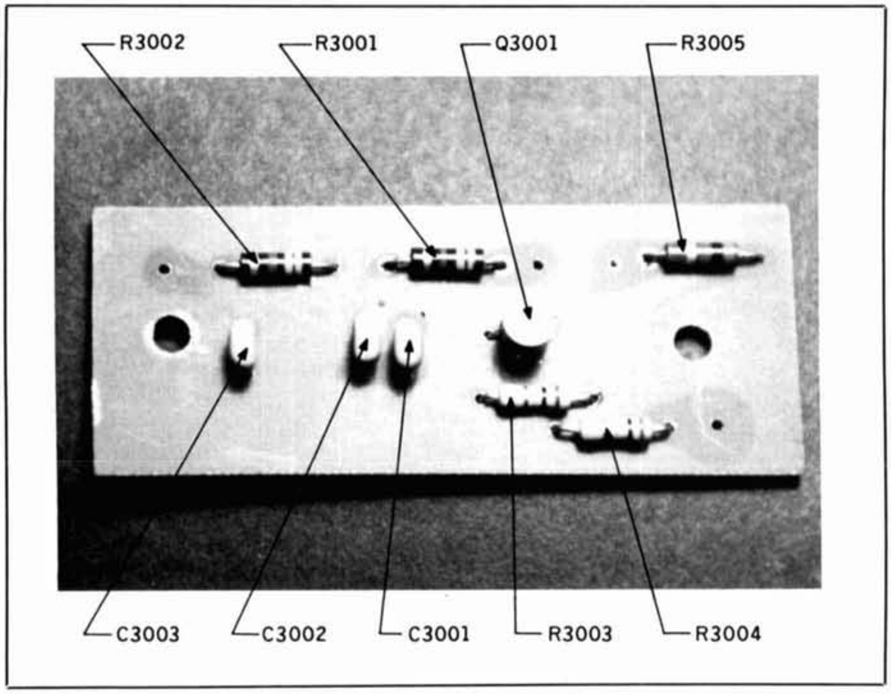

Photo C. Audio card layout.

tighten the standoff into the chassis bushing. After you have set the board in place, you can tighten the retaining screw easily using a screwdriver fed through one of the holes for the speaker grille. With the board mounted, feed the green, blue, and yellow wires between the 1400 card and the chassis. These will be connected to the appropriate points under the chassis in the next step. Leave the red A+ wire above the chassis.

Connect the violet coax that you disconnected from the 1400 card to the input of the new front-end card, with its shield to the ground foil on the 1400 card. Next, solder the free lead of the 0.05-μF capacitor on the front-end card to the junction of R1402 and R1404 on the 1400 card.[4] Solder the ground wire to the ground foil on the 1400 card, then connect the red A+ wire to the receiver A+ line on the 1500 card.[4] After you've made the A+ connection, turn the radio upside down with the front panel facing you. Connect the yellow lead to pin 1 of the AUX-7 card. This is the first card slot from the front of the radio. To find pin 1, count from the far right pin with the radio upside down and the front facing you. Connect the green wire to pin 2 and the blue wire to pin 3 of the AUX-7 card slot—again counting from the far right. Dress the wires into the existing harness with wire ties and turn the radio over.

Carefully reinstall the bandswitch rod into the TR-7. Make sure that you have the rod oriented according to the marks you made prior to its removal. *Be very careful when installing the rod to ensure that all switch wafers are lined up. Never force the rod should it become stuck during reinstallation.* Once you've inserted the rod through all wafers of the switch, reinstall the detent hardware and the bandswitch knob. Reinstall the speaker and reconnect its wires.

Finally, you may want to change the low-frequency antenna connection point on the rear panel. In its stock configuration, the low-frequency antenna was brought out on pin 7 of the ACCESSORY connector. Drake provided a spare phono connector on the rear panel that can be used for this antenna connection. Personally, I detest phono connectors, so I replaced it with a BNC connector (see **Photo G**). This is optional; if you feel that the phono jack will suffice, use it. If you prefer the BNC, you must enlarge the hole.

First, remove the four screws that hold the rear subpanel to the chassis. Gently pull this subpanel away from the radio as far as the leads will allow. If you are going to use a BNC connector, remove the unused phono jack immediately below the EXT RCVR jack. *Using extreme care, ream or drill this hole out to 3/8 inch.* Mount a BNC connector and grounding lug in this new hole. Next, using a narrow soldering iron with a small chisel tip, remove the violet coaxial cable from pin 7 of the ACCESSORY connector, and remove the shield from the

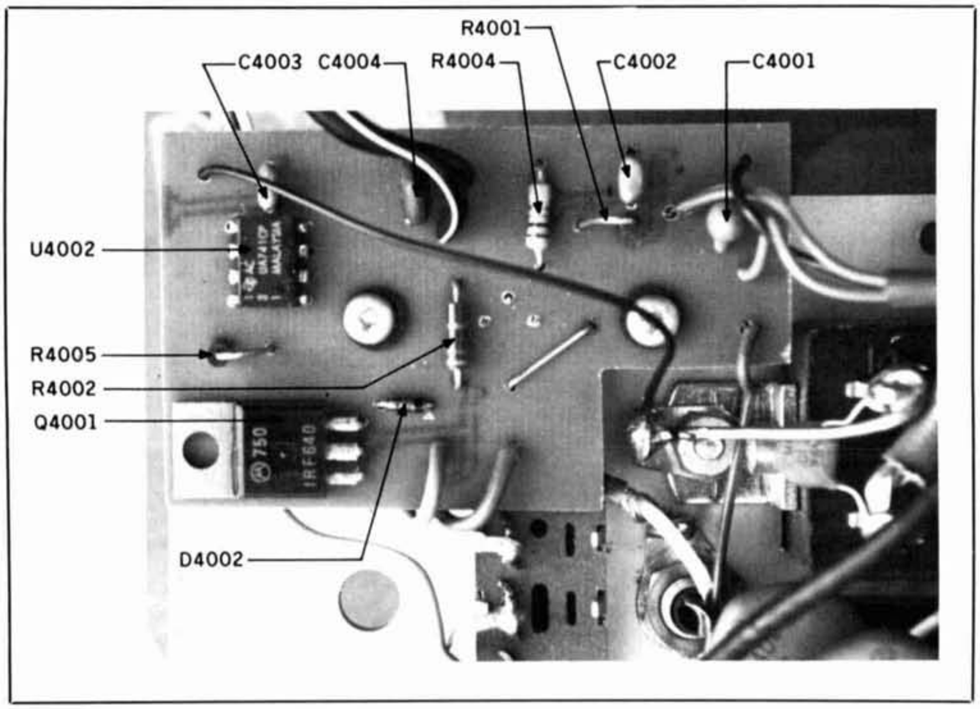

Photo D. Fan control layout.

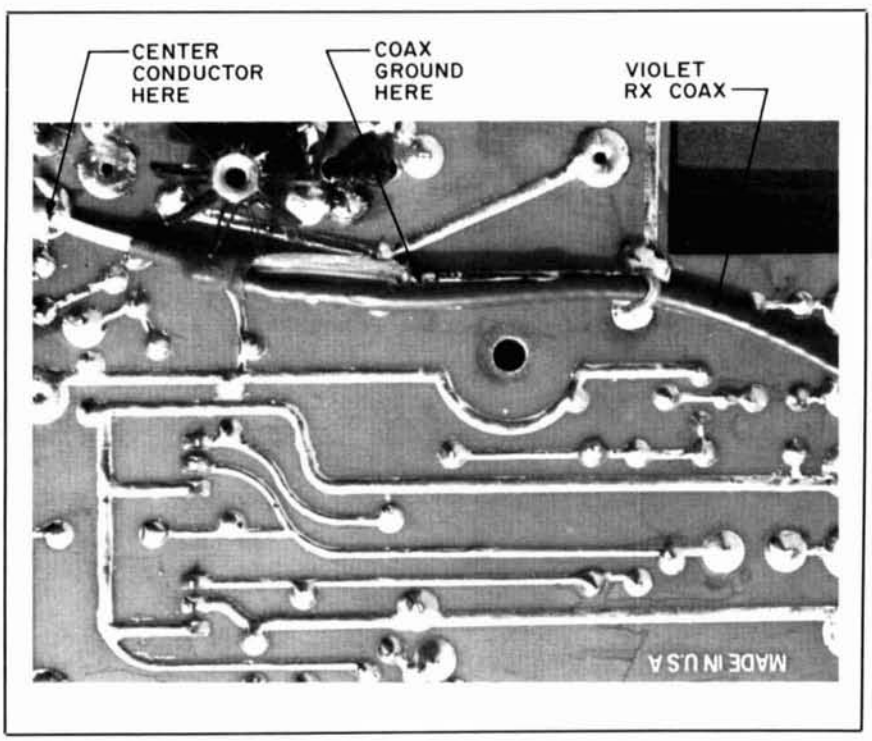

Photo E. QSK card RX coax grounding.

ground lug. Solder the center conductor of this violet coax to the center pin of the BNC connector, and solder the shield to the ground lug. Replace the subpanel, taking care not to pinch any of the wires that lead to it. The front-end card installation is now complete.

Front-end card testing

Note: Do not apply power to the set until instructed to do so, or damage may result.

Carefully inspect the wiring to the front-end card for melted insulation or loose wire strands. Turn the radio upside down and

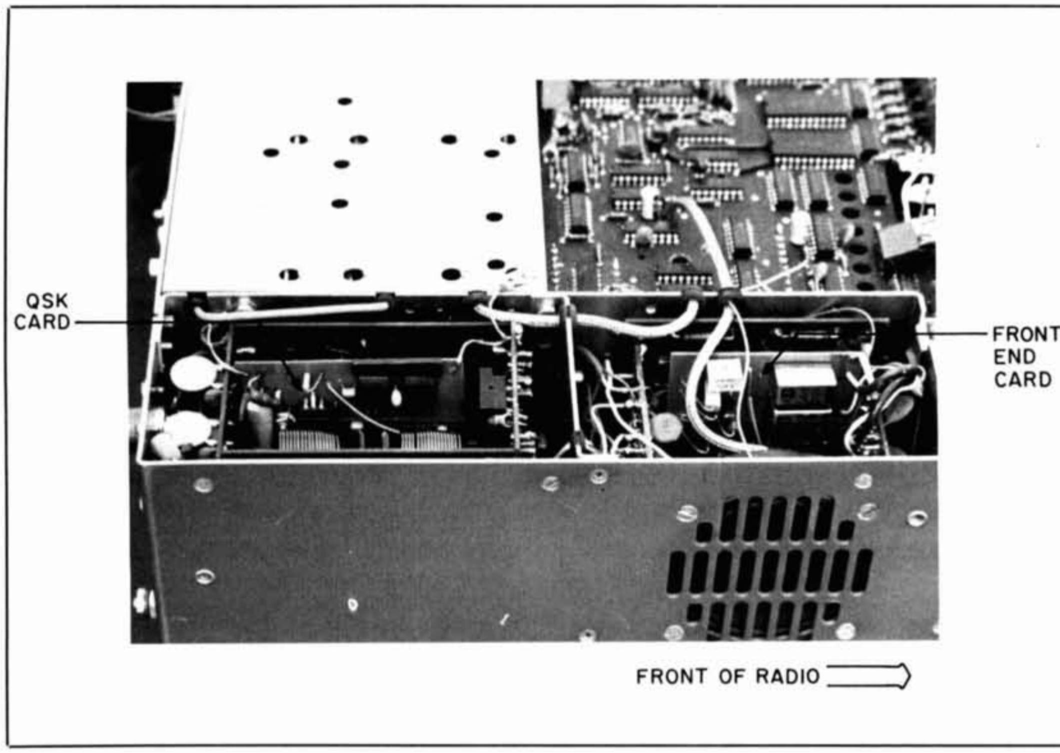

Photo F. QSK card and front-end card location.

shake out any remaining wire or drill fragments.

Preset the front-panel controls as shown in Table 1.

Connect an HF antenna to the rear-panel SO-239 connector. Then connect the power supply, and turn the radio on. The radio will power up on 7.500 MHz, and control functions should be normal. Tune around the 40-meter amateur band and make sure that HF operation of the radio is unaffected.

Place the AUX PROGRAM switch into position 1. Set the bandswitch to 1.5 MHz and set the tuning dial to 500. The display should read 1.500 MHz. Connect a low-frequency antenna to the BNC connector on the rear panel. When tuning down from 1.5 MHz, you should hear numerous broadcast signals. Next, select position 2 on the AUX PROGRAM switch. Tuning from 0.5 to 1.0 MHz, you should again hear numerous broadcast stations. Finally, select position 3 on the AUX PROGRAM switch. Relays K1002 and K1003 should now actuate, and if your antenna is efficient below 0.5 MHz, you should hear numerous aircraft beacons and similar low-frequency stations.

Control	Setting
RF GAIN	Full clockwise
AF GAIN	Full counter-clockwise
MIC GAIN	Full counter-clockwise
CARRIER	Full counter-clockwise
MODE	LSB
AUX PROGRAM	NORM
PBT	OFF
RIT	OFF
CAL	OFF
NB	OFF
PTT/VOX	PTT
REF/FWD	FWD
RCT	OFF
BAND	7 MHz
TUNING	500

Table 1. Front-panel control settings.

In the event any of the bands don't work, recheck your wiring to the AUX-7 card slot. Pins 1, 2, and 3 on the AUX-7 card slot go high whenever positions 1, 2, or 3 are selected on the AUX PROGRAM switch. Whenever any of the three low-frequency positions on the AUX PROGRAM switch are selected, relays K1001 and K1004 should be actuated. Relays K1002 and K1003 should be actuated only when the AUX PROGRAM switch is in position 3.

Operation

Operation of the front-end card is quite simple. To receive in the 10 kHz to 1.5 MHz range, simply select the appropriate 500-kHz band with the AUX PROGRAM switch. You can set the main bandswitch to any band except 21 or 28.5 MHz. The low-frequency antenna can remain connected to the set at all times, if you so desire.

QSK card installation

Installation of the QSK card is slightly more difficult than the front-end card, as it requires the removal of the low-pass filter switch board (Drake Assy #1900).

To begin the QSK card installation, turn the radio around so the back is facing you. Remove the hold-down wire on T/R relay K1901, remove it from its socket, and set it aside. Remove the bandswitch knob, and mark the top side of the bandswitch detent unit on the rear panel for reassembly later. Slowly remove the bandswitch rod from the high and low-pass filter assemblies and set it aside along with the knob and hardware.

Unsolder the three ground braids to the LPF assembly, and gently pull the front end of the LPF unit up so the wires on the 1900 card are accessible.[9] Using a soldering iron with a long, narrow tip, remove the following wires from the 1900 card *in the order listed*:

1) Grey coax at upper left of card.
2) Black coax at upper left of card.
3) Violet coax at upper left of card.
4) Pink wire with ferrite bead in upper right of card.
5) Tan wire in upper right of card.
6) Blue wire in upper right of card.
7) Red wire in upper right of card.
8) White wire in upper right of card.
9) Both orange wires in upper right of card.
10) Green wire in upper left of card.

After removing these wires, carefully pull the wire harness down through the bottom of the radio. This will bring all of the wires listed above (except 1, 4, and 5) into a position where you can connect them to the new QSK card. *Use care when performing this step; it is very easy to get one of these wires caught on the 1900 card switch wafer.* Cut the wire ties that held these wires in a bundle behind the 1900 card. Wires 4 and 5 are too short to be pulled down under the chassis. Instead, remove them from their connection points on the rear high-pass filter switch card (Drake Assy #1500) *making a note of the position of each wire.* Be careful not to lose the ferrite bead off the pink wire. Cut a 12-inch piece of both pink and tan hookup wire, and connect each to the appropriate locations on the 1500 card.[4] Make sure you don't forget to put the ferrite bead on the pink wire. Bundle these two wires together and feed them under the chassis through the LPF compartment. Carefully reinstall the LPF assembly in its compartment and resolder the three ground wires to their connection points.

The QSK card mounts on a standoff inside the low-pass filter assembly (see **Photo F**). The ideal hardware for this is a 1/4-inch hex, 3/4-inch long brass standoff with no more than 1/4 inch of 4-40 thread. Thread the standoff (along with an internal-tooth lockwasher) into the bushing pressed into the side wall of the LPF compartment.

You must solder three coaxial cables to the QSK card prior to its installation. These three cables are the ones that you disconnected earlier from the 1900 card (see **Figure 3**). Connect the center conductor of the black coax to point H on the QSK card, with the shield connected to point I. Disconnect the short grey coax from the 1900 card, this will be reconnected later. Solder the center conductor of this coax to point L and the shield to M. Finally, connect the center conductor of the violet coax to point J and the shield to point K. *Make sure the shield on the violet coax is grounded exactly as shown in* **Photo E**, *or ground loops may occur.*

The most difficult part of the QSK card installation is positioning the card and securing it to the standoff installed previously. Because there isn't any easy way to get a screwdriver into the LPF compartment, I recommend using a 4-40 bolt with a 3/16-inch hex head. You can tighten this type of bolt with a pair of long-nose pliers without too much difficulty. To install the card, carefully slide it into place, making sure you are on the proper side of the small bolts holding in the rear LPF switch wafer. Also make sure you don't damage the piston trimmer capacitor on the rear LPF card with the heat sink stud for D2005.

Install the center hold-down bolt as described above. Reinstall the bandswitch rod, taking care to align the detent unit with the mark made previously. *Never force the bandswitch rod when installing it. The switch wafers are easily damaged by excessive force.* Reinstall the detent-unit hardware and the bandswitch knob. Turn the radio upside down with the rear panel toward you, and connect the following wires to the QSK card (see **Figure 2**):

1) Connect the two orange wires to point A.
2) Connect the white wire to point B.
3) Connect the green wire to point C.

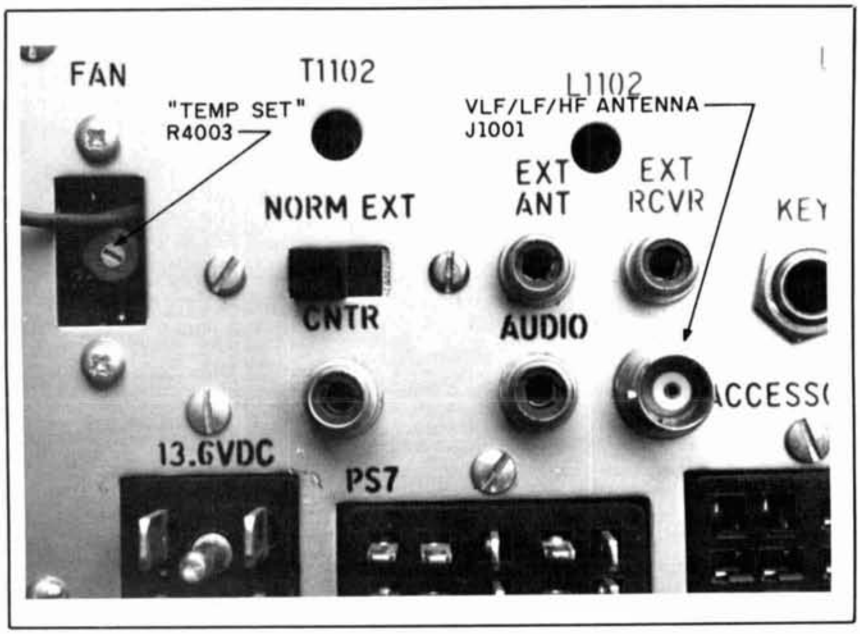

Photo G. Rear panel layout.

4) Connect the tan wire to point D.
5) Connect the pink wire to point E.
6) Connect the blue wire to point F.
7) Connect the red wire to point G.

Finally, turn the radio rightside up and attach a ground lug to the screw near L701 on the RF compartment cover. Next, run a 1/8-inch ground braid from the control ground on the QSK card to ground on the 1900 card. From this same ground point on the 1900 card, run another 1/8 inch ground braid to the solder lug just installed. Likewise, run a 1/8 inch ground braid from the QSK card RF ground to the LPF ground lug in the left rear corner of the LPF compartment (as viewed from the rear, see **Photo F**). Bundle the tan and pink wires together and dress them against the chassis. *Do not bundle the tan and pink wires with the other wiring to the QSK card or spurious emissions may result.* Bundle the orange, white, red, green and blue wires into the main chassis cable harness. Install the heat sink for diode D2005, taking care that it doesn't touch the inductors on the LPF card behind it. Finally, install a 300-uH toroidal choke from the center pin of the rear panel SO-239 antenna connector to ground.

To finish up the QSK modification, you need to change the value of resistor R310 on the Transmit Exciter board to speed up the minimum transmit-receive switching time and add an RF bypass capacitor to the ALC card. To change R310, remove the DR-7 digital display board (refer to the Drake service manual for specifics).[10] Then remove the shield from the VCO compartment (this is necessary to prevent damage to C307). Remove the Transmit Exciter card (the second card forward from the front of the VCO cage) and locate R310 (see **Photo H**).[11] Replace the original 47-k, 1/4-watt resistor with a 22-k, 1/4-watt device. Reinstall the Transmit Exciter card, taking care to line up the pins correctly. Then reinstall the VCO shield and the DR-7. Turn the radio over and remove the rear screw from the ALC (1600) card. Install a ground lug under this screw and solder the negative lead of a 0.1 uF, 35 volt tantalum capacitor to this lug. Solder the positive lead of this capacitor to the lug on the ALC card where the white and violet wires meet (+10 volts transmit). This completes the installation of the QSK card.

QSK card testing

Note: Do not apply power to the set until instructed to do so, or damage may result.

Carefully inspect the wiring to the QSK card for melted insulation or stray wire strands. Turn the radio upside down and shake loose any wire fragments. Once you've done this, preset the controls according to the settings in **Table 1**. Connect the power supply and a 50-ohm, 150-watt dummy load to the radio. Turn on the set, and ensure that the operation of the unit appears normal. The frequency display should show 7.500 MHz, and the unit should be in the receive mode. *Turn the radio off im-*

mediately if there are any unusual indications at this point.

Connect a key to the key jack on the rear panel. Place the MODE switch in the CW position. Connect a voltmeter to the tan wire on the 1500 card. You should measure about 13 volts during receive. Move the voltmeter to the pink wire on the 1500 card and depress the key; you should see about 13 volts during transmit. If either of these two voltages are very low or missing, refer to "QSK card troubleshooting" at the end of this section.

If all tests look good up to this point, hold down the key and slowly turn the CARRIER control clockwise until the internal wattmeter indicates 50 watts output power. Make sure the neon surge protector on the rear panel (above an RF choke near the key jack) is not illuminated. With the CW DELAY control fully counter-clockwise, release the key. The radio should revert to receive instantaneously. Turn the CW DELAY to mid-position, and briefly depress the key. The radio should switch into transmit, hang in this mode for a second, and switch back. If an external wattmeter is available, connect it between the TR-7 and the dummy load. With 50 watts output showing on the front-panel meter, you should read about 50 watts into the dummy load. Advance the CARRIER control until the front panel wattmeter reads 150 watts and the ALC light is on. Hold the radio keyed for two minutes. With the radio unkeyed, check for overheating of any of the components on the QSK card.

If all is well up to this point, you are ready to check the operation of the second set of MOSFET switches. Connect an ohmmeter from pin 11 on the ACCESSORY connector to chassis ground. With the TR-7 in the receive mode you should see a short; in transmit the circuit should be open. Move the ohmmeter from pin 11 of the ACCESSORY jack to pin 9 of the PS-7 jack. Here you should see an open circuit during receive and a short to ground during transmit. If the radio passes all of the tests listed above, the QSK card is working properly.

Operation

To use the TR-7 in the QSK mode, simply turn the CW DELAY control fully counterclockwise. If you wish to operate semi break-in, advance the CW DELAY control clockwise until you achieve the most comfortable "hang time" from transmit to receive. When using the TR-7 into an antenna of unknown impedance, always reduce output power with the CARRIER control. The PIN diode and PA were designed to handle high VSWR conditions, but it's always best to reduce the stress on these components. As a final note, the PIN diode circuitry and heat sink were designed for continuous key-down operation, like RTTY. If you operate in these 100 percent duty-cycle modes, you must have a low VSWR to the radio and a cooling fan for the PA.

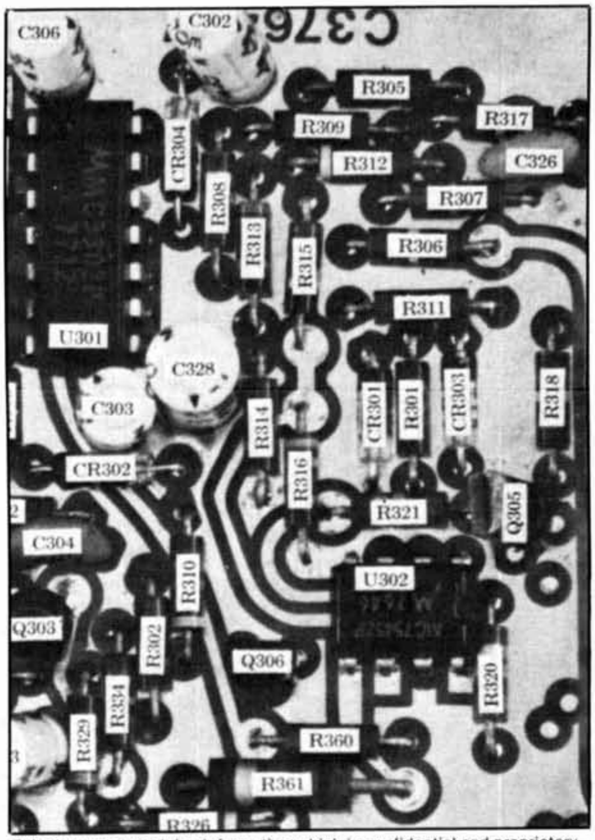

Photo H. Transmit Exciter board.

QSK card troubleshooting

If the QSK card doesn't function properly, check the status of the PA Inhibit line. If it's low, all transmit functions will be inhibited. A PA Inhibit signal is generated when the synthesizer becomes unlocked or the external VFO (if used) is in the SPOT mode. If the status of the PA Inhibit line is correct, recheck all wiring, making sure that the tan and pink wires aren't reversed. Also check transistors Q2002 and 2005 to ensure that they are neither shorted nor open.

Audio filter installation

The audio filter card is very simple to install. It mounts with two 1/2-inch long,

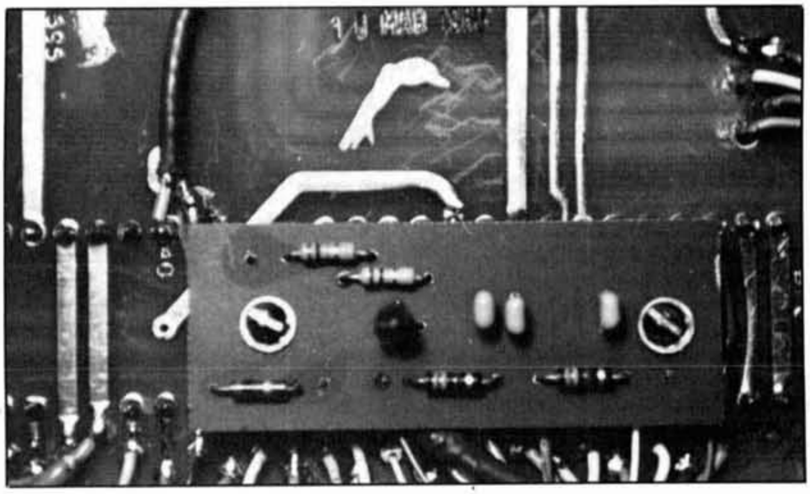

Photo I. Mounting the audio filter card.

4-40 bolts and two 1/4-long nylon standards (see **Photo I**). To install the card, turn the radio upside down with the rear panel facing you. Remove the two 4-40 screws in the right rear corner of the parent board. These screws are near pins 24 and 36 on the second IF/Audio board (card #11) at the extreme rear of the TR-7. Unsolder the green shielded cable from pin 29 of card 11, and unsolder the shield from ground. Cut two short pieces (approximately 2.5 inches) of red, black, and white hookup wire and solder them as follows (see **Figure 2**):

1) White wire to point C.
2) Red wire to point D.
3) Black wire to point E.

With this completed, you're ready to wire the audio filter card. Follow these steps:

1) Solder the white wire to pin 29 of card 11.
2) Solder the black wire to the ground adjacent to pin 29.
3) Solder the red wire to pin 38 of card 11.
4) Solder the center conductor of the green audio cable to point A on the audio filter card.
5) Solder the shield from the green audio cable to point B on the audio filter card.

The audio filter card installation is now complete.

Audio filter card testing

It's simple to test the audio filter card. Turn the TR-7 on and select LSB on the MODE switch. Check the voltage on the collector of Q3001; it should read approximately 0.95 volts. Select the CW position on the MODE switch; the collector voltage should drop to 0.05 volts. Select a narrow CW filter and advance the AF GAIN control. Listen to the noise in the speaker when in the CW position versus LSB. If all is working properly, you'll notice a substantial reduction in high frequency "hiss" when in the CW mode.

If the card fails to function properly, check all wiring, and make sure you didn't short the center conductor of the audio cable to the shield. Also make sure you didn't miscount the pins on the second IF/Audio card.

Fan control card installation

The fan control card is very easy to install. It fits in place of the 117 volt AC plug on the rear apron of the TR-7. To install it, turn the radio around so the rear panel is facing you. *With the power supply disconnected,* remove the four 4-40 bolts holding the center panel of the rear apron to the main chassis. *Carefully* fold the center panel down so it's almost horizontal. Remove the two 4-40 bolts holding the fan power plug to the panel. Cut the wires from this plug at the PS-7 control connector. Remove the fan plug; it is no longer needed.

Next, remove the second IF/Audio card from the radio. Remove the FA-7 fan. Mount a 7805 regulator IC to the chassis wall behind the PA. There's a hole immediately below the two coaxial-cable cutouts along the top of this wall that's in the perfect location for this IC (see **Photo J**). Make sure you apply a thin coating of silicon grease to the back of the 7805 prior to mounting it. The temperature sensor IC mounts within the center of the PA heat sink. I soldered the LM335 to the end of a short piece of #18 speaker "zip-cord." I in-

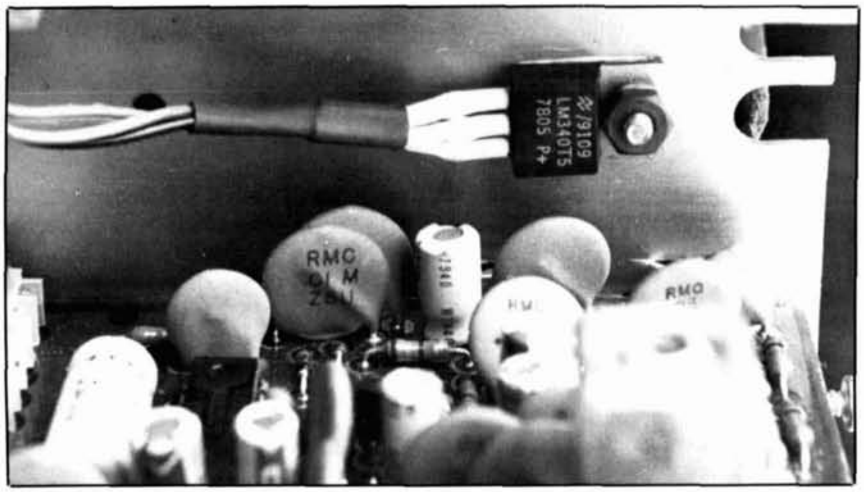

Photo J. 7805 mounting location.

sulated the connections with heat shrink and placed the IC in the center of the heat sink. It isn't necessary to mount this sensor; the temperature rise of the air in the heat sink is quite sufficient. Thread the sensor wires through an appropriately sized grommet and insert this grommet into the unused cutout above the 7805 regulator. Install the new 3-inch, 5-volts DC fan in place of the FA-7. *These fans can be very dangerous; its very easy to accidently get a finger caught in the blades. I strongly recommend the addition of a fan guard.* Thread the wires from the fan through the rectangular fan plug cutout. You are now ready to wire the control card.

Begin wiring by connecting the temperature sensor IC to the card (see **Figure 4**). Solder the lead from the cathode of the LM335 to point A and the anode of the LM335 to point B. Next, solder the positive fan lead to point C and the negative fan lead to point D. Finally, cut an 8-inch piece of red hookup wire, and connect it to point E, and a 3-inch piece of black hookup wire to point F.

Mount the fan control card to the center rear panel using two 3/16-inch hex, 3/8-inch long 4-40 standoffs. These standoffs supply sufficient clearance to mount the circuit board in place of the fan plug. When the card is in place, the fan temperature control will be in the center of the rectangular fan plug cutout (see **Photo G**). After mounting the card, solder the black wire to the ground connection for the four-pin connector, and solder the red wire to pin 9 of the ACCESSORY connector. Carefully replace the center panel and reinstall the four bolts that secure it to the chassis. Installation of the fan control card is now complete.

Testing the fan control card

Before applying power to the TR-7, turn the Temp. Set pot R4003 fully clockwise. This prevents the fan from coming on unless the ambient temperature at the sensor is above 97 degrees Celsius. Turn on the TR-7, and measure the voltage to pin 7 of U4002. This voltage should be 5.0 volts. Next, measure the voltage at pin 2 of U4002. The voltage at this point will vary with temperature, however it should be approximately 3 volts at 25 degrees Celsius. If all checks out up to this point, slowly rotate R4003 (Temp. Set) counterclockwise. If the ambient temperature is above 25 degrees Celsius, the fan will come on at some point in the rotation of this control. To set the pot to correspond to a specific temperature, calculate the LM335 voltage using the following formula:

$V = 10(273.18 + T)$, where T = Temperature in degrees Celsius and V = Voltage in millivolts.

For example, it you wanted the fan to come on at 50 degrees Celsuis, $V = 10(273.18 + 50)$ $V = 3230$ millivolts. R4003 would then be adjusted so pin 3 of U4002 measures 3.23 volts.

In the event the circuit doesn't work, check the voltage from R4003 to make sure it's within the ambient temperature range of the sensor D4001. In this circuit, whenever the voltage on pin 2 of U4002 goes above the voltage on pin 3, pin 6 of U4002 should go high.

Conclusion

These modifications make the TR-7 an excellent performer, both as a "CWbuff's'' radio and as a general-coverage receiver. The independent design of the modifications lets you install them as you see fit. Wherever possible, I tried to use readily available parts. Finally, my design doesn't require the drilling of any holes—allowing you to change the radio back to its stock configuration, should this be required in the future. (A complete set of pc boards is available from FAR Circuits, 18N640 Field Court, Dundee, Illinois 60118 for $16, plus $3.50 shipping and handling. Ed.)

Acknowledgements

I would like to express my sincere appreciation to the people at M/A-COM for their help in the design of the PIN diode portion of the QSK card. I would also like to thank Bill Frost with the R.L. Drake Company for his assistance with documentation.

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2. William E. Sabin and Edgar O. Schoenike, *Single-Sideband Systems and Circuits*, McGraw-Hill, 1987, pages 296-299.
3. Cornell Drentea, *Radio Communications Receivers*, TAB, 1982, page 256.
4. R.L. Drake Company, *TR-7 Service Manual*, 1980, pages 2-87 and 2-104.
5. International Rectifier, *Semiconductor Handbook*, 1981/82, pages A-8 through A-18.
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8. R.L. Drake Co., *TR-7 Service Manual*, 1980, page 2-89.
9. R.L. Drake Co., *TR-7 Service Manual*, 1980, page 2-102.
10. R.L. Drake Co., *TR-7 Service Manual*, 1980, pages 4-4 and 4-5.
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Front-End Card Parts List

Part	Manufacturer	Description
C1001, 1003	(1)	12000 pF, 100 volts DC polyester capacitor
C1002	(1)	18000 pF, 100 volts DC polyester capacitor
C1004, 1008	(1)	4700 pF, 100 volts DC polyester capacitor
C1005, 1007	(1)	3900 pF, 100 volts DC polyester capacitor
C1006	(1)	7800 pF, 100 volts DC polyester capacitor
C1009	(1)	0.05 μF, 100 volts DC mylar capacitor
C1010	(1)	0.22 μF, 25 volts DC tantalum
D1001-D1006	(2)	1N4841 switching diode
K1001-K1004	(3)	subminature SPDT relay (Aromat RSD-12)
L1001, 1002	(5)	22 μH inductor
L1003, 1007	(5)	6.8 μH inductor
L1004, 1006	(5)	10 μH inductor
L1005	(5)	4.7 μH inductor
Q1001, 1004	(2)	2N3906 transistor
Q1002, 1003	(2)	2N3904 transistor
R1001	(6)	1 k potentiometer (Bourns 3386S-1-102)
R1002, 1004		
R1005, 1006		
R1007, 1009	(4)	10 k, 1/4 watt carbon-film resistor
R1003, 1008	(4)	4.7 k, 1/4 watt carbon-film resistor

QSK Card Parts List

Part	Manufacturer	Description
C2001	(1)	1.0 μF, 25 volts DC tantalum capacitor
C2002, 2004	(1)	0.22 μF, 25 volts DC tantalum capacitor
C2003, 2006		
C2007, 2008	(1)	0.1 μF, 100 volts DC monolithic capacitor
C2005	(1)	0.1 μF, 630 volts DC mylar capacitor
D2001, 2003	(2)	1N4148 switching diode
D2002, 2004	(2)	1N4752 33 volt, 1 watt zener diode
D2005	(7)	High-power PIN diode; M/A-COM MA4P4002D
F2001	(8)	1/4 A subminiature fuse
K2001	(9)	SPDT reed relay; Magnecraft W104MIP-42
L2001,2002	(5)	270 μH inductor
L2003, 2004	(10)	300 μH toroid; 70 turns #30 on Amidon T-50-61 core

Part	Manufacturer	Description
Q2001	(2)	2N3904 transistor
Q2002, 2005	(11)	IRF-9531 P-channel power MOSFET
Q2003, 2006	(2)	IRF-640 N-channel power MOSFET
Q2004	(2)	2N3906 transistor
R2001, 2002		
R2003, 2006		
R2008	(4)	1 k, 1/4 watt carbon-film resistor
R2004	(4)	4.7 k, 1/4 watt carbon-film resistor
R2005	(4)	22 k, 1/4 watt carbon-film resistor
R2007	(4)	10 k, 1/4 watt carbon-film resistor
R2009	(4)	100 ohm, 2 watt carbon resistor
Miscellaneous	(14)	heat sink, EG&G Wakefield 201CB

Audio Filter Parts List

Part	Manufacturer	Description
C3001, 3002		
C3003	(1)	0.1 μF, 100 volts DC monolithic capacitor
Q3001	(2)	2N3904 transistor
R3001, 3002	(4)	330 ohm, 1/4 watt carbon-film resistor
R3003, 3004	(4)	10 k, 1/4 watt carbon-film resistor
R3005	(4)	1 k, 1/4 watt carbon-film resistor

Fan Control Parts List

Part	Manufacturer	Description
C4001	(1)	1.0 μF, 25 volts DC tantalum capacitor
C4002, 4003	(1)	0.1 μF, 100 volts DC monolithic capacitor
C4004	(1)	0.001 μF, 100 volts DC disk ceramic capacitor
D4001	(12)	LM335 temperature-sensitive zener diode
D4002	(2)	1N4148 switching diode
FA4001	(13)	3 inch, 5 volts DC fan (Papst 8105G, see text)
Q4001	(2)	IRF-640 power MOSFET
R4001, 4005	(4)	1 k, 1/4 watt carbon-film resistor
R4002	(4)	6.8 k, 1/4 watt carbon-film resistor
R4003	(6)	10 k, 5 turn potentiometer (Bourns 3339P-1-103)
R4004	(4)	10 k, 1/4 watt carbon-film resistor
U4001	(12)	7805 regulator IC
U4002	(15)	UA741 operational amplifier IC

Manufacturers Listing

(1) Mallory Capacitor Co., P.O. Box 1284, Indianapolis, Indiana 46206.
(2) Motorola Inc., Semiconductor Products Sector, 3102 N. 56th Street, Phoenix, Arizona 85018.
(3) Aromat Corp., 629 Central Avenue, New Providence, New Jersey 07974
(4) Allen-Bradley Co. Inc., 1201 S. 2 Street, Milwaukee, Wisconsin 53204
(5) Sprague-Goodman Electronics Inc., 134 Fulton Avenue, Garden City Park, New York 11040
(6) Bourns, 1200 Columbia Avenue, Riverside, California 92507
(7) M/A-COM Inc., 43 South Avenue, Burlington, Massachusetts 01803
(8) Littlefuse-Tracor, 800 E. Northwest Highway, Des Plaines, Illinois 60016
(9) Magnecraft Electric Co., 1910 Techny Road, Northbrook, Illinois 60062-5376
(10) Amidon Associates, P.O. Box 956, Torrance, California 90508.
(11) International Rectifier, 348 Kansas Street, El Segundo, California 90245.
(12) National Semiconductor, 333 Western Avenue South, Portland, Maine 04016.
(13) Pabst/Pamotor.
(14) EG&G Wakefield, 60 Audubon Road, Wakefield, Massachusetts 01880
(15) Fairchild Semiconductors; see National Semiconductor.

■

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SPECIAL COLUMN

QUARTERLY DEVICES

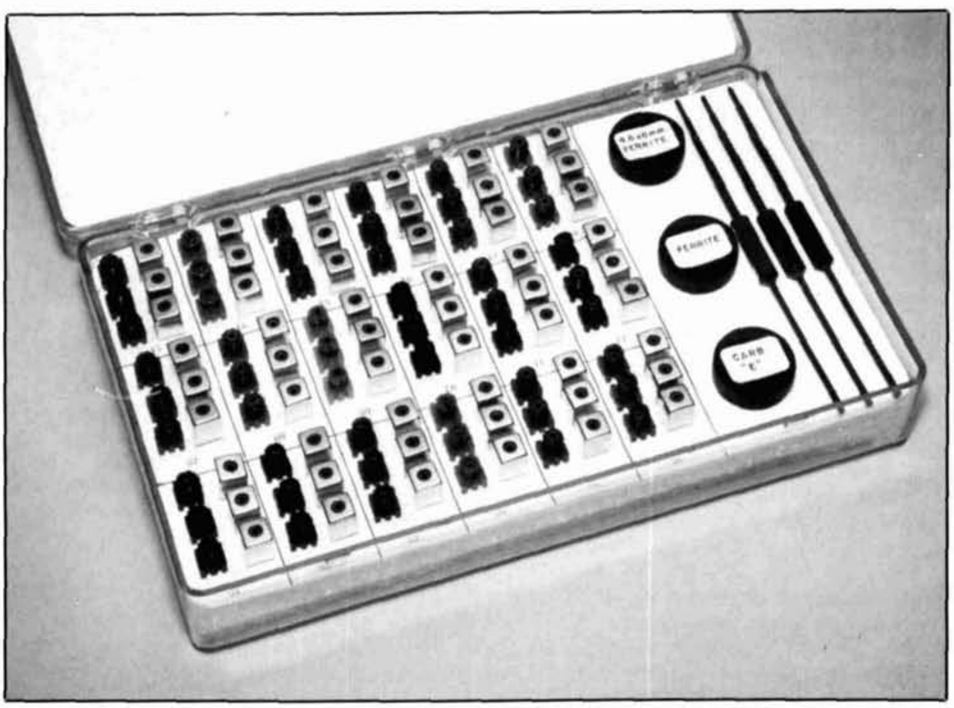
To the vector go the coils

RF network design has never been simpler, thanks to the great selection of user-friendly CAD programs around these days. CAD eliminates hours of pencil biting for the consummate engineer, and opens up new opportunities for those of us who previously avoided mathematical rigor in favor of the cut-and-try approach. Many inexpensive and easy-to-use programs are available. Some were described in detail in the Spring 1992 issue of *Communications Quarterly* by F. Dale Williams, K3PUR, in his article, "Be A CAD (User)."

Since "Quarterly Devices" is a hardware column, I won't stray into the realm of describing design software. Rather, I'll present some especially useful devices that help to make CAD-generated solutions come to life on the bench.

In the past, finding good HF shielded inductors for commercial prototypes or "cosmetically correct" home projects wasn't easy. Domestic coils were expensive and too large for diminutive solid-state packaging. Low-cost miniatures were manufactured overseas and were difficult to get. It's not surprising that hand-wound toroids emerged as the home-builder's inductor of choice during the 70s and 80s.

Fortunately, access to small and inexpensive HF and VHF slug-tuned inductors is improving. Companies such as TOKO now have widespread distribution in the United States, and some domestic manufacturers offer excellent tunables molded from low-



(Photo courtesy of Coilcraft.)

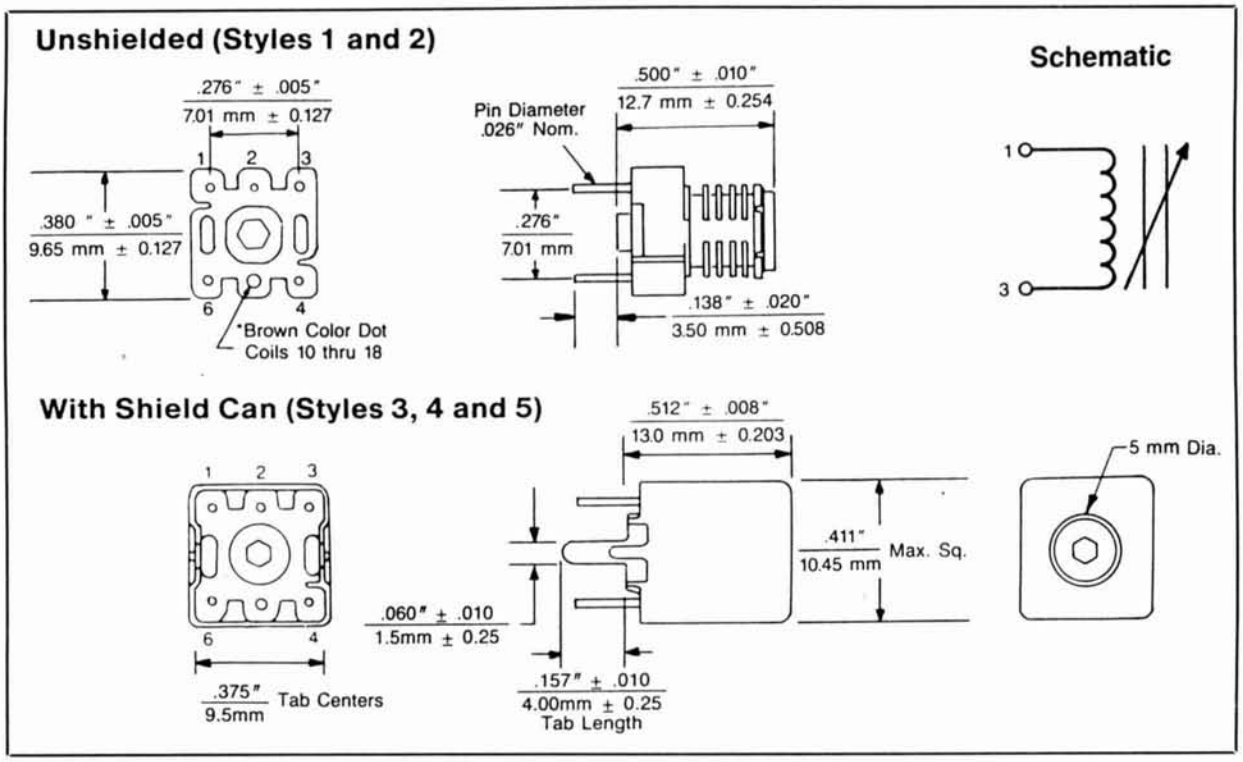


Figure 1. Slotted bobbin enhances temperature stability and supports electromagnetic shielding ring. Standard- sized 10-mm can provides electrostatic shield. (Courtesy of Coilcraft.)

cost thermoplastic materials. At least one American company—Coilcraft Corporation of Cary, Illinois—offers a line of versatile designer kits to professional and amateur builders for very reasonable prices.

"Slot Ten" 10-mm tunable RF coils

For building small-signal CAD-generated filter designs in the 500-kHz to 30-MHz region, Coilcraft's Slot Ten tunable RF inductors are hard to beat. These standard-sized 10-mm devices come in 18 values ranging from less than 1 μH to over 1 mH. Each coil may be outfitted with a variety of shield and core options to suit nearly any application (see **Photo A**).

The Coilcraft Slot Ten M100 designer's kit contains everything you need to make any value and style of coil offered in the Slot Ten line. Six samples of each coil (there are 18) are provided for a total of 108 pieces. Half of these samples are outfitted with shield cans and half are not. Because Coilcraft shield cans aren't crimped to the coil base, you can remove and reinstall them on other coils as needed. In addition to 108 coils and 54 shield cans, you also get a quantity of alternative tuning cores and sleeves which you can use to customize your coils to fit specific applications.

The Slot Ten series offers designers a lot of versatility, and this means making choices. For example, you'll elect whether or not to use an electrostatic shield can, and whether or not to install an electromagnetic

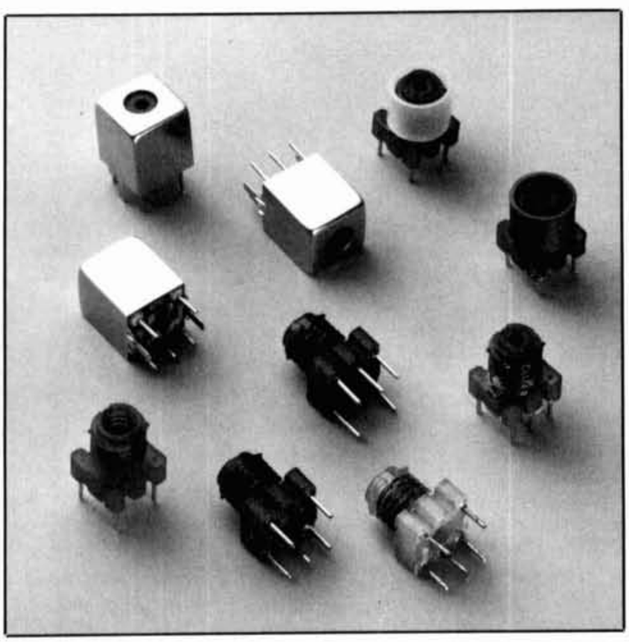


Photo A. Designers may outfit Slot Ten inductors with various shielding and slug options to fit specific applications. (Courtesy of Coilcraft.)

Style 1 — Unshielded with carbonyl E core						
Part Number	Color	L min (uH)	L max (uH)	L nom (uH)	Min Q @ L nom	Test Frequency
SLOT TEN-1-01	Brown	.800	1.2	1.0	48	7.9 MHz
SLOT TEN-1-02	Red	1.2	1.8	1.5	50	7.9 MHz
SLOT TEN-1-03	Orange	1.76	2.64	2.2	56	7.9 MHz
SLOT TEN-1-04	Yellow	2.56	3.84	3.2	54	7.9 MHz
SLOT TEN-1-05	Green	3.6	5.4	4.5	54	7.9 MHz
SLOT TEN-1-06	Blue	5.2	7.8	6.5	51	7.9 MHz
SLOT TEN-1-07	Violet	7.6	11.4	9.5	48	7.9 MHz
SLOT TEN-1-08	Grey	11.2	16.8	14.0	38	2.5 MHz
SLOT TEN-1-09	White	16.0	24.0	20.0	38	2.5 MHz
SLOT TEN-1-10	Black*	22.4	33.6	28.0	38	2.5 MHz
SLOT TEN-1-11	Brown*	32.0	48.0	40.0	38	2.5 MHz
SLOT TEN-1-12	Red*	46.4	69.6	58.0	47	2.5 MHz
SLOT TEN-1-13	Orange*	67.2	100.2	84.0	48	2.5 MHz
SLOT TEN-1-14	Yellow*	96	144	120.0	40	790 KHz
SLOT TEN-1-15	Green*	137.6	206.4	172.0	34	790 KHz
SLOT TEN-1-16	Blue*	200	300	250.0	39	790 KHz
SLOT TEN-1-17	Violet*	280	432	360.0	30	790 KHz
SLOT TEN-1-18	Grey*	416	624	520.0	35	790 KHz

Style 2 — Unshielded with ferrite core						
Part Number	Color	L min (uH)	L max (uH)	L nom (uH)	Min Q @ L nom	Test Frequency
SLOT TEN-2-01	Brown	.80	1.7	1.2	47	7.9 MHz
SLOT TEN-2-02	Red	1.25	2.75	2.0	48	7.9 MHz
SLOT TEN-2-03	Orange	1.75	4.0	2.9	48	7.9 MHz
SLOT TEN-2-04	Yellow	2.4	5.4	3.9	47	7.9 MHz
SLOT TEN-2-05	Green	3.5	7.8	5.6	47	7.9 MHz
SLOT TEN-2-06	Blue	4.7	10.6	7.6	46	7.9 MHz
SLOT TEN-2-07	Violet	7.4	15.6	11.5	38	2.5 MHz
SLOT TEN-2-08	Grey	11	25	18.2	40	2.5 MHz
SLOT TEN-2-09	White	16	35	25.3	40	2.5 MHz
SLOT TEN-2-10	Black	22	50	36.5	40	2.5 MHz
SLOT TEN-2-11	Brown*	33	72	52.5	39	2.5 MHz
SLOT TEN-2-12	Red*	46	103	74.8	51	2.5 MHz
SLOT TEN-2-13	Orange*	66	136	100	40	790 KHz
SLOT TEN-2-14	Yellow*	95	198	146	44	790 KHz
SLOT TEN-2-15	Green*	136	297	216	40	790 KHz
SLOT TEN-2-16	Blue*	198	426	312	45	790 KHz
SLOT TEN-2-17	Violet*	286	630	530	33	790 KHz
SLOT TEN-2-18	Grey*	418	927	790	38	790 KHz

shield sleeve. And, you must choose the most appropriate tuning-core material for your application. To help you make these decisions wisely, let's take a closer look at the basic product and the options Coilcraft provides to enhance it.

The basic Slot Ten coil

The basic Slot Ten molded form has a four-section slotted bobbin. Turns are scramble-wound into each section, providing superior stability over a single-layer solenoid winding. The bobbin ridges between slots equally divide and segment the winding—and serve as supports for the optional electromagnetic shielding sleeve.

The coil base has a standard 5-pin footprint. This provides ample support and prevents insertion errors during pc board assembly operations. Two pins connect to the coil's untapped primary; there is no secondary winding. Use 3.5-mm pin spacing when laying out your pc art (see **Figure 1** for an illustration).

The thermoplastic in each form is pigmented with RETMA color-code for easy coil identification. There are 18 coils in the SLOT-TEN line; coils #01-#09 are solid colored; coils #10-#18 are solid with a prominent brown dot on the base. To illustrate how this identification system works, a plain red form designates a #02 coil and a red form with a brown dot designates a #12 coil.

Coils assigned higher numbers have progressively more inductance. Actual inductance for each coil varies significantly with different combinations of sleeve, core, and shielding installed. Because of this, no value in microhenries is assigned to basic coils. Inductance for any specific combination of coil plus add-ons is found in an inductance chart like the one shown in **Table 1**.

Style 3 — Shielded with carbonyl E core and plastic sleeve

Part Number	Color	L min (uH)	L max (uH)	L nom (uH)	Min Q @ L nom	Test Frequency
SLOT TEN-3-01	Brown	.70	.94	.82	65	7.9 MHz
SLOT TEN-3-02	Red	1.05	1.41	1.23	37	7.9 MHz
SLOT TEN-3-03	Orange	1.5	2.0	1.75	40	7.9 MHz
SLOT TEN-3-04	Yellow	2.04	2.76	2.4	40	7.9 MHz
SLOT TEN-3-05	Green	2.9	3.9	3.4	39	7.9 MHz
SLOT TEN-3-06	Blue	4.25	5.75	5.0	38	7.9 MHz
SLOT TEN-3-07	Violet	6	8.2	7.1	35	7.9 MHz
SLOT TEN-3-08	Grey	9.4	12.6	11.0	25	2.5 MHz
SLOT TEN-3-09	White	12.8	17.2	15.0	25	2.5 MHz
SLOT TEN-3-10	Black*	18.7	25.3	22.0	26	2.5 MHz
SLOT TEN-3-11	Brown*	26.4	35.6	31.0	25	2.5 MHz
SLOT TEN-3-12	Red*	37.4	50.6	44	29	2.5 MHz
SLOT TEN-3-13	Orange*	52.7	71.3	62	30	2.5 MHz
SLOT TEN-3-14	Yellow*	79	105	92	28	2.5 MHz
SLOT TEN-3-15	Green*	108.8	147.2	128	18	790 KHz
SLOT TEN-3-16	Blue*	155	208	182	20	790 KHz
SLOT TEN-3-17	Violet*	230	310	270	16	790 KHz
SLOT TEN-3-18	Grey*	336	450	390	18	790 KHz

Style 4 — Shielded with carbonyl E core and sleeve

Part Number	Color	L min (uH)	L max (uH)	L nom (uH)	Min Q @ L nom	Test Frequency
SLOT TEN-4-01	Brown	.80	1.2	1.0	44	7.9 MHz
SLOT TEN-4-02	Red	1.2	1.8	1.5	45	7.9 MHz
SLOT TEN-4-03	Orange	1.76	2.64	2.2	48	7.9 MHz
SLOT TEN-4-04	Yellow	2.56	3.84	3.2	48	7.9 MHz
SLOT TEN-4-05	Green	3.6	5.4	4.5	46	7.9 MHz
SLOT TEN-4-06	Blue	5.2	7.8	6.5	43	7.9 MHz
SLOT TEN-4-07	Violet	7.6	11.4	9.5	40	7.9 MHz
SLOT TEN-4-08	Grey	11.2	16.8	14	32	2.5 MHz
SLOT TEN-4-09	White	16.0	24.0	20	33	2.5 MHz
SLOT TEN-4-10	Black*	22.4	33.6	28	32	2.5 MHz
SLOT TEN-4-11	Brown*	32.0	48.0	40	32	2.5 MHz
SLOT TEN-4-12	Red*	46.4	69.6	58	40	2.5 MHz
SLOT TEN-4-13	Orange*	67.2	100.8	85	40	2.5 MHz
SLOT TEN-4-14	Yellow*	96	144	120	33	790 KHz
SLOT TEN-4-15	Green*	137.6	206.4	172	28	790 KHz
SLOT TEN-4-16	Blue*	200.0	300.0	250	32	790 KHz
SLOT TEN-4-17	Violet*	288	432	360	26	790 KHz
SLOT TEN-4-18	Grey*	416	624	520	29	790 KHz

Style 5 — Shielded with ferrite core and sleeve

Part Number	Color	L min (uH)	L max (uH)	L nom (uH)	Min Q @ L nom	Test Frequency
SLOT TEN-5-01	Brown	.86	2.0	1.4	43	7.9 MHz
SLOT TEN-5-02	Red	1.27	3.25	2.25	44	7.9 MHz
SLOT TEN-5-03	Orange	1.83	4.6	3.25	41	7.9 MHz
SLOT TEN-5-04	Yellow	2.48	6.4	4.5	40	7.9 MHz
SLOT TEN-5-05	Green	3.58	9.3	6.5	40	7.9 MHz
SLOT TEN-5-06	Blue	5.0	12.7	8.8	38	7.9 MHz
SLOT TEN-5-07	Violet	7.7	18.4	13.0	30	2.5 MHz
SLOT TEN-5-08	Grey	12.0	30.0	20.9	35	2.5 MHz
SLOT TEN-5-09	White	16	42	29.4	34	2.5 MHz
SLOT TEN-5-10	Black*	25	60	42.3	32	2.5 MHz
SLOT TEN-5-11	Brown*	36	90	62.8	30	2.5 MHz
SLOT TEN-5-12	Red*	48	126	87.2	42	2.5 MHz
SLOT TEN-5-13	Orange*	72	163	116	35	790 KHz
SLOT TEN-5-14	Yellow*	102	238	168	36	790 KHz
SLOT TEN-5-15	Green*	147	360	252	36	790 KHz
SLOT TEN-5-16	Blue*	215	522	312	45	790 KHz
SLOT TEN-5-17	Violet*	303	765	530	33	790 KHz
SLOT TEN-5-18	Grey*	440	1143	790	38	790 KHz

*Coils 10 thru 18 have a brown color dot on bottom of coil base.

Table 1. Inductance and Q for each Slot Ten coil will vary, depending on the combination of shielding and core material installed. Values for the five most popular combinations or "styles" are shown. (Courtesy of Coilcraft.)

The electrostatic shielding option

Slot Ten coils may be used with or without a 10-mm electrostatic shielding can, depending on your application. The primary function of an electrostatic shield is to reduce stray coupling between the coil and other radiating elements in your equipment package. In typical applications, a shield can will eliminate unwanted RF pick up in small-signal stages, reduce local oscillator leakage, prevent VFO instability, and increase port isolation in high-gain amplifiers. The down side of electrostatic shielding is Q loss caused by the proximity of the metal can. However, in a "busy" RF environment like a transceiver, concerns over Q reduction are often overshadowed by the need for circuit isolation.

The electromagnetic shielding option

Electromagnetic (EMI) shielding is provided by a tubular sleeve that slips over the coil bobbin. This sleeve is composed of either ferrite or Carbonyl L—a type of powdered iron. The type of tuning slug you select will govern which type of sleeve you install (the M100 kit provides both types). To add an EMI shield to your coil, simply remove the metal can and swap the clear-plastic bobbin sleeve for the electromagnetic sleeve. Then, reinstall the can.

The primary function of EMI shielding is to protect coils from interference induced by stray magnetic fields. These fields are typically generated by adjacent components like transformers, flybacks, loudspeakers, motors, and relay-coils. In transceivers, a strong DC magnetic field or EMI "hit" can easily detune VCO or VFO inductors, causing an unexplained shift or drift in frequency. A strong AC field can also permeate coils and modulate RF signals in the signal path.

Unfortunately, adding an electromagnetic shield can reduce stability in frequency-determining elements. If you install EMI shielding on a VFO coil, use carbonyl E to minimize this effect. However, the good news is that EMI shielding adds a mass of core material around the coil, increasing both inductance and Q. This usually helps to offset the damping effect of the shield can—a factor worth considering when designing electrostatically shielded band-pass filters.

Slot Ten tuning slugs

A choice of ferrite or carbonyl E (powdered iron) slugs is provided with Slot Ten coils. The one you select will depend upon your specific application. Here are some general guidelines to help you choose:

Ferrite: The Slot Ten ferrite slug has more influence over your coil's inductance (Lmin–Lmax) than the carbonyl slug. The ferrite slug also enhances Q in high-inductance coils more effectively than the carbonyl slug. Because of these factors, ferrite slugs are more often paired with the higher inductance coils in the kit. The exception may be for frequency-sensitive applications where the carbonyl slug is typically more stable.

Carbonyl E: The carbonyl E—or powdered iron—slug has less influence over your coil's inductance range (Lmin - Lmax) than ferrite. In addition, the carbonyl slug enhances Q of low-inductance coils more effectively than the ferrite slug. This makes carbonyl a more attractive choice for the lower-inductance coils in the kit. A carbonyl slug will virtually always prove more stable when Slot Ten coils are used for VFOs and other frequency-sensitive applications.

The "crossover zone"—that gray area where the choice between ferrite and powdered iron becomes less clear—occurs around 10 to 15 μH. Reviewing performance specifications in Coilcraft's "Style" charts (**Table 1**) and looking closely at your application may help you decide. For more concise data, obtain a set of Slot Ten Q-plots from Coilcraft.

Both types of Slot Ten slug are adjusted using a standard non-conductive hex-head tuning wand. Two are provided with the M100 kit.

Specifying Slot Ten coils: five styles, eighteen flavors

If you wish to order a specific Slot-Ten item, here's how to designate exactly what you want. Each Slot Ten part number consists of a style number and a coil number. The five style designators are:

Style 1—unshielded with carbonyl core
Style 2—unshielded with ferrite core
Style 3—shielded with carbonyl core (no EMI sleeve)
Sytle 4—shielded with carbonyl core and carbonyl EMI sleeve
Style 5—shielded with ferrite core and ferrite EMI sleeve

The coil number indicates which coil (#01 through #18) you want.

Part numbers are written as SLOT TEN-(style)-(coil number). To illustrate how this works, to order a #03 coil with a shield can, carbonyl slug, and carbonyl EMI sleeve, you'd specify a SLOT-TEN-4-03. Omit the EMI shield, and it becomes SLOT-TEN-3-03. Omit the can, and you have a SLOT-10-1-03.

Teamwork: CAD and the Slot Ten M100 Designer Kit

I use *RF CAD 3.7* to model RF filters and matching networks. This is a very simple and inexpensive DOS-compatible BASIC program written by Gary Field, WA1GRC with assistance from Joe Reisert, W1JR.

To get started, I simply select the design options I want and input the requisite information (including the nominal inductance of the Slot Ten coil I wish to use). Then I let the computer go to work. My PC cranks out a list of over-specific values like, "63.25 pF for Ca and 151.83 pF for Cx." I install the nearest standard-value capacitors to these numbers, and let the tuning slugs take up the slack.

Like any experimenter, I don't always use the first CAD solution I generate. It's always fun to model alternative approaches and play "let's see what happens when I do this." The point is, pairing the CAD program with a versatile HF design kit like the Slot Ten M100 enables me to generate practical network solutions that work properly the first time. Best of all, I don't have to custom-wind anything—or touch a pencil!

The bottom line

The Slot Ten M100 designer kit is priced at $60; not bad considering you get 108 coils plus a large assortment of cans, shields, and slugs in a handy clear-styrene box. Once you purchase the M100 Kit, Coilcraft typically provides replacement coils at no charge. To order, call 800-322-COIL.

In addition to the Slot Ten line, Coilcraft offers a complete line of higher-inductance tunables suitable for upper HF, VHF, and UHF applications. If you wish to continue your design-kit assortment where the Slot Ten line leaves off, consider purchasing the "Unicoil" M102 Kit. This provides an assortment 196 shielded and unshielded coils in both 10 and 7 mm sizes. The inductance range of the coils in the M102 kit is from 1.5 μH to 0.0435 μH. The M102 also sells for $60.

Coilcraft also offers 5 mm "Unicoil" tunables ranging from 14 to 105 nH. In addition, they market a variety of SMT inductors, molded chokes, wideband transformers, and filters. For a complete up-to-date catalog, write to "Coilcraft", 1102 Silver Lake Road, Cary, IL, 60013. ■

PRODUCT INFORMATION

New RF Detector Oscilloscope Probe Kit

Pomona Electronic's new Model 5815 RF Detector oscilloscope probe kit, which features up to 800-MHz bandwidth, can be used for any scope with 10 megohm input. Each kit contains interchangeable oscilloscope probes and accessories and comes packaged in a reusable plastic case. Replacement parts are available.

For a free copy of Pomona's new 140-page catalog, contact: Customer Service, ITT Pomona Electronics, 1500 East Ninth Street, P.O. Box 2767, Pomona, California 91769. Phone: (714) 469-2900. FAX: (714) 629-3317.

New RF Products Selector Guide Available

Motorola has announced the availability of a revised *RF Products Selector Guide and Cross Reference*. The new guide classifies devices to help users select appropriate products for new designs. The classification categories are: preferred, current, and not for new design.

To obtain a copy of the *RF Products Selector Guide and Cross Reference* , call Motorola Literature Distribution at (800) 441-2447, or write: Motorola Inc., Literature Distribution Center, P.O. Box 20924, Phoenix, Arizona 85063. Ask for SG46/D.

Wilfred N. Caron
2945 Camp Joy Road
Grants Pass, Oregon 97526

A SIMPLE AND ACCURATE ADMITTANCE BRIDGE

Build this handy device for frequencies of 2.0 to 30.0 MHz

In articles describing the use of Smith charts and antenna impedance matching, the initial "input" impedance or admittance data is assumed to be accurate. Unfortunately, accuracy is sadly lacking with most, if not all, commercially available amateur radio class RF bridges. I have purchased and evaluated a number of manufactured bridges, and found them to be so grossly inaccurate as to render them virtually useless for their intended application.

During the course of this investigation, I discovered that two conditions contributed to this problem: the unbalanced condition between the primary and secondary windings of the transmission line transformer, and a lack of adequate bridge performance evaluation procedures. The resulting bridge measurements could not be representative of the actual load impedances.

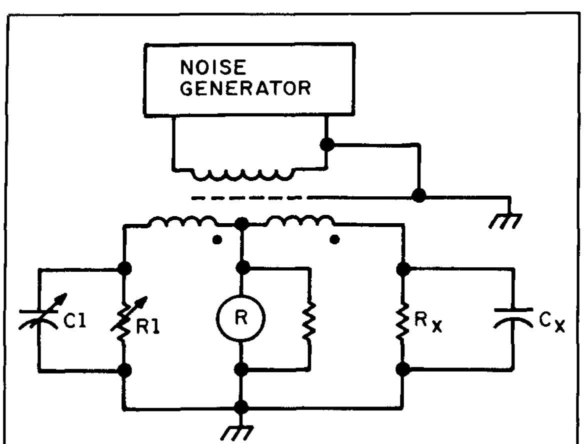


Figure 1. Basic admittance bridge circuit.

In view of the difficulties encountered, my objective became quite clear: to provide a design for a simple and accurate RF bridge; and to provide a procedure for evaluating the accuracy of not only this bridge, but also any other bridge. I selected the admittance rather than the impedance-type bridge because I felt that the accuracy of the bridge would be enhanced by having all the components in parallel and at ground potential. An admittance bridge generates data in the form of parallel combinations of R-C or R-L. The minor inconvenience required to convert parallel combinations to series combinations is offset by the degree of accuracy obtained.

Admittance bridge description

The basic admittance bridge circuit shown in **Figure 1** is simple in the extreme. It operates from a wideband noise generator and contains a specially designed transformer. The two arms of the bridge are driven equally by the transformer. As noted in the figure, there is an electrostatic shield interposed between the primary and secondary windings. This shield effectively suppresses the troublesome capacitive coupling between windings, but permits inductive coupling. When the bridge is balanced, when $R_1 = R_x$ and $C_1 = C_x$, a deep null is detected by the receiver connected at the secondary winding center tap. The center tap is also terminated into a 51-ohm resistor. The resistor prevents the impedance at the receiver terminals from becoming too high. If a mismatch condition exists at this point it is presented to both the reference and unknown ports that would lead to bridge error.

Figure 2 shows the practical circuit of the admittance bridge. Capacitor C1 is switched from one arm of the bridge to the other, depending upon the load reactance. If the load is inductively reactive, the capacitor is placed across the load to neutralize the reactance. For capacitively reactive loads, the capacitor is placed in the opposite arm. A center-off DPDT switch permits the removal of the capacitor from the bridge circuit when purely resistive loads are encountered.

Components and Component Selection

Transformer T1. The accuracy of the admittance bridge is obtained by careful component placement and the design of transformer T1. The secondary winding is made by twisting together two pieces of no. 24 AWG enameled wire, four twists per inch. Three turns of the twisted wires are wound on an Amidon* ferrite balun core, part no. BLN 43-202, as shown in **Figure 3**. The primary winding—a length of RG-174/U miniature coaxial cable—is next wound on the ferrite core. Only one turn is required.

The through holes of the ferrite core have an inside diameter of about 0.150-inch in diameter. It is easier to thread the cable through these holes if you coat it with silicon lubricant. RG-174/U may be difficult to obtain in short lengths. You may substitute a miniature microphone shielded cable with an outside diameter of 0.105 inch (Radio Shack no. 278-510) without degrading the performance of the bridge.

When fabricating the balun it's important that the mechanical symmetry be maintained for the windings. Symmetry is re-

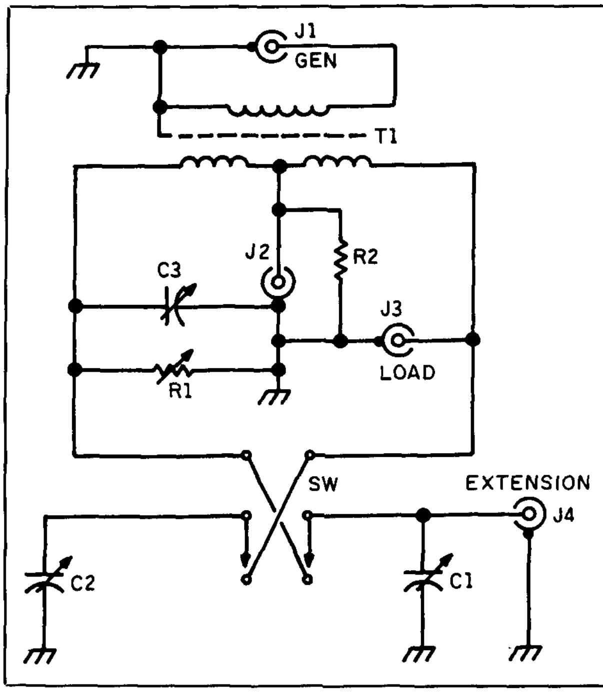


Figure 2. Practical admittance bridge circuit.

quired to achieve an electrical balance to ground over the entire operating frequency range. Lack of symmetry will cause the null point to shift with frequency; this will lead to inaccuracies in bridge measurements.

Standard capacitor. The variable air capacitor used as the standard capacitor is a

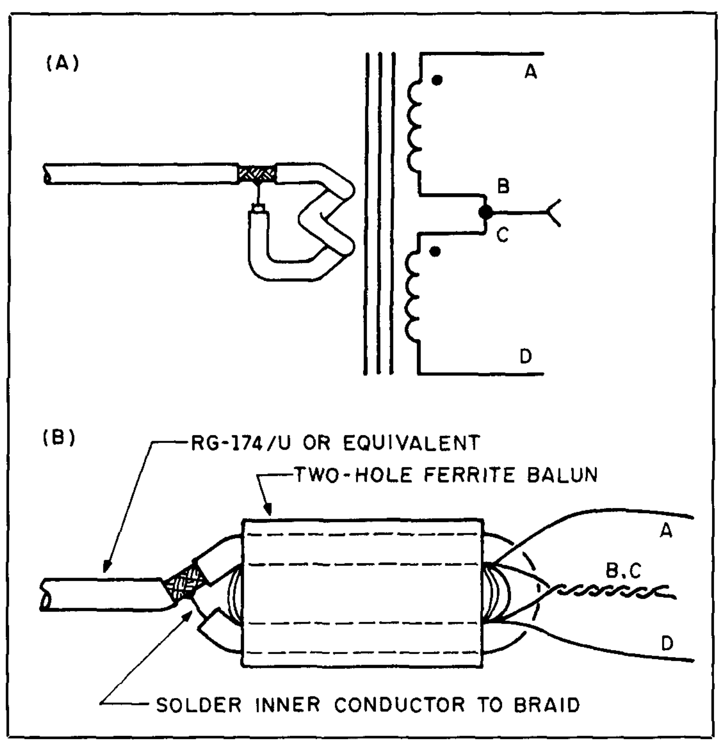


Figure 3. Details of admittance bridge transformer. (a) Circuit diagram. (b) Method of winding transformer.

*Amidon Associates, P.O. Box 956, Torrance, California 90508.

Photo A. Front panel assembly of completed bridge.

single section 365-pF capacitor. Variable capacitors of this size are becoming difficult to find. Antique Electronics Supply** provides a wide selection of single and multiple section variable air capacitors. Ask for part number CV-230.

Standard potentiometer. I used a (Clarostat) Allen-Bradley type "J" as the standard potentiometer. Its nomenclature is RV4NAYSD251A. Its resistance is 250 ohms.

DPDT toggle switch. For the toggle switch used to switch the capacitor from one side of the bridge to the other, I chose a submii.i DPDT center-off switch (Radio Shack no. 274-620).

Bias set capacitor. All variable capacitors have a minimum or residual capacitance. The standard capacitor has a minimum capacitance (in the circuit) of about 15 pF. This minimum capacitance must be tuned out. You can neutralize the minimum capacitance by placing a small trimmer capacitor (C2) in the bridge circuit that is in the opposite arm to the standard capacitor and switching it with the DPDT switch. C2 is a 6 to 50 pF trimmer capacitor (Radio Shack no. 272-1340).

Other bridge circuit components. Even the best of bridges require some form of compensation to neutralize unbalance caused by bridge components and wiring. A small micro-miniature trimmer capacitor of 3.5 to 20 pF across the terminals of the potentiometer was necessary with the prototype bridge. Very little compensation was required. If the bridge doesn't neutralize with the trimmer in this location, the alternative is to place it across the terminals of the load connector.

Standard capacitor extension. The maximum capacitance of the standard capacitor may not be adequate to measure low reactance values at low frequencies. To extend the range of C1, a number of plug-in fixed mica capacitors between the range from about 300 to 1500 pF or greater, in multiples of about 300 pF (300, 600, 900, and so on) are attached to phono plugs (Radio Shack no. 274-339). Each assembly should be tagged with its capacitance value. The value required to achieve the desired bridge null is simply added in parallel to the setting of the standard capacitor to obtain the total capacitance.

**Antique Electronics Supply, 6221 S. Maple Avenue, Tempe, Arizona 85283.

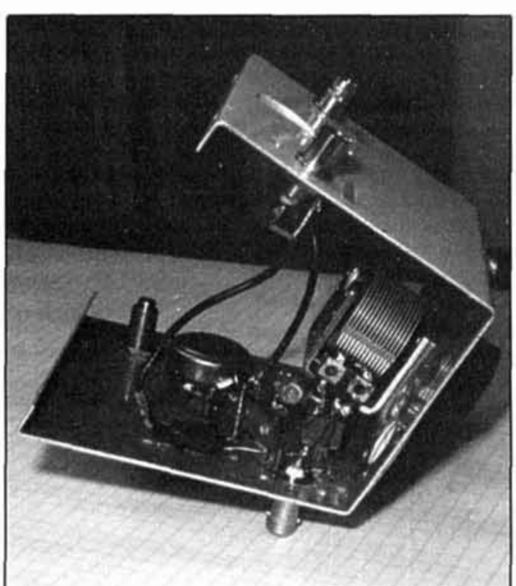


Photo B. Arrangement of components of complete bridge.

Bridge assembly

The completed bridge is shown in **Photos A** and **B**. Note the compact arrangement of the front-panel components. A smaller resistance dial on the front of the cabinet would have shortened the lead to the potentiometer by about one-half inch. However, the present layout works quite well. The dial for the standard capacitor is on top of the cabinet and the placement of the capacitor is such that one of its lugs is soldered directly to the −X +X DPDT switch. The phono jack to the right of the switch accommodates the fixed capacitor pads. The jack is connected directly to another standard capacitor lug, thus eliminating another undesirable length of wire. The capacitor dial can be calibrated in pF and the resistance dial in ohms. However for increased accuracy, I prefer to use calibrated charts (see **Figures 4** and **5**).

Resistance-reactance measurement range

The resistance range of the admittance bridge is determined by the size of the potentiometer, 0 to 250 ohms. The reactance range is determined by the total capacitance and the lowest frequency as indicated by the equation:

$$-X = \frac{1}{2\pi fC} \tag{1}$$

where: $-X$ = capacitive reactance in ohms
f = operating frequency in hertz per second
C = capacitance in farads

The provision to switch the standard capacitor from one side of the bridge to the other reverses the sign of the reactance from $-X$ to $+X$; the $+X$ indicates that the load is inductively reactive.

Calibration procedure

Step 1. Calibrate the resistance dial. An accurate ohmmeter is all that's necessary. This must be done before the potentiometer is wired into the bridge circuit. Generate a calibration chart as shown in **Figure 4**.

Step 2. Calibrate the standard capacitor. With the $-X$ $+X$ DPDT switch set to its center-off position, calibrate the capacitor with a capacitance meter or capacitance bridge. Log the capacitance versus dial reading values.

Step 3. Connect the bridge to a noise source and receiver. Tune the receiver to 30.0 MHz. Connect a precision nonreactive 50-ohm load to the load input connector. Set the toggle switch to its center-off position. Adjust the standard potentiometer to obtain a null. The null may not be very deep. Adjust trimmer capacitor C3 for maximum null depth.

Step 4. Set the toggle switch to either $-X$ or $+X$ position. Set the standard capacitor to its minimum position. Adjust trimmer capacitor C2 to obtain deepest null. This step neutralizes the residual capacitance of the standard capacitor. The minimum effective capacitance is now 0 pF. For a well-balanced bridge, switching from $-X$ through the center-off to $+X$ shouldn't disturb the null depth.

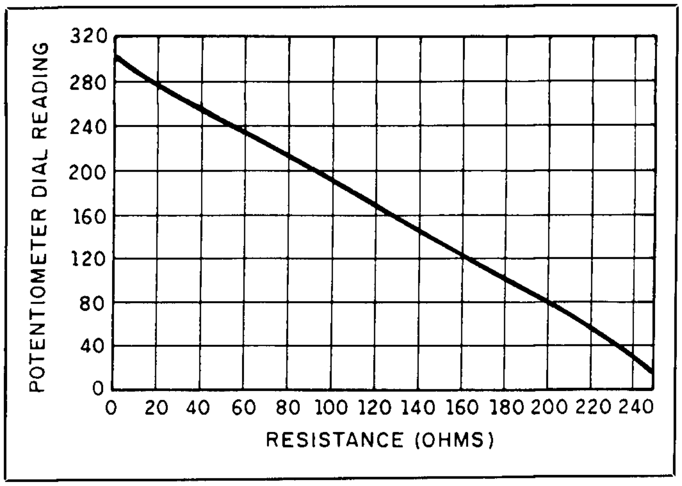


Figure 4. Potentiometer dial calibration curve.

Step 5. Subtract the minimum capacitance of the standard capacitor from all capacitance measurements obtained in **Step 2**. Draw a capacitance versus dial reading calibration curve. This completes the bridge calibration procedure.

Bridge accuracy evaluation procedure

To evaluate the accuracy of a bridge, some standard must be established that will provide a set of calculated accurate impedance points for a specific frequency range. These impedance points are then compared with the measured impedance of the bridge undergoing evaluation.

You know from transmission line theory that when a transmission line with a characteristic impedance of Z_0 is terminated into a load resistance Z_r that is equal to the

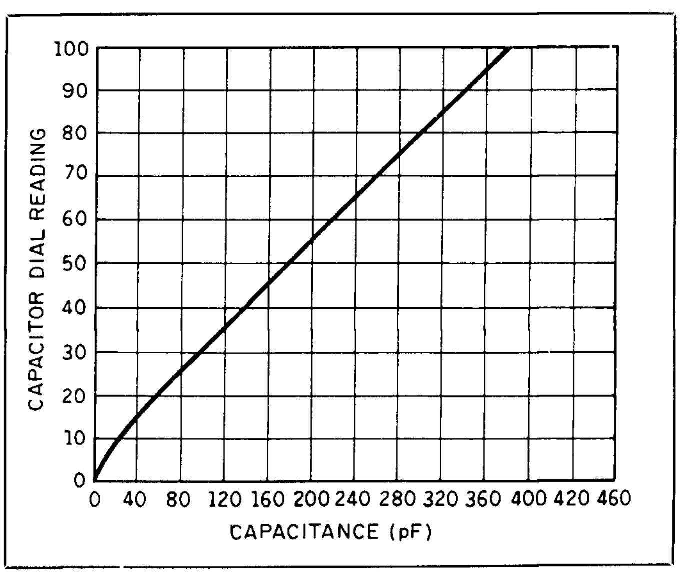


Figure 5. Capacitor dial calibration curve.

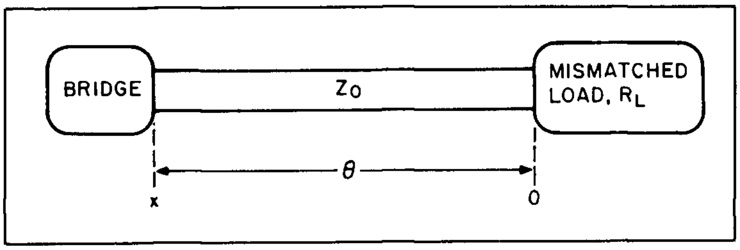


Figure 6. Transmission-line representation for admittance or impedance measurements. The dashed line at 0 is the point at which the load is connected. The reference point x, where the input impedance is determined, is at a distance θ, measured in electrical degrees, from the 0 reference line. For this measurement, load R_L cannot be equal to Z_0; it must be a mismatched load.

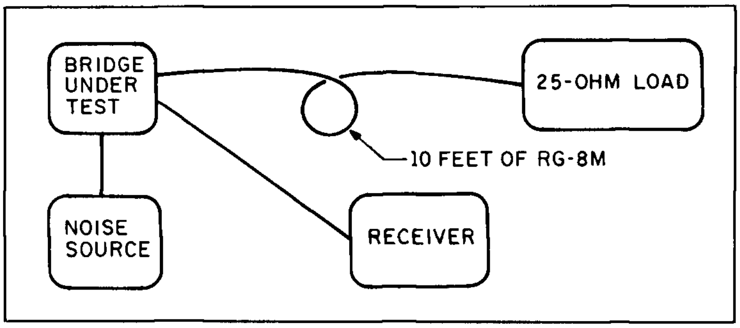


Figure 7. Test setup used to evaluate admittance bridge accuracy.

impedance of the line ($Z_o = Z_r$), the input impedance (Z_{in}) of the line will be equal to Z_o (Z_{in}) = $Z_o = Z_r$. A general sketch of a transmission line is shown in **Figure 6**. The input impedance Z_{in} at some distance θ in electrical degrees from the load Z_r is given by:

$$Z_{in} = Z_o \left[\frac{Z_r \cos\theta + jZ_o \sin\theta}{Z_o \cos\theta + jZ_r \sin\theta} \right] \quad (2)$$

If you divide all of the terms of **Equation 2** by Z_0, you arrive at the so-called normalized form. In this form, the results can be obtained independent of the characteristic impedance of the transmission line itself.

Before you can proceed with the bridge accuracy evaluation procedure, you must have a length of coaxial cable and a 25-ohm precision termination. I evaluated the performance of the prototype bridge using a 10-foot length of RG-8/M coaxial cable (Radio Shack no. 278-979). The first step in the procedure is to determine the magnitude of Z_{in}. A normalized Smith chart is used for this, therefore, Z_{in} is normalized by dividing Z_{in} by Z_0.

Step 1. Determine the magnitude of Z_o. Given: characteristic impedance of RG-8/M is 52 ohms. At resonance, when the cable length is equal to 90 degrees, cos θ = 90° and sin θ = 90°.

$$\frac{Z_{in}}{52} = \left[\frac{25 \cos 90° + j52 \sin 90°}{52 \cos 90° + j25 \sin 90°} \right]$$

$$= \left[\frac{25(0) + j52(1)}{52(0) + j25(1)} \right] = \frac{j52}{j25}$$

Rationalizing:

$$\frac{Z_{in}}{52} = \left[\frac{j52}{j25} \right] \left[\frac{-j25}{-j25} \right] = \frac{108.16}{52} = 2.08$$

Step 2. Draw a circle of magnitude 2.08 on the normalized Smith chart. Next find the resonant frequency of the 10-foot cable.

Step 3. Using the test setup shown in **Figure 7**, but omitting the 25-ohm termination, find the resonant frequency as follows:

a. Set the resistance dial to zero ohms.
b. Set the toggle switch to its center-off position.
c. Tune the receiver until you find a null.

fx	θfx	cosθ°	sin θ°	Calculated R ± jX	Measured R ± jX
2.0	9.825	0.98533	0.17064	0.49 + j0.13	0.49 + j0.13
5.0	24.563	0.81855	0.41569	0.57 + j0.37	0.59 + j0.34
8.0	39.301	0.77383	0.63334	0.70 + j0.54	0.73 + j0.54
11.0	54.039	0.58723	0.80942	0.97 + j0.74	1.01 + j0.72
14.0	68.777	0.36200	0.93218	1.45 + j0.78	1.44 + j0.74
17.0	83.515	0.11294	0.99360	1.99 + j0.36	1.91 + j0.38
18.320	90.000	0	1	2.08 + j0	2.00 + j0
21.0	103.166	−0.22777	0.97371	1.68 − j0.62	1.66 − j0.60
24.0	117.904	−0.46799	0.88373	1.20 − j0.80	1.15 − j0.75
27.0	132.642	−0.67742	0.73560	0.82 − j0.65	0.82 − j0.60
30.0	147.380	−0.84226	0.53906	0.62 − j0.45	0.63 − j0.41

Table 1. Calculated results for the prototype bridge.

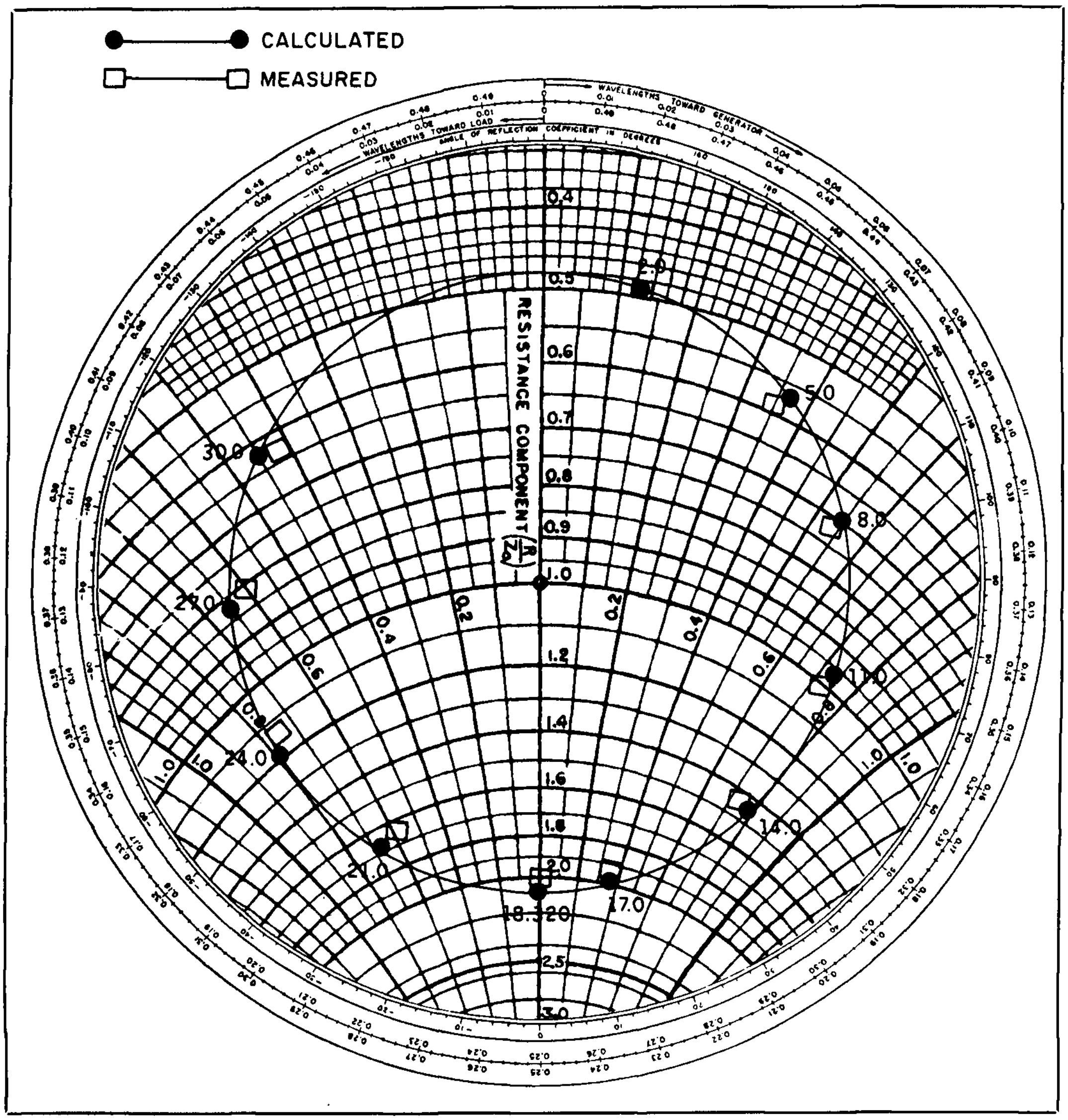


Figure 8. Normalized Smith chart representing measured and calculated impedance points.

d. Readjust the resistance dial and receiver frequency for deepest null. (With the prototype bridge, the null occurred at 18.320 MHz. The cable is 90 degrees long at 18.320 MHz.)

e. The cable length in electrical degrees at other frequencies θfx is determined as follows:

$$\theta fx = \frac{fx}{18.320} \times 90^{\circ}$$

where fx = other frequencies.

Step 4. Prepare a table showing θfx at the frequencies of interest starting at 2.0 MHz and ending at 30.0 MHz. (The calculated results for the prototype bridge are tabulated in **Table 1**.)

Step 5. Compute for Z_{in} using **Equation 2** at each frequency of interest. For 2.0 Mhz with 10 feet of RG-8/M and a 25-ohm termination:

$$Z_{in} = \left[\frac{25 \cos 9.825^{\circ} + j52 \sin 9.825^{\circ}}{52 \cos 9.825^{\circ} + j25 \sin 9.825^{\circ}}\right]$$

$$Z_{in} = 52\left[\frac{25(0.98533) = j52(0.17064)}{52(0.98533) + j25(0.17064)}\right] = 52\left[\frac{24.63325 + j8.87328}{51.23716 + j4.26600}\right]$$

Rationalizing:

$$Z_{in} = 52\left[\frac{24.63325 + j8.87328}{51.23716 + j4.26600}\right]\left[\frac{51.23716 - j4.22600}{51.23716 - j4.26600}\right]$$

$$= 52\left[\frac{1262.138 + j454.642 - j105.085 + 37.853}{2625.247 + j218.578 - j218.578 + 18.199}\right]$$

$$= 52\left[\frac{1300 + j349.557}{2643.446}\right] = 52\,[\,0.49 + j0.13\,]$$

For normalized Smith chart, enter only R = 0.49 and +jX = 0.13.

Step 6. Repeat for all frequencies and enter data on test data sheet and/or Smith chart.

The data obtained in the preceding steps are admittance type data. That is the R ± jX are paralleled combinations. If you want series combinations, follow Step 7.

Step 7. Measure the load admittance using the test setup shown in **Figure 7**. Convert parallel R ± jX data into series R ± jx data as follows:

$$Rs = \frac{R_p \times X_p^2}{R_p^2 + X_p^2} \quad \text{and} \quad Xs = \frac{R_p^2 \times X_p}{R_p^2 + X_p^2}$$

$$z_{in} = RS \pm jXs$$

where Rp = parallel resistance in ohms
Xp = parallel reactance in ohms
Rs = series resistance in ohms
Xs = series reactance in ohms

Normalize by dividing Rs and Xs by 52, the characteristic impedance of the RG-8/M transmission line. Enter the normalized impedance data on the test data sheet or on a Smith chart.

Step 8. Compare measured data against the calculated data to determine the accuracy of the bridge as shown in **Figure 8.**

Summary

This high accuracy admittance bridge is easily constructed and doesn't require specialized components. Calibration of the bridge is an essential step requiring an ohmmeter and capacitance meter or bridge. It's possible to use the calibration curves provided if the bridge is constructed in a manner similar to the prototype bridge. Although this bridge isn't in a class with commercially available laboratory instruments costing several thousands of dollars, it is, by far, more accurate than commercially available radio amateur class RF bridges priced in the neighborhood of $100. You can build this bridge for a fraction of that cost. ■

CORRECTION

Please make this correction

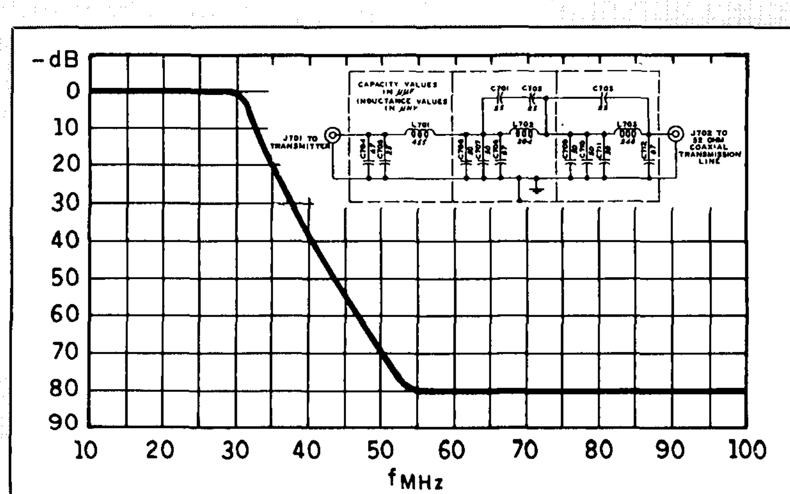
So far, I've received some nice feedback on my article 'Low-pass Filter Performance'' (spring 1992, page 75); however, I do need to make one correction.

In transcribing my raw data on the Collins filter performance, I looked at the top line of my graph format with three lines for 30 MHz, BUT, my scale below required four lines! A sharp-eyed reader caught the ''error of my ways'' and wrote me a nice letter questioning the F_c which, obviously, should have been 30 MHz. Typically, having looked at the data so often for so long, I totally missed it.

Would you be able to publish a correction in the next issue of the quarterly?

Marv Gonsior, W6FR
Fullerton, California

No problem! Readers, here's the revised Figure 7. Ed.



BEST OF THE FOREIGN PRESS

B. Sykes, *G2HCG*
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May 1990

CONTROLLED FEEDER RADIATION

How to improve an antenna's polar diagram by controlling the radiation from its feeder

The use of a balun to feed balanced antennas with coaxial feeder has always been a controversial point, the usual comment being—"it works all right without one, so why should I bother?" The two vital uses of a balun are to ensure that the polar diagram of the antenna is as planned, and to prevent interference pickup on the feeder, or radiation from it.

Interference pickup

The advent of the computer in the shack with its high hash level makes the latter point even more important. Here the difference in hash pickup on the feeder is very noticeable when a coaxial antenna feeder is properly terminated with a balun.

Polar diagram

Control of the polar diagram of the antenna is not, perhaps, so noticeable, but it is very important to know the areas of the world covered by the antenna system. It is, perhaps, even more important to know the areas rejected by nulls in the polar diagram. A dipole erected reasonably in the clear and properly fed with balun and coaxial feeder to the rig will have little or no pickup from the ends, and if oriented in a north-south direction, will provide a useful reduction in QRM from the powerful southern European HF stations. Omit the balun and those nulls will not be in evidence due to uncontrolled radiation and pickup from the feeder. Without a balun, one half of the dipole is connected to the outer of the coaxial feeder, which will radiate in an uncontrolled manner depending on its length.

It may be, however, that you want to have an omnidirectional radiation pattern, or that physical limitations mean the antenna must be erected north-south—although you want to work into Europe to the south. Consider the effect of deliberately controlling the feeder radiation and making use of it. This can easily be achieved by simply moving the balun down the feeder from the antenna feed point by a quarter wave, allowing radiation from the top part of the feeder and using the balun to stop the radiation (and interference pickup) from the lower part of the feeder.

I have called this technique *controlled feeder radiation* (CFR). It should be noted that CFR depends on radiation from the outer shield of a coaxial cable, which is not applicable to balanced feeders.

Balun types

There are many different types of balun available.[1] The simplest and the one applicable here uses the RF choke principle to stop radiation from the outer of the coaxial feeder by simply winding it into a coil or

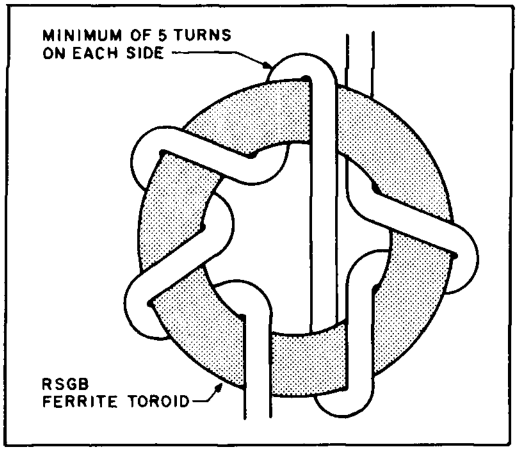


Figure 1. Method of winding choke balun.

onto a ferrite ring, weight and size limitations usually dictating the use of the ferrite. Since high impedance with a minimum number of turns is required, the use of a high-permeability ferrite core is mandatory. Standard black ferrite cores, as used for interference suppression, are ideal, and, since the balun may need to be suspended from the antenna, the use of small coaxial feeder (URM76) is advantageous in the interests of weight reduction. The standard 4 cm O/D core, as supplied by RSGB, will take 11 turns of URM76 and should be wound as shown in **Figure 1** in order to reduce self capacitance. A single core will provide sufficient impedance for 28 to 14 Mhz, and two cores taped together will cover 7 MHz also. If long feeder runs are necessary, the small diameter coax need only be used for the balun and radiating portion, a weatherproof coaxial plug and socket being fitted below the balun to connect to a larger feeder with lower losses. If very high RF voltages are expected, the balun may be wound onto an antenna rod from an old

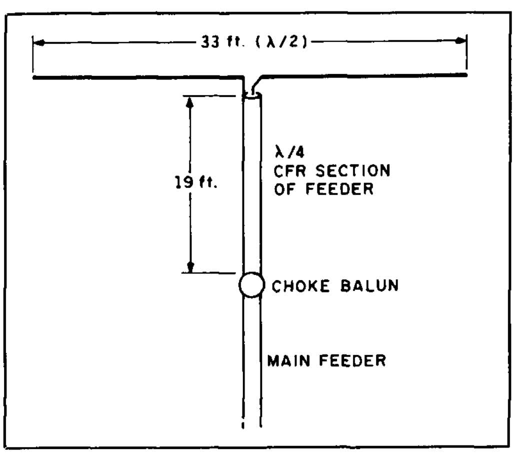


Figure 2. A 14.2-MHz CFR half-wave dipole.

transistor radio, thus physically separating the input and output.

Choice of core

The choke balun operates at high impedance and a relatively low flux in the core, which allows high-permeability materials to be used without fear of core saturation. The transformer balun, usually trifilar wound, operates at low impedance and higher flux densities, often requiring the use of lower permeability ferrites to avoid core saturation and self-resonance effects. The CFR antenna choke core should have a relative permeability of at least 50 at the frequency of use.

Impedance matching and CFR length

A useful advantage of CFR is that the normal 75-ohm impedance of a dipole is reduced to nearer 50 ohms, thus providing a lower VSWR in standard 50-ohm feeder. The physical length of the CFR section will vary with the design of the choke balun. The design shown in **Figure 1** will result in a CFR section length of 0.275 of a wavelength (for example, 19 feet at 14.2 MHz), and this, if added to an existing installation, will not alter the resonant frequency of the system. If a different choke design is used, the CFR section length should be adjusted until the antenna resonant frequency is the same as that of a dipole without the CFR section.

Dipole with CFR

Figure 2 shows a dipole fed with coaxial feeder with the balun placed 0.275 of a wavelength below the feed point, thus providing an omnidirectional vertical quarter-wave radiator in addition to the standard figure-eight pattern of the dipole. The low-angle vertically polarized radiation is a considerable bonus, being achieved without the need for an expensive and complicated system of ground radials. I used this application of CFR this year when operating from southern France back to the UK on 7 MHz, and it has proven very effective indeed. Two days of deliberate operation without CFR, without announcing the fact, resulted in many comments on reduced signal strength.

Multiband CFR

If multiband operation is desired, a trap dipole may be used. A typical example has

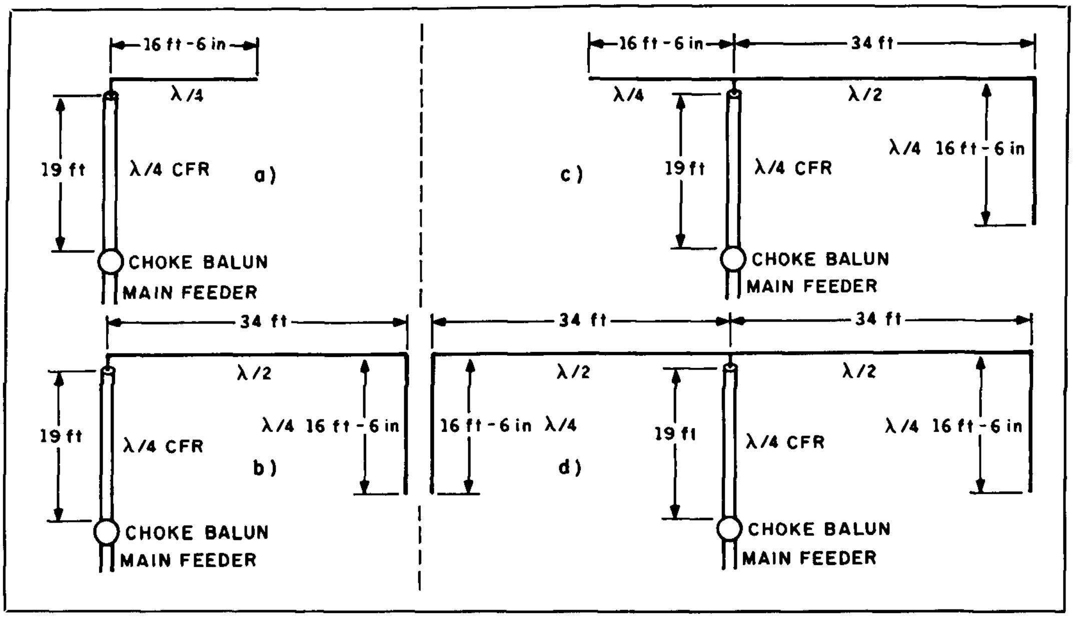


Figure 3. CFR antennas: (A) isotrophic, (B) half square (C) half square, (D) bobtail.

traps for 28, 14, and 7 MHz, and an overall length for 3.5 MHz. The CFR principle can be applied in various ways to this type of antenna, depending on the bands covered.

Placing the choke balun 0.275 wavelength at 28 MHz below the feed point will give optimum all-around DX capability on that band with little or no effect on the other bands, retaining, for example, the QRM-reducing properties of the dipole pattern of 14 and 7 MHz.

Placing the choke balun 0.275 wavelength at 14 MHz below the feed point will give all-around DX coverage on that band with no effect on 3.5 and 7 MHz. The CFR section in this case is a half wave on 28 MHz and, being high impedance, will not accept power from a low-impedance feed.

Similarly, a CFR section having a length of 0.275 wavelength on 7 MHz will give all-around DX coverage on that band with no effect on 14, 28, and 3.5 MHz. The CFR section is high impedance on 14 and 28 MHz.

It is quite feasible to replace the CFR choke balun with a section of feeder wound into a coil and tuned to the required resonant frequency with a capacitor. A number of these resonant traps could be spaced at optimum points along the feeder, thus allowing every possible combination of CFR on the various bands covered.

CFR specials

The concept of controlled feeder radiation means that it is perfectly feasible to have no apparent connection to the outer of the coaxial feeder. The connection exists nevertheless, namely from the inner to the outer surface of the coaxial shield at the antenna end of the feeder. This leads to considerable simplification in the design of a number of antennas and **Figure 3** is perhaps just the beginning of the family of antennas using the CFR principle of making the top section of the feeder into a radiator.

All these antennas, when suitably dimensioned, have been shown to produce a good match to 50-ohm feeder.

The simplest is **Figure 3A**, which consists of a simple quarter-wave, end-fed element combining with the CFR section to produce a right-angled dipole. The antenna will radiate vertically and horizontally polarized signals, or a mixture of both polarizations dependent on the direction from the antenna. With the variable polarizations reflected front the ionosphere, the antenna can be considered to be virtually omnidirectional. Straightening out this antenna results in a very useful low-impedance end-fed dipole, which may be conveniently strung from the window of an upstairs shack to a suitable point in the garden.

Figure 3B could variously be described as a "half square" or a "2/3rd bobtail" and consists of two vertical radiators, fed in phase with equal power. The polar diagram is figure eight at right angles to the wire with a free space theoretical gain of 3 dBd and vertically polarized low-angle radiation. The low angle radiation is particularly useful for DX work, and the gain achieved in practice on a DX signal is considerably more than the theoretical free space 3 dBd.

Figure 3C is a half square with an extra quarter-wave horizontal section, which results in the addition of horizontally polarized radiation to the original vertical radiation. DX signals after reflection from the ionosphere are of varying polarization and the ability to handle all polarizations may well be an advantage.

Figure 3D can only be described as a modified "bobtail." The standard bobtail is end-fed at the high-impedance point at the end of the center radiator. This necessitates a resonant feed system, either a link-coupled tuned circuit or a tapped quarter-wave stub. The CFR system is a low-impedance feed with the center radiator consisting of the CFR section of the feeder. Current distribution in the three verticals is 50 percent in the center and 25 percent in each of the verticals. This is identical to the standard bobtail and results in a free space theoretical gain of 3 dBd with an exceptionally clean figure-eight polar diagram. Comparison tests between this antenna and the two-element version of **Figure 3B** showed identical performance on stations within the beam, and the exceptionally clean pattern of the three-element was a noticeable advantage in reducing the QRM, but conversely disadvantageous if the wanted station was out of the main beam. Basically, the third element was not worth the extra space required, unless the antenna could be accurately oriented onto the wanted station, when the reduction in QRM from the sharp clean polar diagram could be appreciated.

Feeder voltages

It is interesting to consider the voltages on the feeder at the antenna side of the balun. The outside shield of the coaxial cable is behaving as a low-impedance fed quarter-wave radial with an end impedance of some 3000 ohms, which at a power level of 50 watts into that element means some 400 volts at the end furthest from the feed point. The inside of that same shield is at zero potential, being the outer of a coaxial feed line at an impedance of 50 ohms. There is, thus, a potential difference of 400 volts across less than 1 mm of copper at the choke. A dramatic example of skin effect in practice.

Acknowledgement

Initial testing of the CFR principle was conducted at 435 MHz on my "antenna range," but the final full-size tests with much DX operation were carried out by a near neighbor, Bill Wheeler, G3BFC, whose cooperation, patience, and encouragement is gratefully acknowledged. ■

REFERENCES

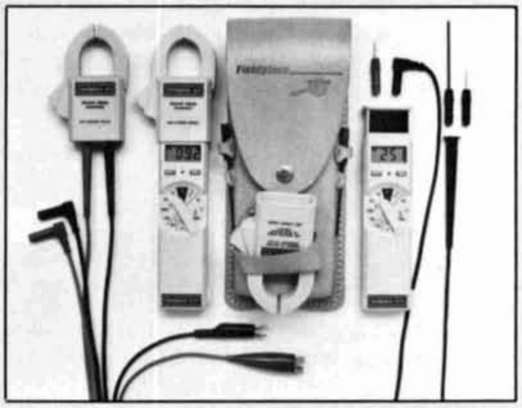
1. Ian White, G3SEK, "Baluns—more than a match," *RadCom*, Radio Society of Great Britain, December 1989.

PRODUCT INFORMATION

New high-resolution "stick" DMM

Fieldpiece Instruments, Inc. has introduced the HS24 heavy-duty DMM for measuring AC current (amps) with resolution to 0.1 A and temperature with resolution to 0.1 degree. The HS24 is also offered as a kit that consists of accessories and a leather case.

The HS24 "stick" style DMM combines the functions of a digital multimeter, a voltage checker, a capacitance meter, a continuity checker, and a current clamp meter in a fully sealed drop-proof, contamination resistant Valox housing. Metal oxide varistors (MOVs) provide transient voltage protection. Jacks on the top of the meter accept probe tips, test leads, alligator clips, or an accessory current clamp head. Also included are a continuity beeper, a high-voltage indicator (blinking red LED and a beeper), and a HOLD button to freeze the reading.



The unit measures AC voltage to 750 volts, DC voltage to 200 volts DC, capacitance to 200 μF, and resistance to 2000 ohms.

For more information, contact Fieldpiece Instruments, Inc., 8322B Artesia Blvd., Buena Park, California 90621.

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SUPERGAIN ANTENNAS

Possibilities and problems

In principle, any desired amount of gain can be developed from an antenna of arbitrary size. The phenomena of high gain from very small antennas is called "supergain."

The optics approach of supergain

To ascertain the validity of supergain, recall the construction used in optics known as Huygens Principle. This principle states that every point on a wavefront can be regarded as a source of radiation, as shown in **Figure 1**. At the end of a short period of time, the envelope of all of these individual wavelets forms the new wavefront. For example, this construct explains why a shadow is not perfectly sharp, and why interference fringes form.

The construct is applicable to all wave propagation—including the signals from antennas. But for this discussion, let's reverse the direction of application of the construct. In **Figure 1**, the right-most wavefront is regarded as existing at some instant. Each point on the front has been created by the sum of the individual point waves of an earlier wavefront. Thus, the left-most wave represents the position of the front at an earlier time. This process can be repeated again and again, until the source of the wave is reached.

There's no restriction on the number of steps taken when tracing the front back toward the source. Therefore, a given pattern can be generated by many different antenna sizes. But this is just a different way of stating my opening sentence.

The end-fire antenna approach to proof of supergain

We can reach the same conclusion using a completely different approach. Consider the case of two parallel ideal half-wave dipoles fed 180 degrees out of phase, at various spacings. These form the basic end-fire array, also known as the flat-top beam or 8JK antenna.

Following Kraus[1] in his book on antennas, the radiation pattern of this antenna can be regarded as the product of two terms—one for the coupling between the antennas, and one for the phase relations of the two sources. The complete expression for the gain in the plane at right angles to the antenna is:

G = SQR(2*R11/(R11 – R12))*SIN(D/2*SIN(ANG)) **(1)**

Where G is the gain
R11 is the radiation resistance of an element
R12 is the mutual resistance between elements
D is the element spacing
ANG is the angle from the line perpendicular to the elements

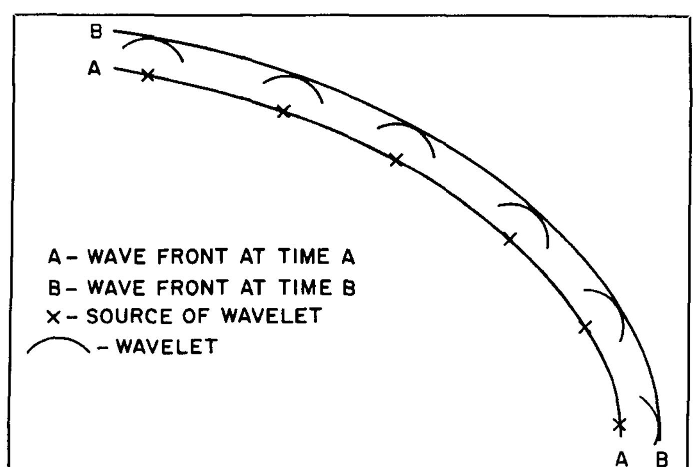

Figure 1. Illustrating Huygens Principle, where a new wave front is created by the envelope of wavelets originating from each point on the original front.

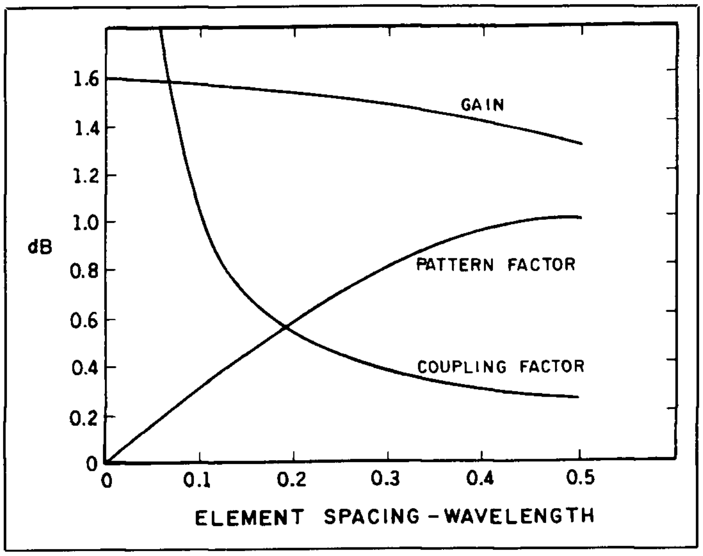

Figure 2. Theoretical gain of a two-element end-fire array, or 8JK antenna, showing the pattern factor, the coupling factor, and the overall gain. Element resistance is neglected.

The mutual resistance is zero at very large distances. It increases cyclically as the separation decreases, then monotonically, and is equal to the radiation resistance of a single element when the spacing is zero. Consequently, as the spacing of the dipoles is reduced, the magnitude of the coupling term grows larger, while the angle term becomes smaller. The overall effect is illustrated in **Figure 2**, which shows the magnitude of the two terms as a function of distance, and the magnitude of their product, the gain. The surprising result is that an array of two dipoles at near zero separation has a gain of 4 dB—nearly 3 dB more than that of a single dipole occupying essentially the same space.

Now let's follow an approach used by Schelkunoff and Friis.[2] Suppose we have a two-element array, with some spacing, say a quarter wavelength, and that these are fed out of phase with a current of one ampere. The currents can be represented by the number pair 1, −1. In the plane at right angles to the antennas, the radiation pattern is the figure 8 of the end-fire array.

Now suppose that each element is replaced by a pair of elements, one fed by a current 1, −1, the other by −1, 1. Let the spacing be half the original spacing, or 1/8 wavelength. Suppose the total length of the array is kept the same as the original, or 1/4 wavelength. Now two elements are at the same place at the center of the array, so they can be replaced by a single element carrying twice the current. The array now consists of three elements, with currents of 1, −2, 1. The pattern is the figure 8 of the original antenna pair pattern multiplied by another figure 8 due to spacing between the two pairs of antennas. The directivity of the antenna has been increased, even though there has been no change in overall dimensions.

In principle this process can be extended infinitely by adding elements and changing the excitation currents. The successive currents are:

Two elements,	1, −1
Three elements,	1, −2, 1
Four elements	1, −3, 3, −1
Five elements	1, −4, 6, −4, 1
Six elements	1, −5, 10, −10, 5, −1
etc.	

Each time there is a further increase in gain.

The minus currents mean that the particular element is delivering power to the exciting system. This is not a common design problem, but with inductors and capacitors of low resistance it can be handled with standard design equations.

While these approaches aren't rigorous in the mathematical sense, they can be made so. Thus, the statement that any desired amount of gain can be developed from an antenna of arbitrary size is soundly based.

Antennas that develop gain or directivity beyond that which is normal for their size are commonly called "supergain" antennas, and are said to be in the supergain regime.

Dimensions of supergain antennas

Work by Harrington[3] and Chu[4] provides a method of determining when an antenna is normal and when it falls into the supergain category. (A summary of this is found in Hansen[5]). Let the antenna be enclosed in the smallest sphere that just surrounds all the antenna elements. The maximum gain that can be developed by an antenna operating by the normal range is:

Gnmax = (2*PI*A/LAM)squared + 4*PI* A/LAM **(2)**

Where A is the radius of the enclosing sphere, and LAM is the wavelength. Any antenna having greater gain is in the supergain region. This equation is plotted in **Figure 3**.

As an example, consider a half-wave dipole. The radius of the enclosing sphere A/LAM is 1/4, so:

Gnmax = (6.28/4)squared + 12.56/4 **(3)** or a gain of 5.6 relative to isotropic, or 7.4 dB. Because the known gain of a practical dipole is 2.14 dB, it definitely falls within the normal gain range.

However, assume that the dipole is shortened to a length of 1/16 wavelength by loading. For this case, from the figure, Gnmax is 0.93 relative to isotropic, or to −0.27 dB. Now the gain of a shortened dipole approaches the constant value of 1.5 or +1.76 dB as the dipole length is shortened. Thus markedly shortened dipoles are in the supergain regime.

Now consider the two-element flat-top beam. When the elements are very close together, the sphere of containment is essentially the same as for a dipole—1/4 wavelength in radius. The supergain point is essentially 7.4 dB. Because the gain of a two-element flat-top beam approaches 4.0 dB, it also operates in the normal regime, but by a smaller margin than for a full size dipole. Decreasing the length of the elements in the array will place it in the supergain regime, as will using the method of element doubling described above.

For a one-wavelength loop, the basic element of quads, the sphere radius is half the diagonal of the loop or 0.185 wavelengths. From **Figure 3**, Gnmax is about 4, relative to isotropic, or 6 dB. Because the gain of a single quad loop is 3.4 dB, it is operating in the normal regime. As for the dipole, the gain becomes constant as the size is decreased, so very small loops are also supergain regime devices.

It seems clear from these examples that common basic antennas operate in the normal regime. It also seems clear that antenna designs can be thrown into the supergain regime without difficulty.

Supergain problems—drive resistance

In a pair of coupled circuits driven by a voltage in one circuit, the drive impedance is:

$$Z_{in} = Z11 - Z12^*Z12/Z22 \qquad \textbf{(4)}$$

Where:

- Z_{in} is the input impedance
- $Z11$ is the self impedance of the first circuit
- Z22 is the self impedance of the second circuit
- Z12 is the mutual impedance

When the coupled circuits are two identical antennas, as in the flat-top beam example, Z11 and Z22 are equal. Also, as the antennas are moved closer together, Z12 approaches Z11 in value. Accordingly, the drive resistance decreases markedly for elements that are very close together.

Actually, this isn't altogether bad. The distant field of an antenna is proportional to the current flowing in that antenna. For a given power, the current increases as the drive resistance falls. Thus, the gain also increases.

The phenomena of drive resistance decrease and current increase is even more marked when adjacent antennas are driven in the opposite polarity sense—a characteristic of the element doubling technique. Hansen[5] reports a calculation

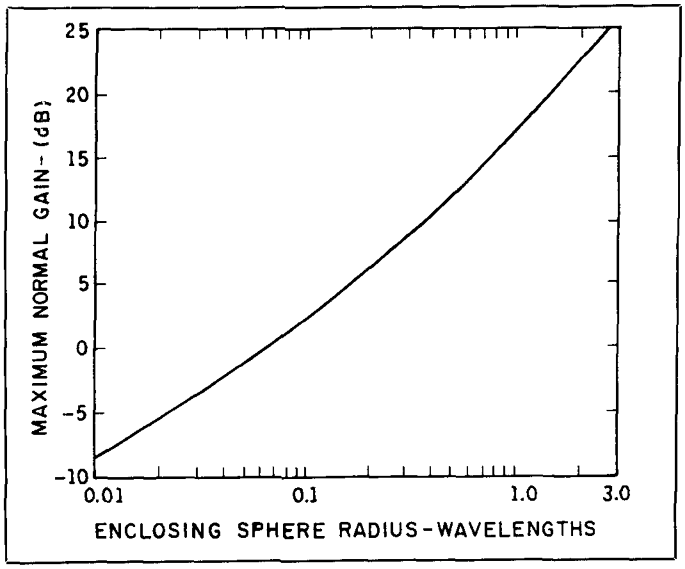

Figure 3. Maximum normal gain of an antenna from Harrington and Chu as a function of the radius of the sphere just enclosing the antenna. If the gain is less than the curve value, the antenna is in the normal (gain) range, if greater, in the supergain regime.

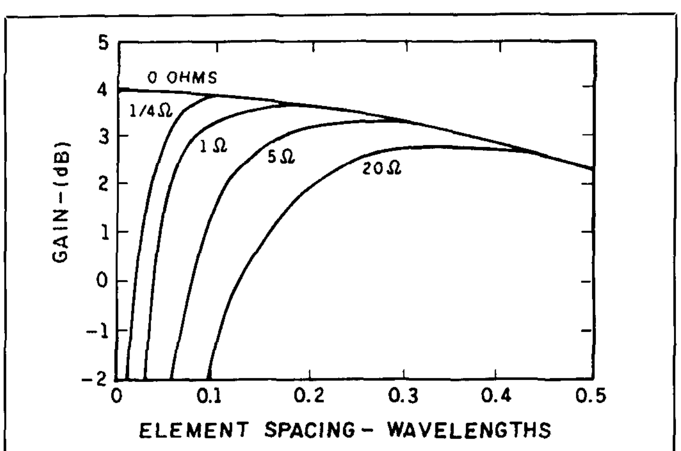

Figure 4. Effect of element resistance on the overall gain of a 8JK array. Compare to *Figure 2*. Calculation is based on resistance being distributed along the antenna length. Usual calculation assumes that the resistance is concentrated at the center of the element.

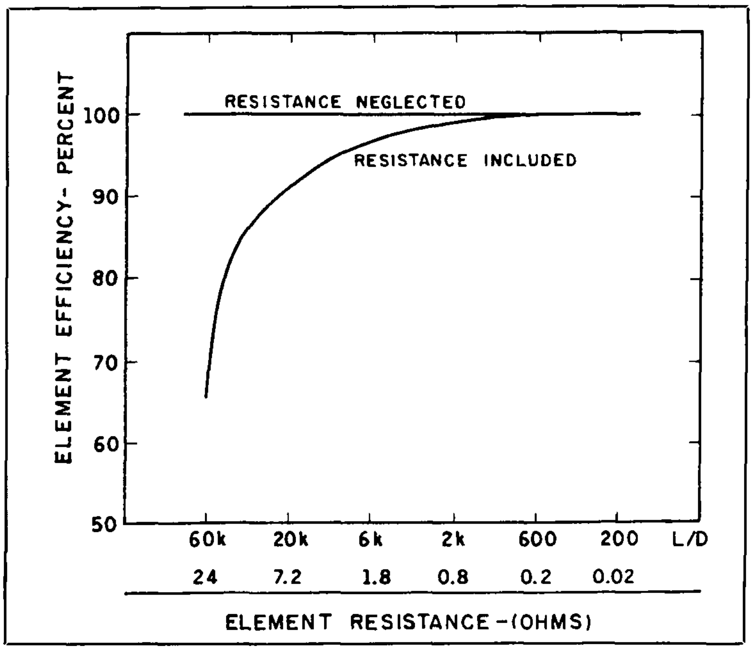

Figure 5. Calculated efficiency of a dipole as a function of element size. Based on a 2-meter long element at 75 MHz. Curve is approximately correct for any dipole of the same length/diameter ratio, but the exact value varies due to skin effect on element resistance.

made by Yaru, for a nine-element symmetrical array, compressed to 1/16 of its original size. The calculated currents in amperes are:

I0, I8	8,893,659,368.7
I1, I7	– 14,253,059,703.2
I2, I6	7,161,483,126.6
I3, I5	– 2,062,922,999.4
I4	260,840,226.8

The net current is only 19.5 amperes. To sustain the required currents, component values must be maintained to one part in 10,000 million.

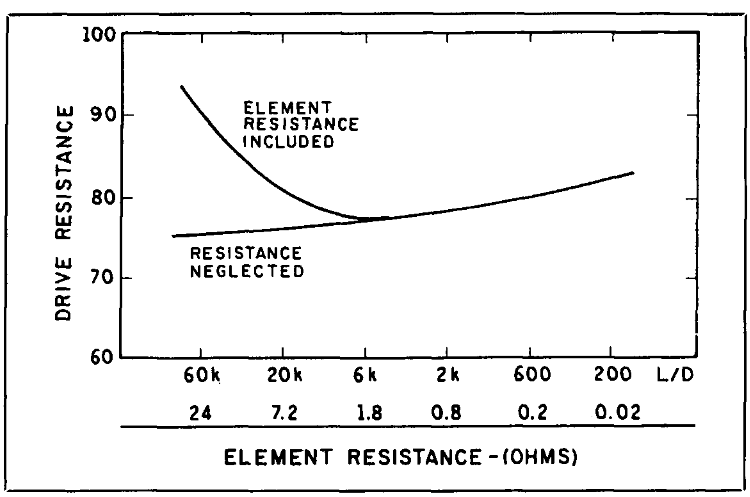

Figure 6. As for *Figure 5*, but showing the effect of element size on drive point resistance. See text.

Note the alternation of phase from one element to the next characteristic of this method of supergaining. Variations of the technique are found in the literature, with phase difference of less or more than 180 degrees. The characteristics of marked phase change between adjacent elements, and power being fed from some elements to the source network, exists in these variations.

The current values per element are much lower with fewer elements, as tabulated above, and the accuracy requirements are less severe. The limit of reasonable design is considered in the next section.

Supergain problems—element resistance

Real world antenna elements have some resistance, which must be taken into account. In **Equation 1** this enters into the coupling term, as:

$$G = SQR(2*(R11 + R1L)/(R11 + R1L - R12)) *SIN(D/2 \times Sin(Ang)) \quad (5)$$

Where R1L is the loss resistance.

The result of this resistance is shown by the curves of **Figure 4**. For a given value of resistance there is a spacing giving maximum gain. At lesser spacings the gain falls rapidly. The rapidly increasing currents in the elements cause an equally rapid increase in power lost in the element resistances.

One repercussion of the element resistance is that the antenna efficiency is less than 100 percent. The radiating efficiency is:

$$EFF = R11/(R11 + R1L) \times 100 \text{ percent} \quad (6)$$

This calculation is valid if the current on the elements is a half sine curve. While this is never exactly true, the departure from the sine curve is small in practical antennas.

Computer programs for calculation of antenna performance close to and in the supergain region should include element resistance as a factor. When using MININEC, this can be accomplished by introducing a loading resistance in the pulse at the center of the radiator. With standard MININEC, this operation must be performed manually. It's easy to write a small routine that will add this automatically, based on the length and diameter of the element.

A suitable relation for copper tubular elements, from Terman,[6] is:

$$R = 83.2*SQR(FREQ)/DIA \quad (7)$$

Where:

R is the resistance in milli-microhms per centimeter of length
FREQ is the frequency in Hertz
DIA is the conductor diameter in centimeters

For example, the resistance of 1-inch tubing at 14 MHz is 3.6 milliohms per foot. For aluminum, this value is multiplied by the conductivity ratio, or by 1.6. Check the value for the alloy used.

Slightly better resistance modeling divides the load resistance among the segments used in the model. This can also be done via a short routine added to MININEC. Some versions, and other antenna modeling programs have this included.

In the regime of normal operation, element resistance isn't of great importance unless the elements are very thin. **Figure 5** shows a MININEC calculation of efficiency for a dipole as a function of element diameter. Normal sizes of tubing and even wire don't have great loss. The situation is different, of course, with the high currents of the supergain regime.

Texts normally neglect other effects of element resistance. One is the fact that current distribution changes. **Table 1** shows the theoretical sine wave currents at several points, plus MININEC calculated currents with and without element resistance included. For the normal range of operation, the current change itself is negligible. However, there can be changes in drive resistance and reactance as shown in **Figures 6** and **7**. In particular, the change in reactance also means a change in resonant frequency. **Figure 8** illustrates this point for one particular wire size. The effects are appreciable in the wire sizes found in "invisible antennas."

Supergain problems—Q and bandwidth

Because individual element currents are high while the effective radiating current remains low, a large amount of energy is stored in the space surrounding a supergain antenna. This is a way of saying that the Q of the antenna is high. This, in turn, means that the usable bandwidth of a supergain antenna will be low.

The previously mentioned work of Chu and Harrington provides data on this situation. They consider that the radiation leaving the enclosing sphere can be described by a set of spherical harmonics. (The colored multi-segment bouncing balls found on computer displays are an example of these.)

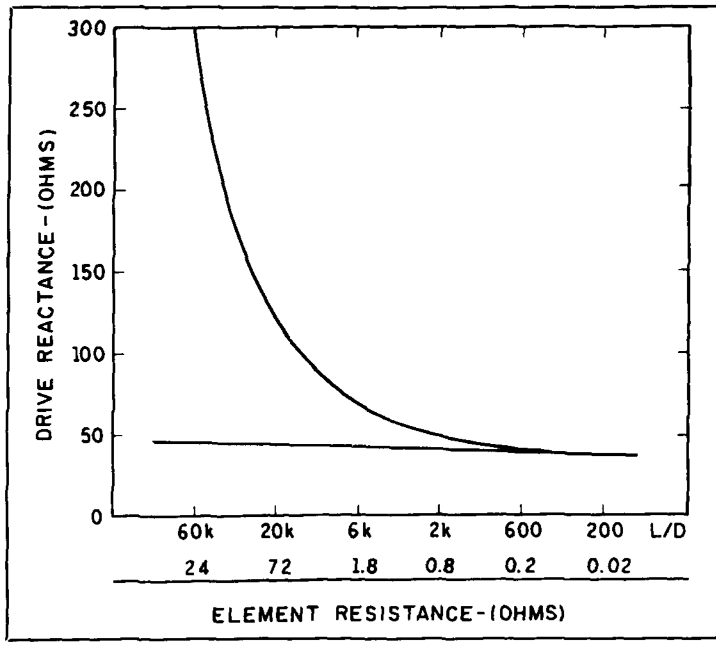

Figure 7. As for *Figure 5*, but showing the effect of element size on drive point reactance.

A large number of terms in the harmonic series is necessary to describe the pattern of narrow beam, high-directivity antennas.

If point-source antennas were really available, the number of sources in any size sphere would be unlimited. But in the real world, antennas must have physical size, preferably around a half wavelength long. Thus, practically, the size of the sphere is determined by the number of elements in the array.

The derived relation between antenna Q and the size of the enclosing sphere is shown in **Figure 9**. The number N on each curve is the number of terms in the harmonic series. For our purposes, the number of terms is also the number of radiating

Table 1		Dipole	Current		
		Percent	Length		
% Max. Current	0.00	12.50	25.00	37.50	50.00
Ideal	0.00	40.67	74.31	95.10	100.00
No Res.	0.00	41.66	73.93	84.31	100.00
With Res.	0.00	43.36	76.45	96.17	100.00

Condition: L=2 Meters, L/D=20,000, F=75 MHz, Material=Copper

Table 1. Distribution of current in a dipole for three calculation conditions. Ideal assumes sine-curve distribution, as usually found in textbooks. No Res. is calculated by MININEC with no element resistance. With Res. is calculated by MININEC with element resistance divided among the 40 segments used in calculation. See text.

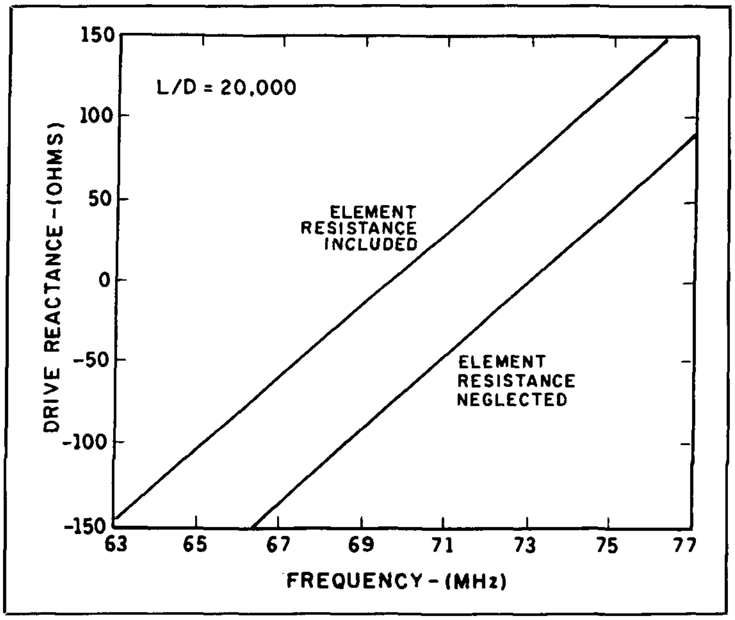

Figure 8. Change in reactance with frequency for a fine wire element with element resistance neglected and considered. The resonant frequency is affected, with a small change in reactance slope. The effects in *Figures 5* through *8* are commonly neglected in antenna textbooks.

elements in the antennas, each being a half-wave long—or at least a large fraction of this length.

The intersection of each curve with the Q = 1 axis is the size of the enclosing sphere for an antenna that lies just on the boundary between normal and supergain regimes. This point can also be taken as the normal design point for an antenna of N elements.

If supergain operation is attempted by reducing the length of the elements or moving them closer together, operation moves along the numbered curve to the intersection with the new enclosing sphere size. Accordingly, the antenna Q increases. This is the consequence of the increase in energy stored in the space surrounding the antenna. The nature of the curves is such that the Q change for antennas with a large number of elements becomes very high if appreciable size change is attempted. This is clearly evident in **Figure 10**, which shows the Q for a 2:1 improvement in gain directivity as a function of the original enclosing sphere size.

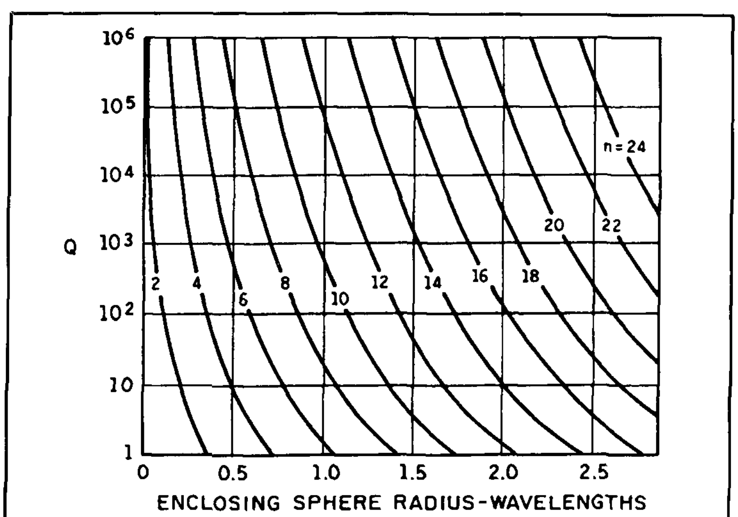

Figure 9. Quality factors for antennas, specifically for the TMmn and TEmn propagation modes, as a function of the enclosing sphere radius in wavelengths. The quality factor is the ratio of stored to dissipated energy, essentially the common Q. See text for usage. After Harrington and Chu.

Given the present state of knowledge and technology, size reduction and supergaining must be limited to antennas that are initially small; that is, antennas with relatively few elements. The situation may change if room temperature super conductors become available, but the matter of bandwidth will still need to be considered.

To appreciate the importance of this fact, recall that a Q of 1000 would mean that an antenna on 80 meters would have a bandwidth of 3.5 kHz. Even a Q of 100 is troublesome if variable frequency operation is needed. Perhaps a new term would be in order: instead of being rock-bound, the supergain antenna station would be wire-bound to a single frequency.

This antenna bandwidth limitation isn't the same as that usually encountered in feeding the antenna. Antennas themselves are wideband when operating in the normal radiation regime. The proof lies in the fact that retuning of the "match-box" restores full operation. Because the antenna hasn't changed, the usual limit is due to the matching system, not the antenna.

Supergain and the dipole

As I noted earlier, physically small antennas are easy to supergain. This is true of short dipoles. As the length of a dipole decreases (in fractions of a wavelength), its gain drops slowly from the initial 2.15 dB to a constant value of 1.76 dB. Even an elemental dipole has this gain. From **Figure 3**, any dipole which fits in a 0.08-wavelength sphere (that is, one less than 0.16 wavelengths long) is operating in the supergain regime. The same is true for a ground-mounted vertical less than 0.08 wavelengths high.

The practical aspects of using short dipoles are well worked out. Just to review, the large reactance at the feed point can be canceled by a local reactance of opposite sign by adding a loading coil, or by adding capacitance at the dipole end. Often a combination of the two is used. This leaves a resistance: the sum of the low radiation resistance of the dipole, plus the loss resistance of the dipole and the added elements. Separate impedance transformation may be used, or it may be combined with the reactance-canceling elements. Note that these steps don't change the operating mode of the antenna. For practical detail, see, for example the work of Hall[7] on dipoles and Sevick[8] on verticals.

A typical 2-meter "Rubber Duckie" is 6 inches long, about 0.16 wavelength. Assuming the body of the transceiver is the effective ground, this is operating in the normal regime. The Short Duckie, 3 to 4 inches long, is just at the edge of normal operation.

Antennas in the 160-meter range, in particular, are often in the supergain regime. For example, a 40-foot vertical has a length of about 0.075 wavelengths, just within the supergain regime. The feed resistance is about 2.3 ohms—not too difficult to feed. But an 8-foot whip has a length of 0.015 wavelengths. Reflecting the penetration into the supergain regime, the feed resistance is about 0.2 ohms. This antenna system will be low efficiency because of loss in matching and in element resistance. However, it will have the same gain as the larger vertical—1.5 dB. The hoped-for room-temperature superconductor would have a big impact on the whip system.

Top hat loading is used to increase the feed resistance. Consider a common situation: a 20-meter dipole at 40 feet, with the feed line ends tied together and fed against ground for operation on a lower band. On 160 meters, the enclosing sphere radius has increased to 0.08 wavelengths—just out of the supergain regime. MININEC gives a gain of 1.8 dB for this condition, with a drive resistance of 4.8 ohms.

Top loading an 8-foot whip also helps, even though operation may still remain in the supergain regime. This is the principle of the DDRR antenna (described in *The ARRL Antenna Handbook*[9]).

The fact that a simple element can be operating in the supergain regime means that there are four array regime types:

Normal array, normal elements
Normal array, supergain elements
Supergain array, normal elements
Supergain array, supergain elements

A close-spaced array of short loaded dipoles would be an example of the last array.

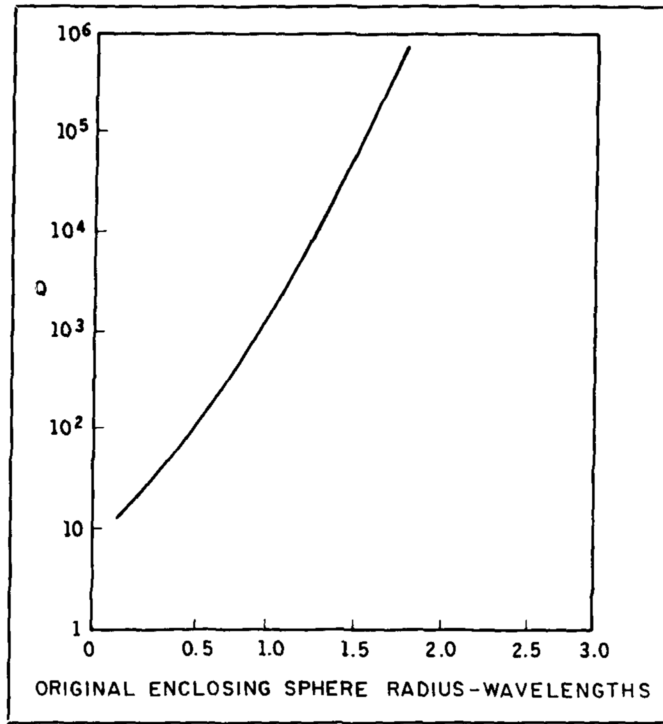

Figure 10. Quality factors for antennas reduced 2:1 in size, as a function of the original enclosing sphere radius. Note that reducing the size of large antennas means that they rapidly become narrow-band devices.

Supergain and end-fire arrays

The earlier discussion of end-fire arrays promised an investigation of practical operation. Let's start with the two-element array fed out of phase—the 8JK array. The general results with this beam alone are shown in **Figure 11**, and are based on data in Lawson.[10] The figure illustrates the ex-

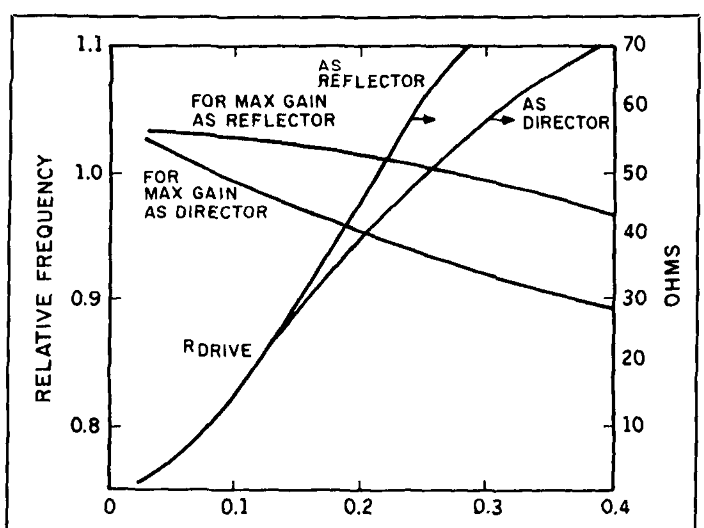

Figure 11. Effect of element spacing of a two-element Yagi, both elements 0.5 wavelengths long. Gain increases slowly as spacing is reduced, but drive resistance decreases rapidly, becoming zero at very close spacing.

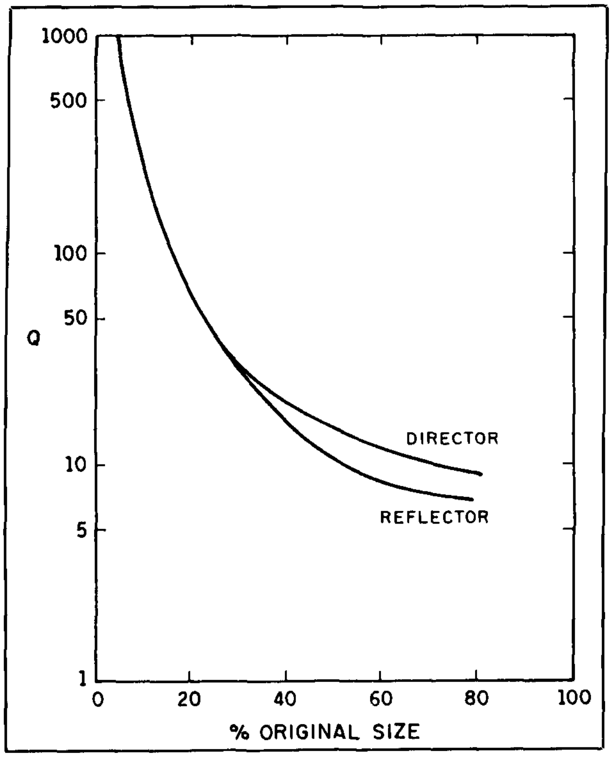

Figure 12. Quality factor for a two-element Yagi, assuming that original size is 0.5 wavelength spacing, and that spacing is gradually decreased. This is the drive point Q, the ratio of drive reactance to drive resistance, and is not the same as that shown in *Figure 9*.

pected gain increase and resistance decrease as spacing is reduced. Element resistance is not included.

It's convenient to use the ELNEC version of MININEC to investigate the effect of conductor resistance on supergain operation. ELNEC is set up for current feed, while the voltage feed of standard MININEC requires successive approximations to set currents precisely.

The example is based on 2-meter long elements, with a copper element radius of 0.00625 meter, about one quarter inch, operating at 73 MHz. Calculations were made with and without element resistance included. Results are essentially the same for a 20-meter element one inch in diameter.

For a single 8JK, with a spacing of one meter and no element resistance, the drive impedances of each of the two elements are 33.4 +j47.8 ohms. Gain is 5.7 dB, with a beamwidth of 90 degrees. With resistance included, the impedances are 33.6 +j47.8 ohms. This antenna is operating in the normal regime, with no special problems and good efficiency.

Now suppose two of these antennas are reduced to half spacing and combined into a three-element array of the same boom length, as described above. Feed currents are 1, −2, 1 amperes. With no resistance, the drive resistances are −16.8 −j6.5, 10.4 +*17.6* and −16.8 −j6.4 ohms. The minus resistances mean that the two end elements are capturing power and feeding it to the center element. The gain has increased to 7.3 dB, and beamwidth has decreased to 56 degrees.

With the effect of resistance included, the drive resistances are −16.7 −j6.4, 10.5 +j17.6 and −16.7 −j6.4 ohms. Gain has decreased, but only to 7.0 dB. Except for the matter of designing three feeds, this amount of supergain appears practical.

With four elements at 0.333333 meters spacing and currents of 1, -3, 3, −1 amperes, the no-element resistance drive impedances are −2.7 −j24.3, 0.33 −j0.93, 0.33 −j0.93, and −2.7 −j24.3 ohms. Gain has increased further, to 8.4 dB. Beamwidth is 48 degrees.

With element resistance considered, the impedances are −2.6 −j24.3, 0.45 −j0.93, 0.45 −j0.93, and −2.6 −j24.3 ohms. Gain falls dramatically from the ideal condition, to 0.70 dB, with a beamwidth of 90 degrees. This amount of supergain has become impractical.

An additional amount of supergain could be attained if the initial pair spacing was increased, or if the element resistance were decreased by a larger diameter or by silver plating. But the complexity of multiple feeds and extremely low driving resistance doesn't make the attempt to increase the amount of supergain appear practical.

However, there's another feed technique that avoids part of the problems. The starting point is the unidirectional end-fire array, where the phasing between elements is equal to their spacing in degrees. For the four elements at 0.33333 meter spacing, the element currents are one ampere at 0, 29, 58, and 87 degrees.

With no element resistance, the drive impedances are 229 −j116, 225 +j40, 211 −j61, and 107 −j130 ohms. The main lobe gain is 3.9 dB, with a beamwidth of 76 degrees, and the back lobe is −2.9 dB. With element resistance included, the element resistances are again 229 +j116, 225 +j40, 211 −j61, and 107 −j130 ohms. Gain and beamwidth remain the same. Except for the multiple feeds, this antenna is quite practical.

Hansen and Woodward[11] showed that the directivity of this type of antenna could be increased if the phasing between elements were increased. They concluded that the best design was obtained if the total phase shift across the antenna was equal to the boom length in degrees plus 171.9 degrees (3 radians), with the shift between each element in proportion. For the four element 0.3333 meter spacing, the phases are 0, 87, 174, and 261 degrees.

With these conditions and no element resistance, the computed element impedances are −24.2 +j77.0, 31.1 +j36.9, 21.6 +j21.0 and 49.5 +j1.0 ohms. The main lobe gain is 7.8 dB, with a beamwidth of 60 degrees. The backlobe gain is 1.6 dB. The backlobe level is higher than theoretically obtainable. The reason for the high level wasn't investigated, but may be related to the use of spacing below the best value, or to the rounding of the phase relations.

The inclusion of element resistance changes drive impedances negligibly to

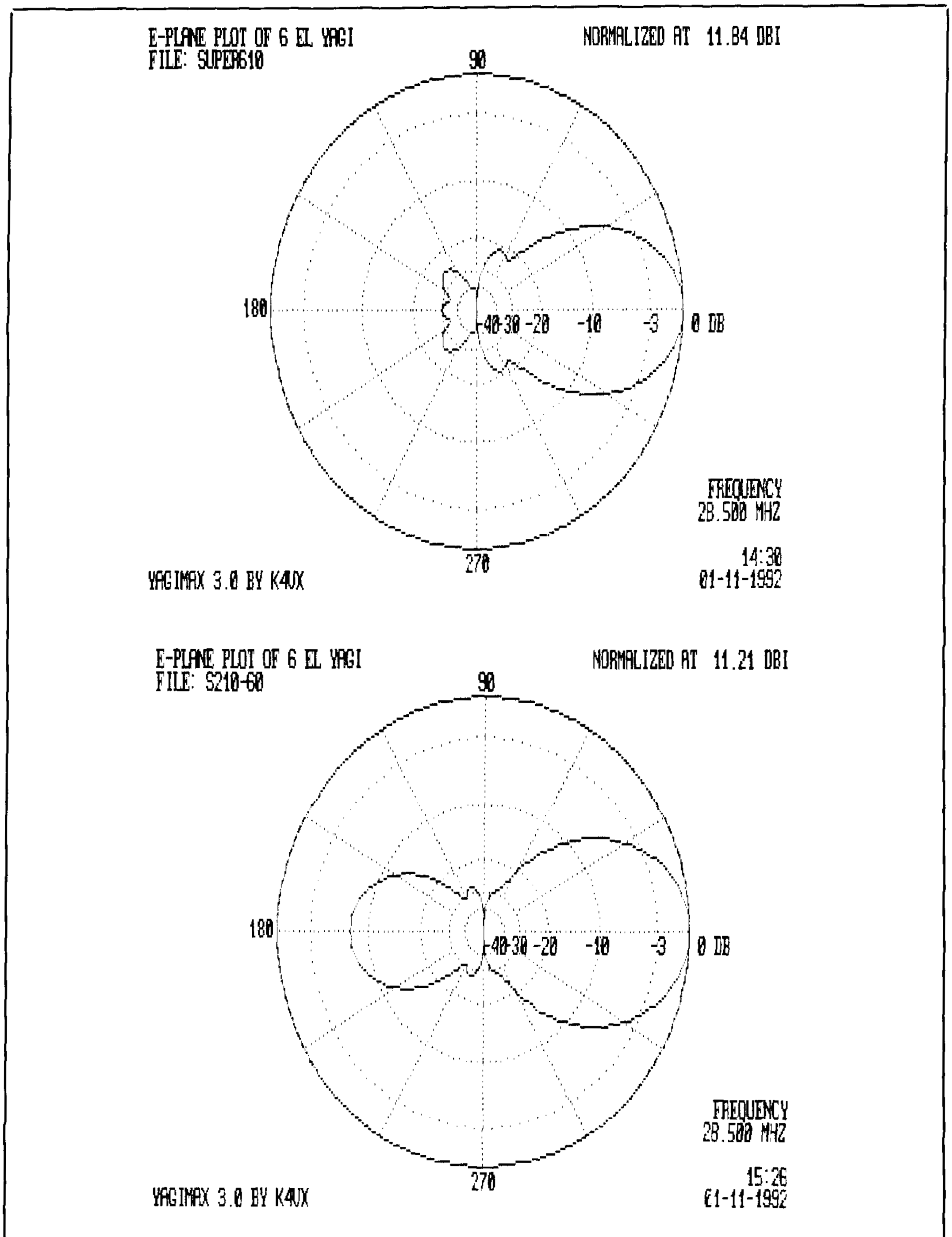

Figure 13. Effect of shrinking the boom length of a six-element Yagi to 60 percent of its original size, and retuning some elements. A is for the original size, B for the reduced size. Further element adjustment can increase the smaller antenna gain. See text.

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FREQ (mHz)	GAIN (dBi)	F/B (dB)	IMPEDANCE (ohms)
28.000	11.50	20.91	24.94-j19.35
28.050	11.54	21.93	24.59-j18.47
28.100	11.59	23.14	24.15-j17.50
28.150	11.63	24.64	23.64-j16.43
28.200	11.67	26.55	23.08-j15.25
28.250	11.71	29.15	22.51-j13.96
28.300	11.75	33.09	21.96-j12.57
28.350	11.78	40.85	21.45-j11.10
28.400	11.81	45.83	21.02-j9.58
28.450	11.83	34.64	20.70-j8.05
28.500	11.84	29.87	20.49-j6.57
28.550	11.85	26.89	20.39-j5.21
28.600	11.84	24.82	20.37-j4.09
28.650	11.82	23.36	20.33-j3.34
28.700	11.79	22.39	20.05-j3.06
28.750	11.75	21.92	19.14-j3.24
28.800	11.69	22.06	17.17-j3.51
28.850	11.63	23.08	14.04-j3.15
28.900	11.54	25.79	10.30-j1.44
28.950	11.42	31.63	6.87+j1.66
29.000	11.22	24.90	4.31+j5.62

Center VSWR Frequency (mHz) 28.500

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Table 2. Gain and drive impedance versus frequency for a six-element one-wavelength boom Yagi (Yagimax file SUPER10).

−24.0 +j77.0, 31.2 +j36.9, 21.7 +j21.0, and 49.6 +j1.0 ohms. Main lobe gain remains at 7.8 dB (actually a reduction of 0.018 dB) with a beamwidth of 60 degrees. The back lobe increases by 0.02 dB.

This method of increasing directivity is practical in this size antenna if single-frequency operation is needed. The technique may be extended to more elements, but feed network complexity seems too great to expect extensive use.

Supergain and the Yagi

We can begin our study of Yagis in the supergain regime by reviewing the extensive data in Lawson's book.[10] His **Figures 2.4** and **2.5** show that the gain of a particular two-element length combination is low at small and large spacings, with a maximum at some intermediate spacing.

If element lengths are adjusted at each spacing, the gain increases as the spacing is reduced, as shown in his **Table 2.1**, and summarized in **Figure 11**. The gain increases at least down to 0.025 wavelength spacing, with a limit of about 7.5 dB. As the spacing decreases, the drive resistance decreases, from about 30 ohms at a typical 0.15 wavelength spacing to 1.1 ohm at 0.025 wavelength.

At close spacing, the enclosed sphere radius is 0.25 wavelength, so the maximum normal gain is 8.5 dB. The close-spaced two-element Yagi is just outside the supergain regime. If the elements are shortened to less than about 0.2 wavelengths by loading, the beam will enter the supergain regime.

Figure 12 shows the Q of the two-element Yagi as calculated by Lawson. The plot is based on the assumption that the initial spacing is 0.5 wavelength, and that the smaller sizes result from attempts to move into the supergain regime by reducing the element spacing. The Q is the ratio of element reactance to resistance, and is not precisely the same as that of **Figure 9**. However, this Q is a true measure of array bandwidth.

Consider now a multi-element beam, using **Figure 2.9** in Lawson. He shows a four-element beam with a 0.7-wavelength boom to have a gain of 11 dB if designed for maximum gain. The enclosing sphere radius is 0.43 wavelengths, with the maximum normal gain being just over 11 dB. This antenna is at the boundary between the normal and the supergain regimes.

Suppose this antenna is shrunk until the boom length is 0.4 wavelength. For this condition, the enclosing radius is 0.32 wavelengths, with a maximum normal gain of 8 dB. For the tuning conditions used by Lawson, he calculates a gain of 9.5 dB. The shrunk beam is in the supergain regime.

Lawson doesn't report the drive resistance for these beams. It isn't possible to determine if the loss of gain in shrinking the beam is due to element loss, or if different element lengths would restore the original gain. In general, Lawson slanted his study to a balance of good performance features: gain, F/B ratio, and easy drive. It appears that he avoided the supergain conditions purposely, because this demands "single-minded" attention to gain.

We can investigate Yagi supergain by using the computer program YAGIMAX, which includes routines for gain optimization. One design in the files included with the program is SUPER610, a six-element beam on a 1-wavelength boom. The program gives a calculated gain of 11.84 dB at 28.5 MHz, with a reduction of 0.62 dB or less at 28 and 29 MHz. F/B ratio is nearly 30 dB.

The enclosing sphere radius is 0.56 wavelengths, for a maximum normal gain of 13 dB. The antenna is in the normal regime. The calculated pattern of the array is shown in **Figure 13A**.

Suppose the boom is shortened to 60 percent of original length. With no other change, the effect is relatively small. The gain reduces to 10.35 dB, a loss of 1.5 dB. F/B ratio drops to 15 dB. The drive resistance also decreases, from the original 20.7 ohms to 12.5 ohms. Because the enclosing sphere radius is 0.39 wavelengths, the normal maximum gain is 10.5 dB. As also shown by the relatively high driving resistance, the antenna remains in the normal regime, but by a small amount.

Changing to the supergain condition requires a retuning of elements. Using the optimization routine included in the program on the reflector has a small effect, changing from 10.36 dB gain to 10.41 for a length reduction of 4.5 inches. F/B increases slightly, to 16.2 dB. Drive resistance decreases to 10.22 ohms.

A small additional effect is obtained by leaving this condition set and changing the first director. A 5.8-inch increase in length boosts the gain to 10.44 dB, and reduces the drive resistance to 2.5 ohms.

The second director clearly illustrates the problems of supergain. A 20-inch reduction in this director length increases the gain to 11.22 dB. F/B ratio drops to 7.3 dB, but the drive resistance drops to about 0.8 ohm. The step also produces pattern changes, the most marked being an increase in back lobe size (see **Figure 13B**).

The detail performance is evident when one compares **Table 2**, for the original beam, and **Table 3**, for the size-reduced retuned beam. The steps have converted an easily fed Yagi of good performance to a narrow-band, hard-to-feed design. These are penalties incurred for maintaining the same gain in a beam of 60 percent of the original boom length.

It's probable that additional gain could be developed by tuning the other directors. However, another effect must be watched. All analysis programs I've used become inaccurate when the degree of supergain grows large. The most noticeable effects are very high values of gain, sometimes reaching *100 times (+20 dB)* as compared to usual values, and negative values of drive resistance.

A large part of the problem is the very high element currents calculated, and the mutual effects of these currents. The problems can be alleviated by using double-precision arithmetic and a large number of segments in moment-analysis programs. However, the steps aren't really worth while, because the drive resistance moves out of reasonable range. In fact, as Kraus noted, supergain drive resistances can drop below values that can be reached by the best transmission line short known.

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FREQ (mHz)	GAIN (dBi)	F/B (dB)	IMPEDANCE (ohms)
28.000	10.10	10.14	4.43-j31.80
28.050	10.17	10.42	3.77-j29.77
28.100	10.25	10.59	3.21-j27.68
28.150	10.34	10.64	2.72-j25.57
28.200	10.44	10.55	2.31-j23.43
28.250	10.56	10.33	1.95-j21.29
28.300	10.68	9.97	1.65-j19.14
28.350	10.82	9.49	1.39-j16.99
28.400	10.95	8.88	1.16-j14.84
28.450	11.09	8.17	.98-j12.69
28.500	11.22	7.38	.82-j10.55
28.550	11.32	6.51	.69-j8.41
28.600	11.38	5.59	.58-j6.28
28.650	11.37	4.61	.49-j4.15
28.700	11.25	3.59	.43-j2.02
28.750	10.98	2.52	.38+j9.41
28.800	10.51	1.41	.35+j2.21
28.850	9.80	0.25	.34+j4.32
28.900	8.85	-0.95	.34+j6.42
28.950	7.67	-2.19	.36+j8.53
29.000	6.30	-3.47	.39+j10.63

Center VSWR Frequency (mHz) 28.500

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Table 3. Gain and drive impedance for the antenna of *Table 2* reduced to 60 percent of its original length, with reflector and directors 1 and 2 retuned. See text for steps followed.

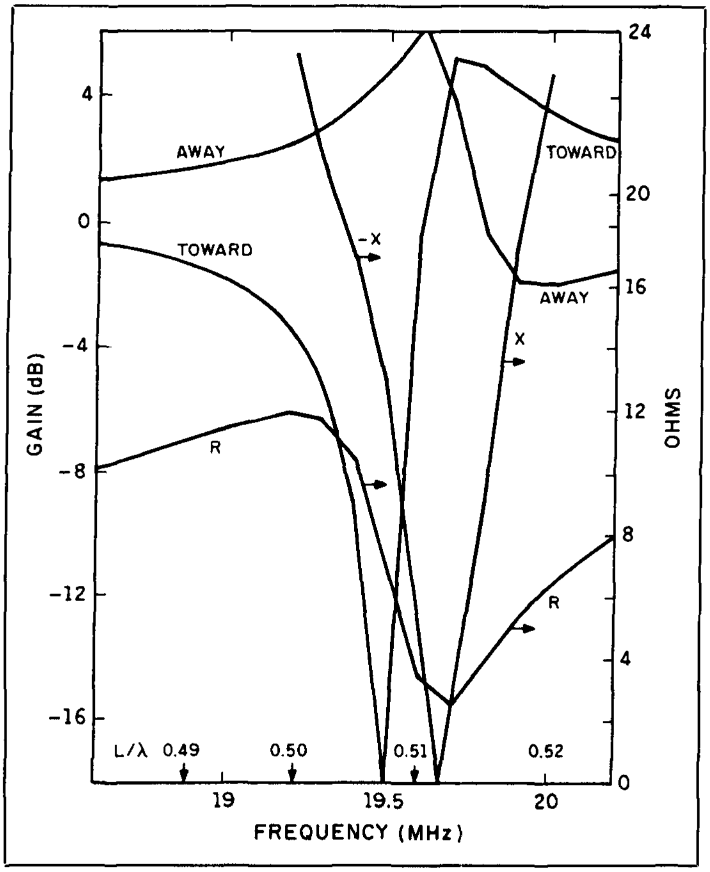

Figure 14. Swept frequency performance of a close-spaced two-element open-loop quad. Elements are of copper tubing and identical in length. The marked change in drive resistance is characteristic of close-spaced arrays, and is not normally found at wide spacing. This antenna performs quite well in a small space.

The overall conclusions for Yagis are:

- Conventional Yagi maximum or near maximum gain designs are at or close to the lower limit of the normal operating regime. Quite small changes will throw the design into the supergain regime, with its attendant problems.
- If the problems can be accepted, boom lengths can be shortened appreciably.
- Although not shown here, it appears that a better supergain design compromise can be reached if the gain goal can be secured with director tuning only; this leaves reflector tuning for F/B ratio adjustment.

Another point to remember is that the single feed point of the Yagi is much simpler than the multiple feeds necessary with end-fire arrays.

Stacking

There is, however, another important possibility that we have ignored so far. The antennas considered here have been one or two dimensional, but the enclosing sphere is three dimensional. It would seem possible to obtain better gain, while avoiding the problems of supergain, by going to three dimensional antennas. For Yagis, this means stacking.

Lawson includes some data on this topic. He shows a three-element 0.3-wavelength boom Yagi at one wavelength elevation above ideal ground as having a gain of 14.10 dB with a drive resistance of 16.6 ohms. Two such antennas spaced 3/4 wavelength, with the center at one wavelength, give a gain of 16.27 dB. Because the gain due to ground reflection is very nearly 5.0 dB, the free space gain of these antennas would be 9.10 and 11.27 dB.

The enclosing sphere for the single Yagi is about 0.29 wavelength; for the stacked pair, it's about 0.47 wavelength. The maximum normal gains are about 9.0 and 11.5 dB, respectively. The stacked pair remains in the normal operating regime, with none of the supergain problems. As compared to a long boom design of the same gain, the problems of boom length and rigidity have changed to a problem of vertical stacking and adjustment of double the number of feeds. The stacked array will usually have fewer elements for the same gain.

The overall conclusion seems to be that stacking should be considered for Yagis before supergain is attempted.

Supergain and the quad

In one sense, the quad in supergain conditions will behave the same as the Yagi. In the supergain regime, both antennas depend on the mutual impedance approaching isolated element impedance at close spacings.

However, there is one difference. The quad is already a three-dimensional antenna. Further, the gain of an individual quad loop is greater than that of a dipole. In the usual configuration, the boom length is the same as in Yagis, but the element span is half that of the Yagi, and the height is much greater.

A typical compromise-design three-element quad in free space will have a gain of 8.7 dB with a boom length of 0.45 wavelength. The enclosing sphere radius is 0.29 wavelengths, for a maximum normal gain of 9 dB. Such a compromise design is in the normal regime.

Suppose the boom length is shrunk to 0.288 wavelength (a 20.5-foot boom on 20 meters). If elements remain the same, gain will drop slightly to 8.37 dB. Drive resistance will also drop to 84 ohms. Gain can be increased by retuning the elements or by adding reactance. Reducing the reflector perimeter and increasing that of the director (i.e., bringing the element resonances closer) will provide gains as high as 9.8 dB. For this condition, the enclosing sphere radius is 0.23 wavelengths, with a maximum normal gain of 7 dB. The antenna is in the supergain regime. The drive resistance drops appreciably for this tuning, but is still 12.2 ohms and not particularly troublesome to feed.

A second way to reduce the size of the enclosing sphere is to reduce the size of the elements by loading. One can do this by using inductive loading at or near high current points, capacitor loading between high voltage points, and "linear loading" by folding the sides to resemble a three-wire transmission line. For example, it's not difficult to make a 20-meter size loop that's resonant on 40 meters. There's some loss in gain, but even very small loops have a gain of 1.5. As for the dipole, these loops go into the supergain regime. One result is that the four classes of antenna arrays tabulated above also exist in the quad family.

As an alternative to loading, suppose that the loop is opened at the side opposite the feed point. The loop is now parallel resonant at its original operating frequency. Series resonance occurs at half this frequency. In one sense, the series resonant quad loop has been converted to a bent dipole of square configuration, with sides of 0.125 wavelength. Such a loop fits in a sphere of 0.088 wavelength radius, for a maximum normal gain of 1.5 dB. Because the calculated gain of the bent dipole around the resonant point is about 0.2 dB, the loop is in the normal regime. Drive resistance is about 10 ohms.

Arrays of these open loop quads can have good performance. Consider two identical open loops of 0.00625 meter radius copper tubing, each 2 meters on a side, spaced 2 meters to give a cubic antenna. The 8-meter conductor length means that the resonance point will be around 18.75 MHz.

The MININEC-calculated performance of this antenna, including element resistance, shows a maximum director gain of 5.2 dB at 19.7 MHz, with a drive resistance of 2.5 ohms. F/B ratio is poor at only 1.8 dB. The maximum reflector gain is just over 6 dB, with a F/B of 6.3 dB, and a drive resistance of 3.4 ohms. Maximum F/B occurs for reflector action, and is 23.3 dB at 19.6 MHz. Gain, however, has dropped to 5.0 dB, with a drive resistance of 7.2 ohms. **Figure 14** shows the variation in gain toward and away from the parasitic, and the drive resistance versus frequency. The antenna is narrow band, the gain dropping to 3.8 dB at 19.4 and 19.7 MHz.

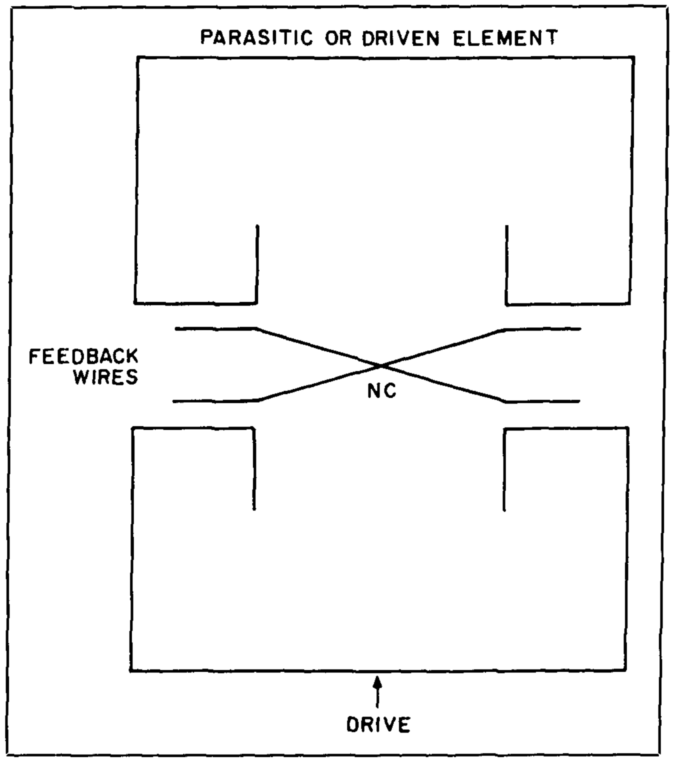

Figure 15. Principle of "antenna neutralization" as developed by Moxon, G6XN. Two dipoles are shortened by folding the element ends, resulting in over-coupling and low drive resistance. The coupling is opposed or "neutralized" by introducing a pair of short elements that couple small out-of-phase components into the wire ends. See text.

The antenna enclosing sphere is just 1.73 meters in radius, or 0.11 wavelengths at the best gain point. The maximum normal gain is 3.5 dB. The antenna is in the supergain regime. Despite this, the feed resistance isn't impossibly low. The calculated antenna efficiency is good—92 percent. This doesn't include the loss in the matching section.

A trial at a spacing of one meter showed essentially the same performance, with a gain of 6 dB at 19.7 MHz. However, drive impedance was markedly reduced to 0.9 ohm, and efficiency was 87 percent.

This design seems to be a good one for minimum space beams. These antennas could be scaled to 20 and 15 meters, by the multiplying ratios 1.39 and 0.925. It would probably be worthwhile to increase tubing size. It would be possible to compensate for the relatively narrow bandwidth by designing for the top end of the band and adding capacitance plates at open element ends for tuning. Experimental work on the matching system seems to be indicated; gamma and delta matches may be possible. It might be good to use voltage feed to the open end of the driven element.

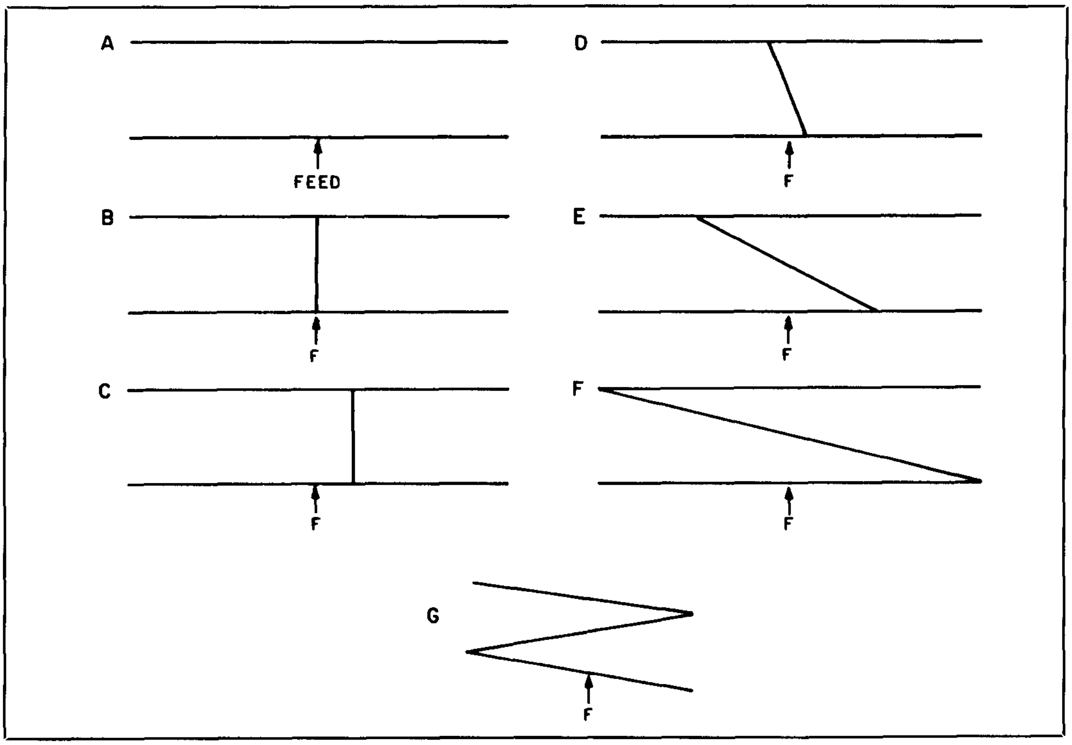

Figure 16. Schematic of a family of direct-coupled antennas. A, no coupling; B, neutral coupling; C, positive coupling; D, E, F, negative coupling of varying magnitudes; G, a variation of F. See text for concept.

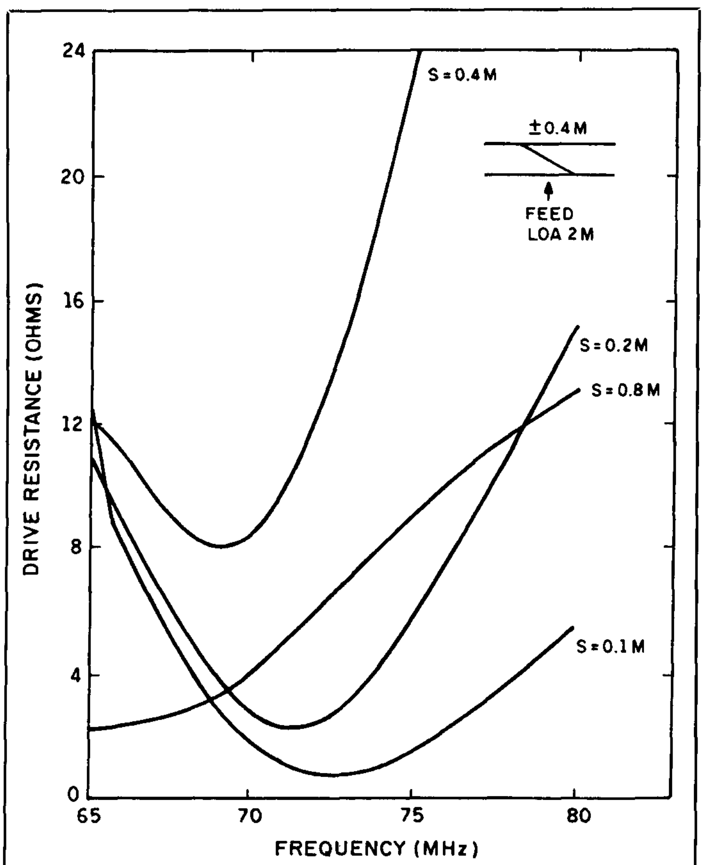

Figure 17. Drive resistance of negative direct-coupled antennas of varying spacing. Both elements identical in size.

Some data on multi-element design and other variations at or near the supergain regime is given in Haviland.[12]

Neutralization and the antenna

British amateur Les Moxon, G6XN,[13] has concluded that the reason for the low driving resistance of the size-reduced family of antennas is over coupling of the elements. This led him to the concept of antenna neutralization. The concept is the same as the neutralization employed in RF amplifier design, in which an out-of-phase voltage is coupled from the drive to the driven circuit.

Moxon's proposed physical concept for a two-element end-folded shortened beam is shown in **Figure 15**. An additional pair of elements are introduced. These are not resonant, but are regarded as transmission line sections feeding the small voltages capacitance coupled from the antenna element to the transmission line ends. The diagonal placement puts the voltages out of phase, as required for neutralization. Another physical layout is shown in the references.

This is not an easy design to analyze. A large number of wires are involved because of the number of bends. For a high degree of accuracy there should be a minimum of four segments per wire in the neutralizing area, preferably more. Also, the problems encountered in the analysis of very closely spaced wires appear.

So far, the results of MININEC analysis of capacitance-coupled neutralization have been inconclusive. Increases in drive resistance by a factor as large as 3:1 as compared to no neutralization were found. However in all cases tried, this increase was found at frequencies well away from those giving maximum or near maximum gain. At and around the maximum gain frequency, the drive resistance varied by no more than a few percent among the conditions tried. This included various spacings, plus direct connection with resistive, capacitive, and inductive isolation elements.

It's not apparent whether this finding is a real reflection of antenna performance, whether it's due to the known analysis limitations of MININEC, or whether a larger computer would give a confirmable answer. In an effort to gain an understanding of the concept, several approaches were tried.

In one, a small additional amount of excitation was introduced at the center of the second element of a two-dipole array. The magnitude of this excitation was varied up to 50 percent of the main excitation, both in and out of phase to the main excitation. Results as to drive resistance increase were again inconclusive, possibly because only 0 and 180 degree phase differences were used.

Figure 16 shows the genesis of a different approach. At **A** two parallel dipoles are shown, the basic Yagi or 8JK beam, de-

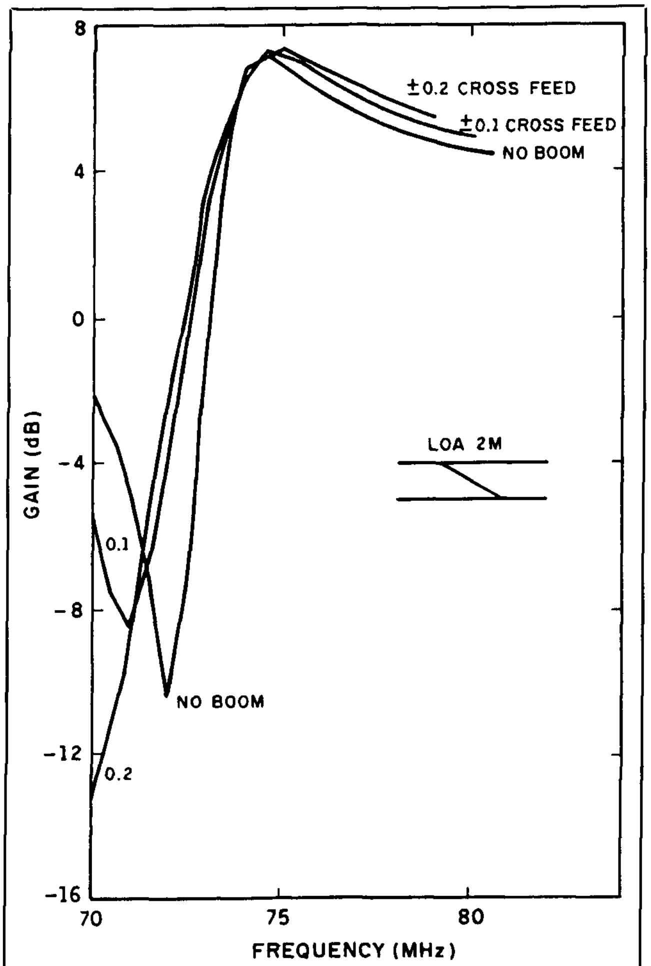

Figure 18. As *Figure 17*, but for gain away from the parasitic element (reflector action). Over much of the frequency range, this mode gives the best forward gain.

Antenna Type	Element Spacing 0.10 Gain	Res.	.20 Gain	Res.	.40 Gain	Res.	.80 Gain	Res.
No Boom	7.30	1.15	7.40	3.30	7.50	11.70	6.20	28.00
N1	7.20	1.00	7.20	3.50	7.10	11.00	6.10	30.00
N2	7.00	1.60	7.20	3.50	7.10	11.00	6.00	29.00
N3	7.40	.98	7.30	3.00	6.90	9.60	5.90	24.00
N4	7.40	1.00	7.20	2.80	6.70	8.40	5.70	22.00
N8	7.40	10.40	6.50	9.30	6.40	11.00	6.10	39.00
Nend	4.70	1.10	4.30	32.00	4.30	172.00	4.80	309.00

Condition: Max. Calculated Gain any Frequency
Element Length= 2 Meters
L/D= 20,000
Spacing in Meters
Neutralize Element Span= 1/ 10 Net Number, Meters
Copper
Frequency 75 MHz +/- 5
Gain, DB
Res.,Ohms

Table 4. Gain and drive resistance for a family of two direct-coupled dipoles, with isolated element conditions for comparison. Note that the effects are small until the end separation of the neutralizing element becomes large, as in E and F of *Figure 16*. MININEC calculations with element resistance included. These results were the first indication that the concept of antenna neutralization can increase the drive resistance.

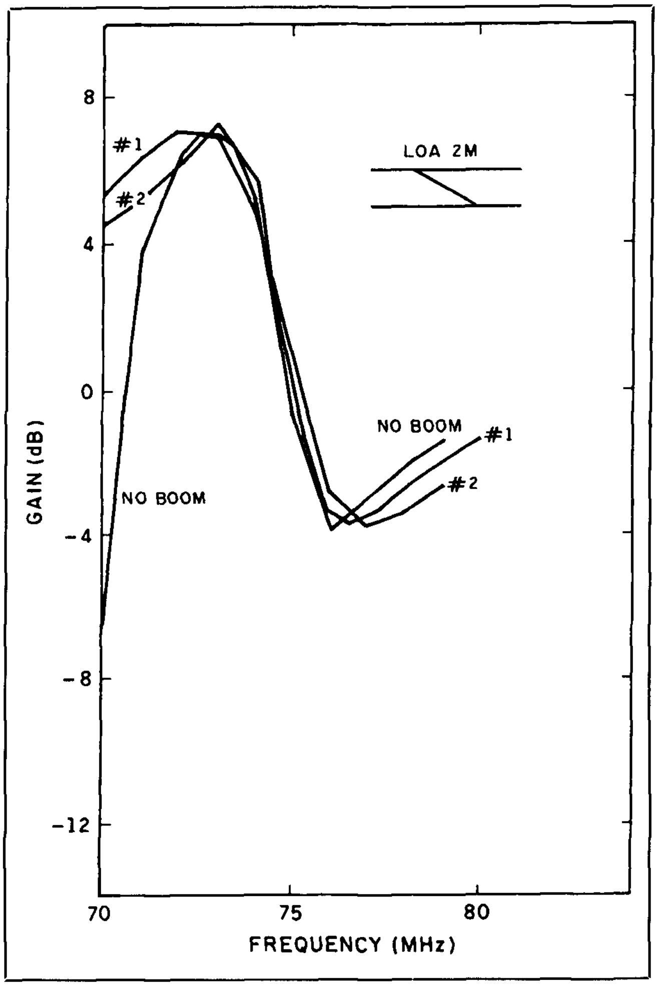

Figure 19. As *Figure 17*, but for gain toward the parasitic element (director action).

pending on excitation. If a connecting boom is introduced, as at **B**, and this boom is symmetrical, it has only a small effect—primarily equivalent to shortening the elements. (See the discussion in Lawson[10] about this.)

However, if the boom isn't symmetrical, it will introduce an additional current path. This path can be in the direction of increasing element coupling, as at **C**, or of reducing it, as at **D**. The amount of coupling can be changed by changing the intersection points, as in **E** and **F**. A further change can be made in the length of the coupling diagonal, which requires canting the elements, as shown in **G**.

Analysis of this family of designs seems to show convincingly that the concept of antenna neutralization is valid. The first indication found is summarized in **Table 4**. Here, the gain and drive resistance are tabulated for four spacings, with no boom and with symmetrical cross connection at five points along the element—the last being at the ends. It was found that there is no particular effect when the connecting points are close to the center. But as shown, with widely separated diagonal connections, there can be marked increase in drive resistance—in some cases with gain loss, in others gain improvement.

Figures 17 to **19** show typical overall effects. **Figure 17**, drive resistance, shows a minimum near the single element resonant frequency, the minimum resistance increasing and the frequency of occurrence decreasing as the element spacing increases. **Figures 18** and **19** show the effect on gain away from and towards the undriven element, respectively, for two of the spacings. The most noticeable gain effects are on the low frequency side of the peak gain point, where the parasitic is acting as a director. Overall conditions can be found to give a marked increase in drive resistance, plus a small improvement in gain with little change in F/B ratio. This is generally Moxon's claim.

Much more analytical and experimental work on this concept is needed before design rules can be presented. This brief account is included here in the hope that it will stimulate such work by others.

The Zig-zag Beam

A number of years ago, Cumming[14] described the antenna of **Figure 16G**. He regarded it as a special case of the Helix antenna, wound on an ellipse of zero minor axis. It can also be regarded as the end point of a series of cross-diagonal feeds. In the form Cumming depicted, the antenna is used with a reflecting screen. The data given here is for the zig-zag (Z-Z) element alone, with no reflector, and for feed at the center of one end element.

The Z-Z is an end-fire array. The direction of maximum radiation is a function of element length and spacing. The gain in the direction away from the feed (taken from the center of the antenna) is shown in **Figure 20**. At each spacing, there is a change from reflector to director action. At close spacing, the change is quite rapid as frequency is varied. (Note that there is a change in element length as the spacing is varied.)

The gain in the direction of the feed is shown in **Figure 21**. This also shows the change from director to reflector action, but this is much less marked than for the other lobe. Maximum gain is a function of spacing, reaching 8.6 dB for the case having 0.1 wavelength between adjacent element ends.

Figure 22 shows the calculated drive resistances. The curves indicate a resistance minimum, which varies with spacing. Most, if not all, of the changes in the frequency of the minimum are due to the lengthening of the element as spacing increases, since the span of the antenna, rather than the element length, was kept constant. The magnitude of the resistance at minimum increases as spacing increases, reflecting the reduction in coupling. Note that the antenna approaches a long wire at larger spacings.

This data was developed using MININEC 3.12. It's likely that calculated values would be different with earlier versions that lack the small angle correction feature. The regular variation of the curves in **Figures 20** through **22** is an indication that the calculated results are trustworthy. Marked discontinuities over a range of conditions indicates that results are suspect, and should be checked. One way to do this is to increase the number of segments used in analysis.

The enclosing sphere for these antennas is around 0.25 wavelengths radius. The maximum normal gain is thus around 7.5 dB. At the wider spacings, the antenna is in the normal regime, but at the close spacings it shows supergain. This is indicated further by the calculated decrease in bandwidth. For example, for the 0.4-meter spacing, the gain varies by 1.4 dB over the range from 61 to 78 MHz, while for a spacing of 0.1 meter, the frequency range is from 72.5 to 76.5 for the same variation. Drive resistance also changes with spacing and frequency.

Cumming shows a design using two mirror-image zig-zag elements stacked closely, using a reflector screen. A measured gain of 6.2 dB at 64 MHz, increasing to 10.2 dB at 86 MHz was obtained. SWR was an average of 1.8 and a peak of 2.2 over this band using 300-ohm feed. Cumming notes that the antenna family can be regarded as a traveling wave antenna of the slow-wave class, with a propagation velocity of 0.91. This suggests that very wide band operation could be obtained if the length of each rod was decreased progressively, as in the log periodic.

The zig-zag is another antenna where more analysis and experiment is needed to develop practical design rules.

Summary

I started the work reported here when I encountered a supergain phenomena in a comprehensive study of quads.[12]

In the process of study, several unex-

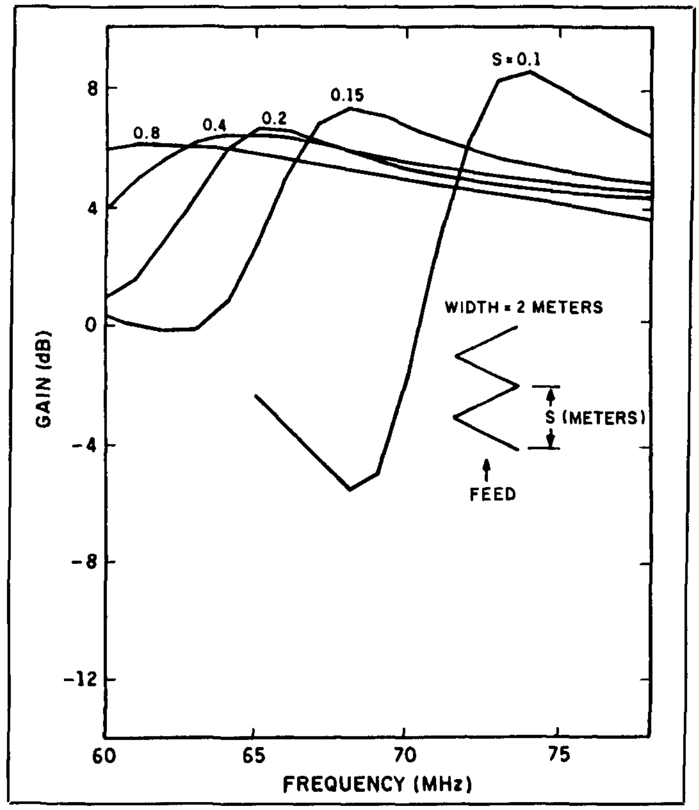

Figure 20. Gain away from the feed point for a zig-zag antenna of several elements and spacings. Note that the span is kept constant, which means that the element length increases as end spacing increases.

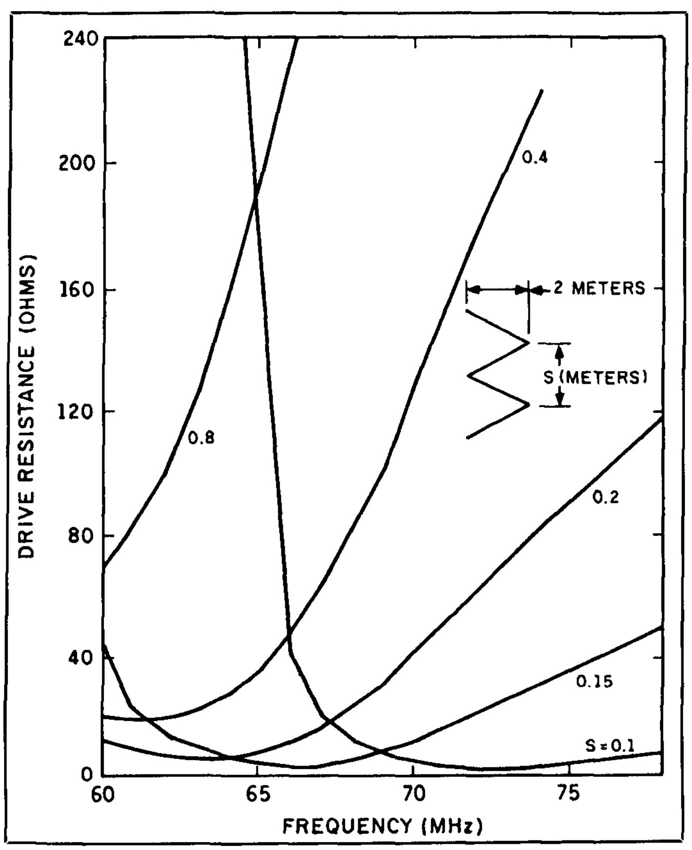

Figure 21. As *Figure 20*, but for the gain away from the feed point. In this direction, the antenna gives good wide-band performance.

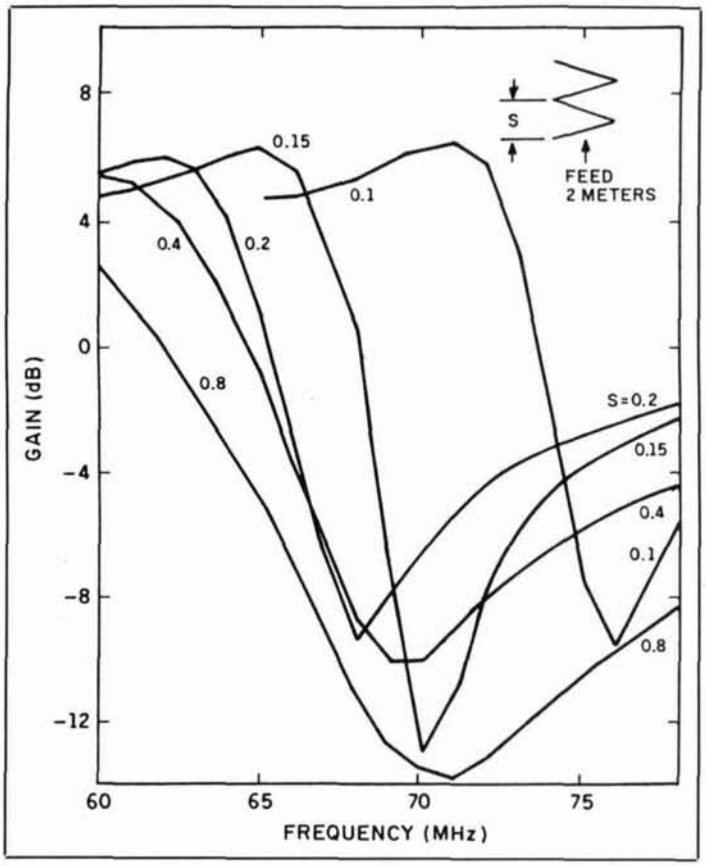

Figure 22. As *Figure 20*, but showing the variation in drive resistance.

pected factors surfaced. One is the degree to which quite common antennas are either in the supergain regime, or are close to it. I had not previously encountered the matter of regarding short dipoles as supergain antennas. Another factor that emerged was the importance of the enclosing sphere in determining whether attempts to achieve supergain were worthwhile. This was particularly noticeable in the consideration of high gain Yagis.

One feature that can only be called "disappointing" was the speed with which conditions became unattractive or even impossible in the practical sense as the supergain regime was reached. The dream of a small high gain and highly directive beam remains "pie in the sky."

On the other hand, some promising avenues for research have appeared. One is the importance of stacking, both as a way of increasing performance while skirting the supergain regime, and in securing increased performance with short boom antennas. Others are the interesting miniaturization possibilities of the open loop quad and the zig-zag antenna family. All of these, and in particular the possibility of antenna neutralization, need more work, both analytical and experimental. ■

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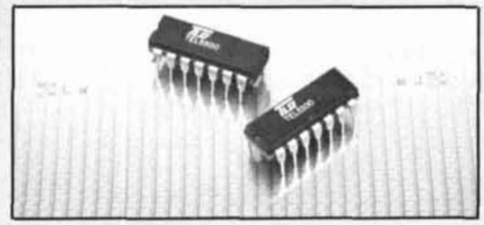
PRODUCT INFORMATION

Programmable ICs for telecom applications

Telephonics subsidiary TLSI, Inc. will market a pair of programmable integrated circuits designed to handle telecommunication line voice and data signals.

The TEL5100 programmable gain/attenuator and the TEL5500 programmable equalizer ICs have application in PBXs, central office equipment, and telephone switches. Both circuits can be used in line cards. The TEL5100 is also suitable for volume controls, automatic gain control (AGC), peak detectors, DSP hardware, and analog multipliers. The TEL5100 and TEL-5500 can be used individually or together, depending on the application.



The TEL5100 and TEL5500 are fabricated using low power CMOS technology. Both devices are offered in 14-pin plastic and ceramic DIP and in 16-pin small outline surface mount packages (SOIC), as well as in die form. Sample quantities are available for evaluation and qualification. MIL-STD-883 screening is available.

For more information contact TLSI, Inc., 815 Broad Hollow Road, Farmingdale, New York, 11735.

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HOW SHORT CAN YOU MAKE A LOADED ANTENNA?

Efficiency levels in inductively shortened antennas

The declining sun spot cycle is (or soon will be) making the higher frequency bands less useful. However, not everyone has the room for a forty or eighty-meter half wave antenna, and you may find your interest turns to short inductively loaded antennas. Here's my report on a study I made to determine the effects of shortening, loading-coil position, and coil Q on:

- inductance required
- efficiency
- signal strength relative to a full-size half wave resonant antenna
- feedpoint resistance at resonance
- SWR to be expected at the input of an arbitrary 50-foot coax line

I'll also discuss some feedline possibilities you can try that should satisfy your SWR-sensitive transceiver.

Assumptions

Because efficiency is a prime consideration, I assumed no. 10 for antenna wire—both to reduce wire loss and, as a consequence of its smaller length-to-diameter ratio as compared with the more common no. 14, to reduce the required inductance for lower coil loss. I chose to study coil Qs of 300 and 100. This represented a reasonable range of achievable Q, which is a function of coil dimensions, wire size, and turn spacing.[1]

For antennas with length-to-diameter ratios of about 2,000 to 20,000, the curve of free space R_r versus L/D is quite flat—within about one ohm of 64 ohms. But this curve ignores the end loading effect of insulators, so I used a nominal value of 65 ohms in calculations. Feedpoint resistance at resonance varies above and below the "free space" feedpoint impedance as a function of height above earth. Feedpoint resistance at resonance of antennas at heights and in surroundings other than those exhibiting the free space value differs somewhat from the calculated values. However, the difference is small in view of the cumulative effects of various approximations common to all antenna calculations, so I assumed free space value for the calculations. Antennas are assumed to be at a height exceeding 0.2 wavelengths (28 feet at 7.15 MHz, for example), where it is generally accepted that earth losses are negligible.

Antenna length and coil position notation

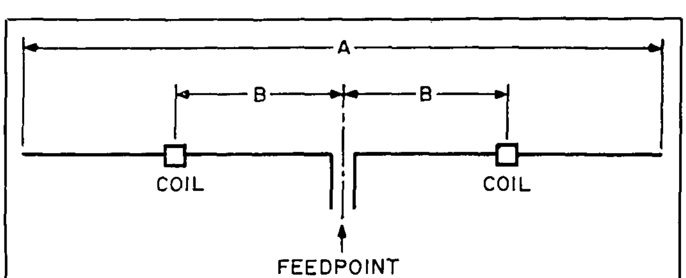
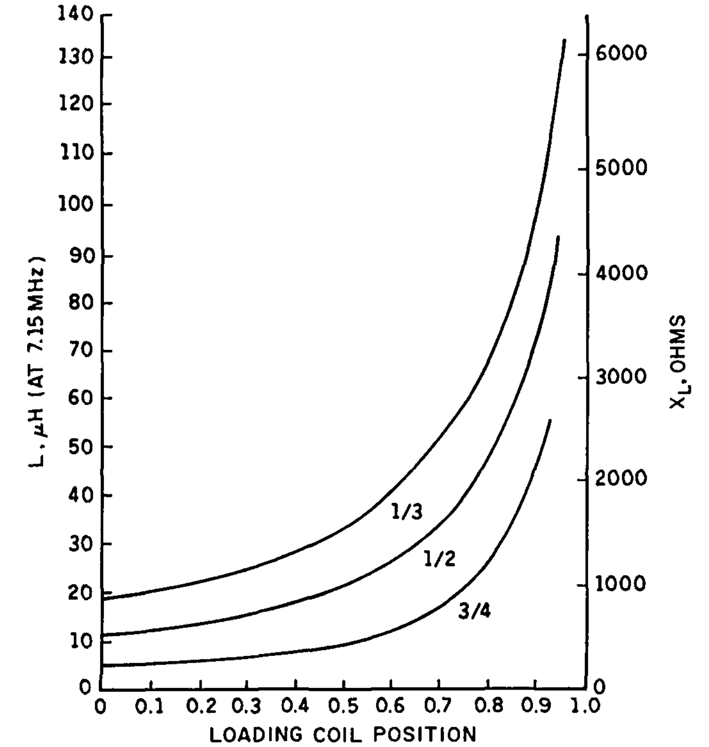


Figure 1. Dimensions relating to shortened dipole.



Position L in μH — **Right scale data**

(3/4 length curve)		(1/2 length curve)		(1/3 length curve)		Left	Right
						μH	X_L, ohms
0	5.43	0.0	11.5	0.0	18.6	0	0
0.25	7.0	0.25	14.7	0.25	23.8	22.3	1000
0.4	8.4	0.4	17.9	0.4	28.8	44.5	2000
0.5	10.0	0.5	21.0	0.5	33.8	66.8	3000
0.6	12.4	0.6	25.8	0.6	41.1	89.0	4000
0.75	20.43	0.75	40.2	0.75	62.2	111.3	5000
0.9	54.6	0.9	93.1	0.9	136.3	133.5	6000

Figure 2. Inductance and reactance of loading coils required (at 7.15 MHz in text example) for dipoles shortened to 3/4, 1/2, and 1/3 times length of full-size *resonant* half wave dipole as a function of coil position (decimal fraction of B/(A/2). If reactance is used (right scale) curves in *Figures 2, 3,* and *4* are frequency independent.

are as shown in **Figure 1**. The horizontal axis in the curve plots (see **Figures 2** through **6**) is the ratio of B (the distance of the coil from the feedpoint) to A/2 (the dipole half length).

I considered shortening to 3/4, 1/2, and 1/3 length relative to a full-size **resonant** half wave dipole. A resonant half-wave antenna is approximately 95 percent of the 180-degree length with a resultant pure resistance feedpoint impedance less than the 73.2 + j42.5 ohms of an infinitely thin 180-degree dipole.

The appendix at the end of this article shows the method I used to calculate the effects discussed here. If you are interested in considering other degrees of shortening, coil positions, or coil Qs, you should find this information helpful.

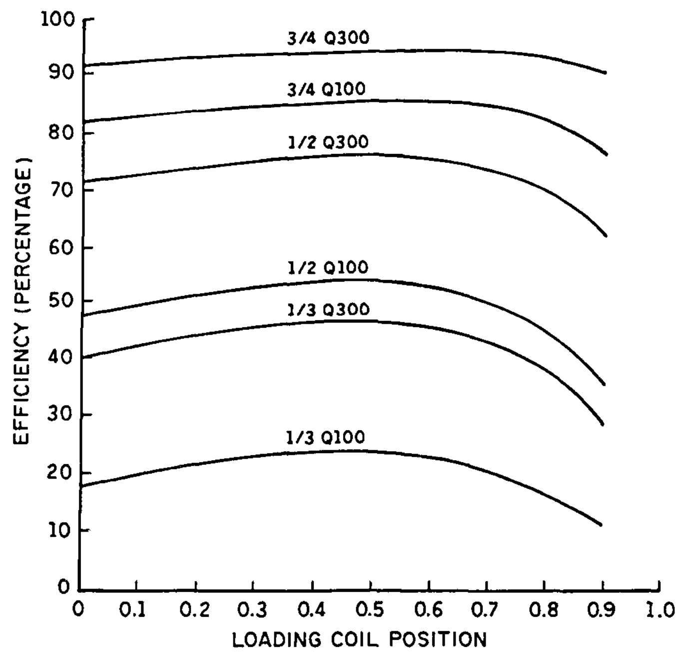
Results

Figure 2 shows that the inductance required increases with the degree of shortening. Depending on the shortening, it increases rather rapidly for coil positions beyond about 0.5 to 0.6.

The right-hand vertical scale indicates the coil reactance required. If this is used to calculate the coil inductance required at other frequencies, the curves of **Figures 3, 4,** and **5** are frequency independent. Due consideration should be given to the problem of attaining a selected Q with high inductance coils, particularly at lower frequencies where the required inductance becomes larger than those indicated for this 7.15-MHz example. Remember that the conventional formulas for coil inductance yield the "low frequency" inductance, and coils will exhibit somewhat higher reactance because of distributed capacity of the coil.

Figure 3 is rather surprising. Efficiency is remarkably independent of coil position over a rather wide range. It is, however, a function of shortening and coil Q. The amount by which efficiency is less than 100 percent represents the power that, except for feedline losses and a slight antenna wire loss, must be dissipated in the two coils.

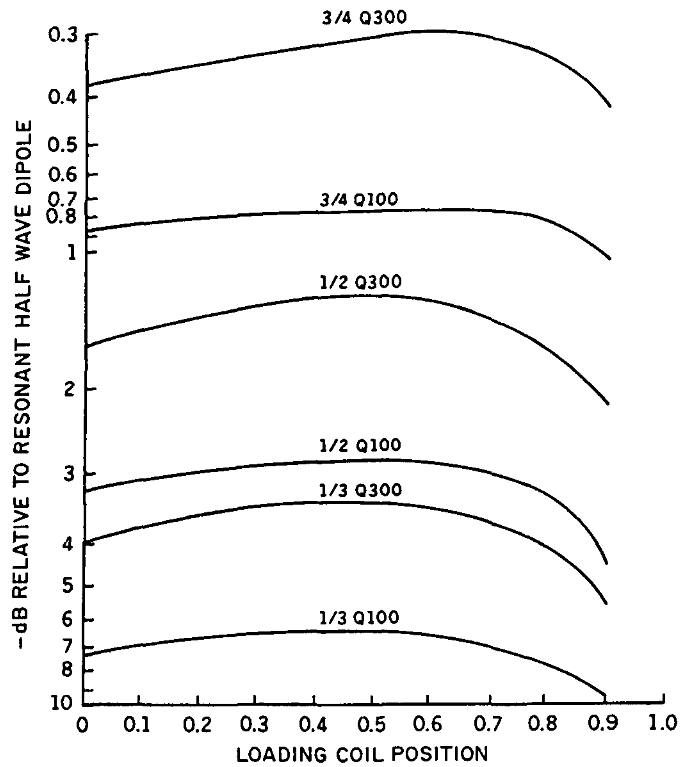
Figure 4 translates the efficiency to terms of effect on signal strength relative to a full-size half wave resonant antenna. As a practical matter, a 3/4-size antenna with the coils as poor as Q of 100 at positions of 0 to 0.9 will be only 1 dB down, or so. A 1/2-size antenna with coils with Q as low as 100 will be down only 3 dB for positions



Position Efficiency (percent)

Position (3/4 length)	Efficiency	Position (1/2 length)	Efficiency	Position (1/3 length)	Efficiency
Q = 300		**Q = 300**		**Q = 300**	
0.0	92	0.0	71	0.0	40
0.25	93	0.25	75	0.25	45
0.4	93	0.4	75	0.4	47
0.5	93	0.5	76	0.5	46
0.6	94	0.6	75	0.6	45
0.75	93	0.75	72	0.75	41
0.9	90	0.9	61	0.9	29
Q = 100		**Q = 100**		**Q = 100**	
0.0	82	0.0	48	0.0	20
0.25	83	0.25	52	0.25	23
0.4	84	0.4	53	0.4	23
0.5	85	0.5	53	0.5	23
0.6	85	0.6	52	0.6	22
0.75	84	0.75	48	0.75	19
0.9	77	0.9	35	0.9	12

Figure 3. Efficiency of shortened antennas relative to full-size half wave resonant dipole as function of coil position.



Position -dB

(3/4 length)		**(1/2 length)**		**(1/3 length)**	
Q = 300		**Q = 300**		**Q = 300**	
0.0	0.36	0.0	1.48	0.0	3.93
0.25	0.33	0.25	1.28	0.25	3.44
0.4	0.31	0.4	1.21	0.4	3.12
0.5	0.30	0.5	1.2	0.5	3.33
0.6	0.29	0.6	1.23	0.6	3.43
0.75	0.31	0.75	1.48	0.75	3.88
0.9	0.44	0.9	2.15	0.9	5.45
Q = 100		**Q = 100**		**Q = 100**	
0.0	0.84	0.0	3.22	0.0	7.1
0.25	0.79	0.25	2.88	0.25	6.47
0.4	0.74	0.4	2.78	0.4	6.33
0.5	0.72	0.5	2.78	0.5	6.36
0.6	0.71	0.6	2.85	0.6	6.2
0.75	0.76	0.75	3.21	0.75	7.18
0.9	1.12	0.9	4.59	0.9	9.26

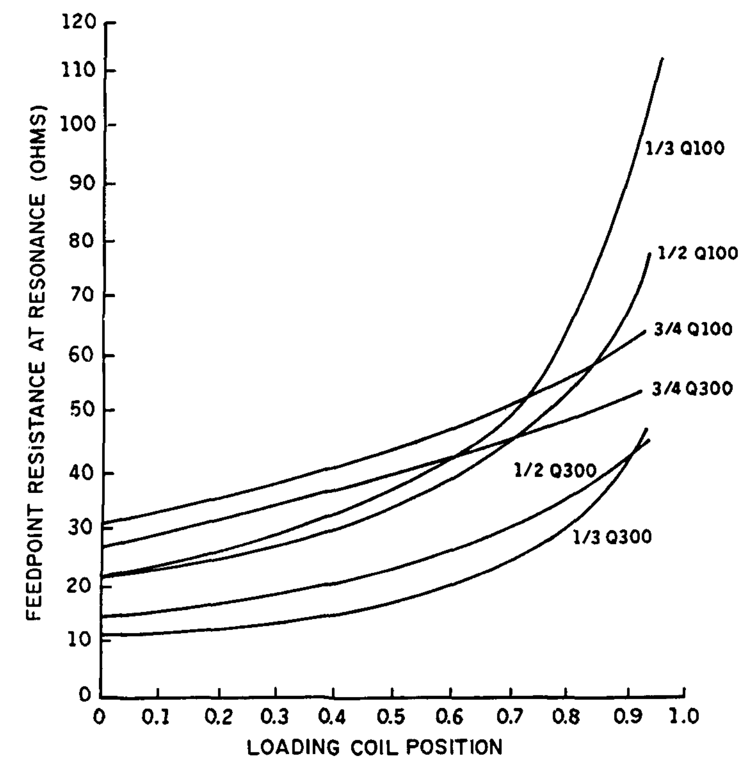
Note: These curves are plotted on semi-log paper with the vertical logarithmic scale increasing downward.

Figure 4. Signal strength of shortened antennas relative to full-size half wave resonant dipole as function of coil position.

from 0.25 to 0.7. Of course, that's half the power delivered to the feedpoint and may present a problem of coil heating depending on the power level involved. Note that with coil Q of 300, shortening to 1/3 length results in only about 3.5 dB loss, while the same shortening with coil Q of 100 results in more than 6 dB loss. In this case, the coils would have to dissipate 75 percent of the power input to the antenna.

Figure 5 is the calculated feedpoint resistance at resonance. It could be useful in determining how to feed the antenna. For cases that depart significantly from the 52-ohm value, you might consider open-wire line, series-section-transformers, or other methods of matching to coax.

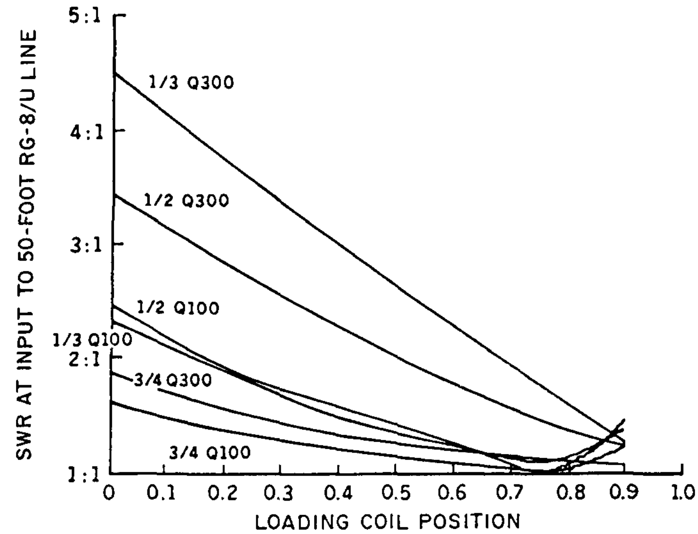
As part of the study, I calculated SWR expected at the input end of an arbitrary 50-foot length of RG-8/U coax (see **Figure 6**). Although not plotted, the difference in line loss at 7.15 MHz compared to that of a perfectly matched line (so called "excess loss") did not exceed 0.4 dB for positions 0 to 0.9 in the shortening cases studied. This difference is considerably less; in fact, it is mostly less than 0.1 dB for all cases nearer the 0.5 position.



Position R_{fp}

(3/4 length)		(1/2 length)		(1/3 length)	
Q = 300		**Q = 300**		**Q = 300**	
0.0	27.8	0.0	14.3	0.0	10.4
0.25	34.85	0.25	19.0	0.25	13.9
0.4	38.5	0.4	22.5	0.4	16.6
0.5	41.7	0.5	25.0	0.5	18.9
0.6	44.5	0.6	27.9	0.6	21.7
0.75	48.0	0.75	33.4	0.75	28.5
0.9	55.1	0.9	45.2	0.9	47.2
Q = 100		**Q = 100**		**Q = 100**	
0.0	31.0	0.0	20.9	0.0	21.5
0.25	38.7	0.25	27.6	0.25	27.9
0.4	43.1	0.4	32.3	0.4	33.2
0.5	45.9	0.5	35.9	0.5	37.9
0.6	49.1	0.6	40.5	0.6	44.2
0.75	54.5	0.75	50.7	0.75	60.6
0.9	64.4	0.9	79.4	0.9	113.5

Figure 5. Feedpoint resistance at resonance.



Position SWR

(3/4 length)		(1/2 length)		(1/3 length)	
Q = 300		**Q = 300**		**Q = 300**	
0.0	1.30	0.0	3.29	0.0	4.37
0.25	1.46	0.25	2.54	0.25	3.38
0.4	1.33	0.4	2.19	0.4	2.85
0.5	1.23	0.5	1.98	0.5	2.57
0.6	1.16	0.6	1.79	0.6	2.26
0.75	1.06	0.75	1.51	0.75	1.75
0.9	1.06	0.9	1.14	0.9	1.09
Q = 100		**Q = 100**		**Q = 100**	
0.0	1.62	0.0	2.34	0.0	2.28
0.25	1.32	0.25	1.82	0.25	1.79
0.4	1.19	0.4	1.56	0.4	1.52
0.5	1.12	0.5	1.41	0.5	1.35
0.6	1.06	0.6	1.27	0.6	1.16
0.75	1.04	0.75	1.02	0.75	1.16
0.9	1.22	0.9	1.49	0.9	2.08

Figure 6. SWR at input end of 50-foot RG-8/U coax used in text example at 7.15 MHz.

When operating off the resonant frequency, SWR and "excess loss" will increase, and inductively loaded antennas have narrower bandwidth than ordinary dipoles. You may need to add a transmatch or other matching system to placate your SWR-sensitive transceiver. On the other hand, open-wire line or twin lead with a balanced matching system driven through a 50-ohm balun is a good solution.

I made one set of calculations for an antenna shortened to 1/5 size with the coils at the 0.5 position. At 7.15 MHz, this required a coil of 55.2 μH and, with Qs of 300 and 100, resulted in an efficiency level of 16 percent and 6 percent with signals relative to a full-size half wave resonant antenna down 8 and 12 dB, respectively.

Conclusions

Efficiency and, hence, signal strength loss is fairly independent of coil position over a surprisingly wide range of possible coil positions for a given shortening. This indicates that it is feasible to choose the coil position to better accommodate mechanical considerations of a particular installation, or to minimize the difficulty of achieving a suitable coil inductance and Q.

In most cases, SWR at resonance won't cause excessive coax line loss on the lower frequency bands. However, inductively loaded antennas have narrower bandwidths than full-size antennas depending on the shortening, coil position, and Q. Consequently, operation off resonant frequency with severely shortened antennas will probably require some matching technique—possibly using open wire or twin lead lines.

Shortening does reduce antenna efficiency. However, shortened antennas can provide suitable alternatives to full-size models. So, how short can you make a loaded antenna? It all depends on the level of signal strength loss you are willing to accept and the power loss your coils can survive. However, loss can be reduced if you use coils with higher Q.

Appendix

$$L_D = (491.8 \times 0.95)/f \quad \textbf{(1)}$$

where:

L_D = approximate length of normal resonant half wave dipole, feet.
f = frequency, MHz

$$A = S \times L_D \quad \textbf{(2)}$$

where:

A = length of loaded dipole, feet
S = shortening factor, (1/2, 1/3, etc...)

$$X_L = 60 \{[Ln(48((A/2) - B)/d) - 1)\tan \Theta_2] - [Ln(48\,B/d) - 1]\tan \Theta_1]\} \quad \textbf{(3)}$$

where:
X_L = reactance of loading coils, ohms
Ln = natural logarithm
A = length of loaded dipole, feet
B = length from feedpoint to coil, feet
d = diameter of antenna wire, inches
Θ_1 = 0.366 f B, angular length from feedpoint to coil, degrees
Θ_2 = 0.366 f ((A/2) – B), angular length from coil to end of antenna, degrees*

*Note that 360/983.6 = 0.366. (The customary 0.95 multiplier for "end effect" has already been incorporated in scaling from the resonant dipole length in **Equation 1** as in the text material.) If **Equation 1** and **Equation 2** are skipped (designed to a specific length for A of **Figure 1**, then use 0.366/0.95 = 0.385 in place of 0.366 in Θ_1 and Θ_2 expressions.

$$L = X_L/(2\,\text{pi}\,f) \quad \textbf{(4)}$$

where:

L = inductance of each loading coil, μH

$$R_c = X_L/Q \quad \textbf{(5)}$$

where:

R_c = AC resistance of coil, ohms
Q = quality factor

$$R_{cfp} = 2\,R_c \cos^2\Theta_1 \quad \textbf{(6)}$$

where:

R_{cfp} = total coil resistance reflected to feedpoint, ohms

$$R_r = 65\,[\sin\Theta_1 + \cos\Theta_1\,((1 - \cos\Theta_2)/\sin\Theta_2)]^2 \quad \textbf{(7)}^2$$

where:

R_2 = radiation resistance of loaded dipole at feedpoint, ohms

$$R_{fp} = R_r + R_{cfp} + R_w \quad \textbf{(8)}$$

where:

R_w = AC resistance of wire reflected to feedpoint, ohms
Value of 0.6 ohms assumed for this study.

$$\text{Eff} = 100\,R_r/R_{fp} \quad \textbf{(9)}$$

where:

Eff = efficiency, percent

$$\text{dB} = 10\,\text{LOG}_{10}\,(\text{Eff}/100) \quad \textbf{(10)}$$

where:

dB = signal strength of loaded antenna referenced to full-size half wave resonant dipole

$$\text{SWR} = R_{fp}/Z_o \quad \textbf{(11)}$$

Or inverse if result is less than 1.
Equation 11 is true only because, in this case (resonance), the input impedance is pure resistance.

where:

SWR = SWR at antenna feedpoint at resonant frequency

Z_o = characteristic impedance of feedline, ohms

SWRI = SWR at input to line (see *The ARRL Antenna Book*, 15th edition, page 24-14.

Procedure used in text material:

Use **Equation 1** to calculate length of normal resonant half wave dipole. Select shortening factor. Use **Equation 2** to calculate length of short antenna, then select distance from feedpoint to coil, B. Enter parameters into **Equation 3** yielding X_L. Use **Equation 4** to calculate L. Select Q that is reasonable to achieve at the chosen frequency and use **Equation 5** to calculate R_C. Refer this to the feedpoint using **Equation 6**. Calculate R_r using **Equation 7**. Select a reasonable wire loss resistance (zero if wire loss is considered negligible). Use **Equation 8** to calculate the feedpoint resistance at resonance. **Equation 9** will yield the efficiency and **Equation 10** the expected signal strength compared to a full-size half wave resonant antenna—not including feed line losses, of course. See the footnote following **Equation 3** for designing to a particular loaded antenna length rather than to a fractional size of a resonant half wave.

Choice of feed method depends on the particular installation, consideration of the feedpoint resistance at resonance, and the bandwidth over which operation is anticipated. ■

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2. Author's extension to the general case of J. Evans, NV1W, derivation for the exactly center-loaded case, based on the method of W.J. Byron, W7DHD, *Ham Radio*, May 1983, pages 36-40 and June 1983, pages 17-20.

PRODUCT INFORMATION

New HP Portable Oscilloscopes

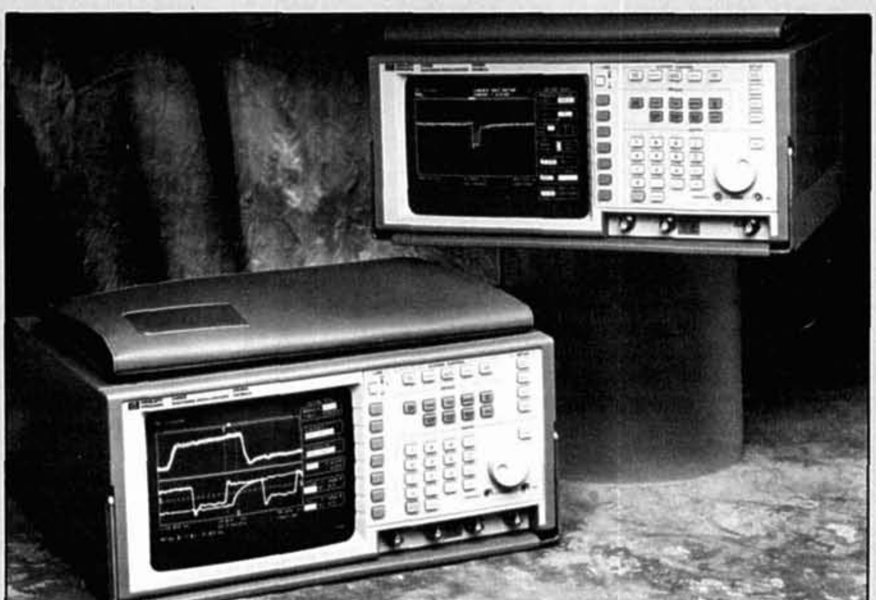
Hewlett Packard offers two new portable oscilloscopes—the two-channel HP 54505B and four-channel HP 54506B.

Both oscilloscopes provide simultaneously sampling at MSa/s without any sample-rate versus channel-count tradeoff. The scopes have 8-K points of memory per channel, 125-MHz single-shot bandwidth, and 330-MHz repetitive bandwidth.

Key features include:

- FFTs for frequency-domain analysis;
- enhanced sequential single-shot for high-acquisition throughput applications;
- advanced logic triggering;
- mask testing with auto-mask generation and mask editing;
- integration and differentiation for power and rate-of-change computations;
- measurement statistics such as maximum, minimum, mean, and most recent;
- TV/video triggering on existing standards and a user-definable mode for nonstandard applications;
- real-time clock for time and date-stamping printouts and waveform memories.

For more information on either of these products contact: Hewlett-Packard Company Direct Marketing Organization, P.O. Box 58059 MS51L-SJ, Santa Clara, California 95051-8059.



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HF RECEIVER DESIGN

Tracing the development of the high frequency communications receiver

This article traces the evolution of the high frequency communications receiver from the earliest TRF design, through the development of the multiple conversion superhet. It describes, in some detail, the evolution of the modern fully synthesized high performance communications receiver.

Over the years, many changes have taken place in the basic configuration of the communications receiver. I'll discuss the reasons behind this continued growth and also the revolutionary changes in design philosophy that have occurred in the last two decades. These changes followed the introduction of integrated circuits, digital techniques, and microprocessor control.

Basic requirements of a communications receiver

Some of the more significant fundamental requirements for a communications receiver include:

- The receiver must be tuneable across a wide range of frequencies.
- The receiver must be capable of being set quickly and accurately to any required frequency. The frequency display must be of adequate resolution and accuracy.
- The receiver must be capable of demodulating all required transmission modes (AM, CW, SSB—USB and LSB—and possibly ISB, RTTY, FM, etc.).
- The receiver must be sensitive, to enable reception of very weak signals.
- The receiver must be selective enough to receive the wanted signal, while rejecting unwanted adjacent signals.
- A number of different selectivity filters will be required for different transmission modes and propagation conditions.
- The receiver must provide adequate re-

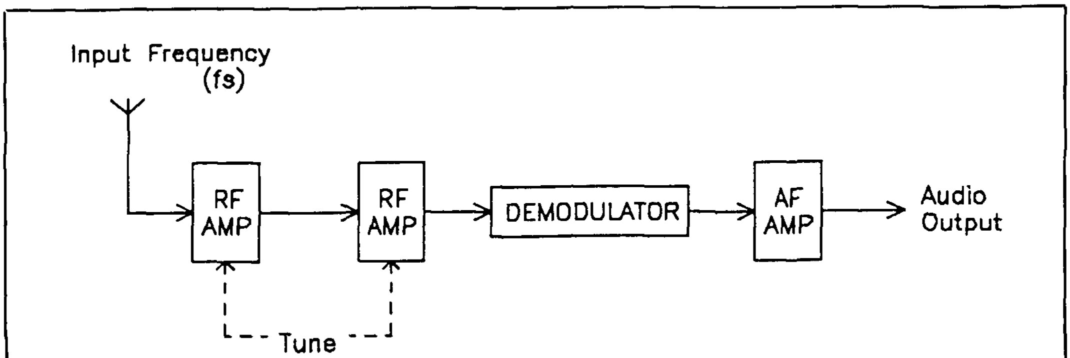

Figure 1. The TRF receiver.

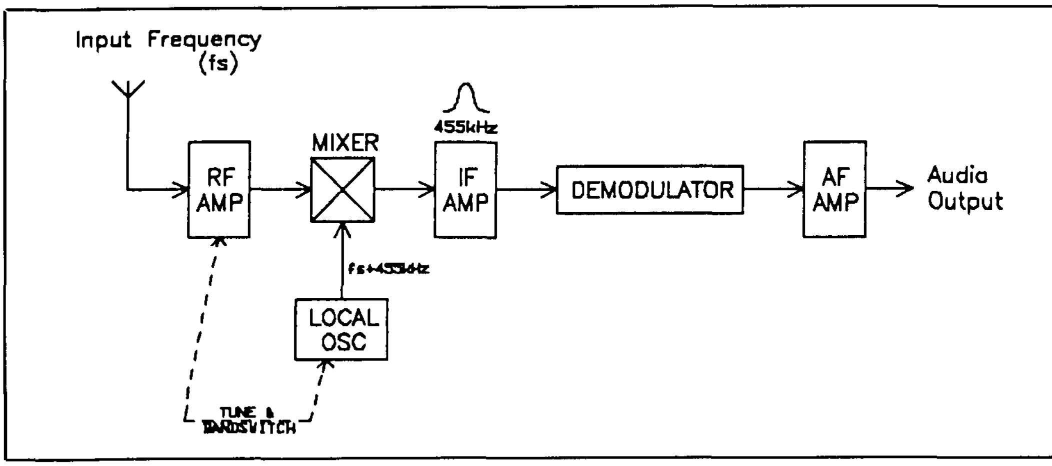

Figure 2. Single-conversion superhet receiver.

jection of all unwanted external spurious signals. This includes image frequency rejection and IF rejection.

- The receiver must have high performance with regard to dynamic interference effects like intermodulation, cross modulation, blocking, reciprocal mixing, and so on.
- All internal noise produced by the receiver must be at an acceptably low level (e.g., below the noise floor). Synthesizer phase noise must be at low levels for good reciprocal mixing performance.
- The receiver must possess good frequency stability over a wide temperature range.
- The receiver must be ergonomically designed with regard to the layout of controls, and must be simple and straightforward to operate.
- The receiver must have high reliability over the complete range of environmental conditions for which it is designed.
- The receiver must be easy to maintain. Diagnostics using Built-In-Self-Test (BIST), Built-In-Test-Equipment (BITE), and easily changeable Line Replaceable Modules (LRUs), may be a requirement.
- Computer-controlled remote receiver operation may also be a requirement.

Basic principles—an historical overview

The TRF receiver. One of the simplest forms of receiver, dating back to the very origins of radio (and hence of electronics), is the "straight" or Tuned Radio Frequency (TRF) receiver. This type of receiver consists of a tuned radio frequency (RF) amplifier, a demodulator, and an audio amplifier (see **Figure 1**).

The RF amplifier could be fixed-tuned in a single channel receiver, but was more often variable-tuned to allow for reception of stations across a band of signals. The selectivity achieved by this arrangement depends solely on the selectivity (sharp tuning, or high "Q") provided by the RF amplifier and, in practice, often left a lot to be desired. While it's true that RF amplifiers can be made quite selective, especially at frequencies below about 1 MHz, the task becomes more difficult at higher frequencies—even with multi-stage designs employing positive feedback to increase "Q." (Who remembers trying to reduce interference on the old "regenerative set" by turning up the feedback until it squealed, then backing off a bit? It certainly wasn't very user friendly!)

Because of these selectivity problems, the tuneable TRF is of very limited use as a communications receiver, as it is virtually impossible to make the tuned RF stages sufficiently selective across a wide range of input frequencies. (It is interesting to note, however, that it's quite possible to design very simple single channel low frequency TRF receivers with very high performance. These receivers never drift, and never suffer from oscillator/synthesizer noise! An example dear to the heart of any ex-Merchant Navy radio officer is the "reserve receiver" used in the ship's radio room to monitor the international distress frequency of 500 kHz.)

The superhet receiver. The superheterodyne receiver evolved out of a need for a tuneable receiver with adequate and consistent selectivity. In the superhet, the incoming frequency is first converted (heterodyned) to a single fixed frequency, known as the intermediate frequency (IF). At this fixed frequency, the required selectivity could be obtained easily with two or more stages of amplification using conventional tuned circuits, provided the IF chosen was

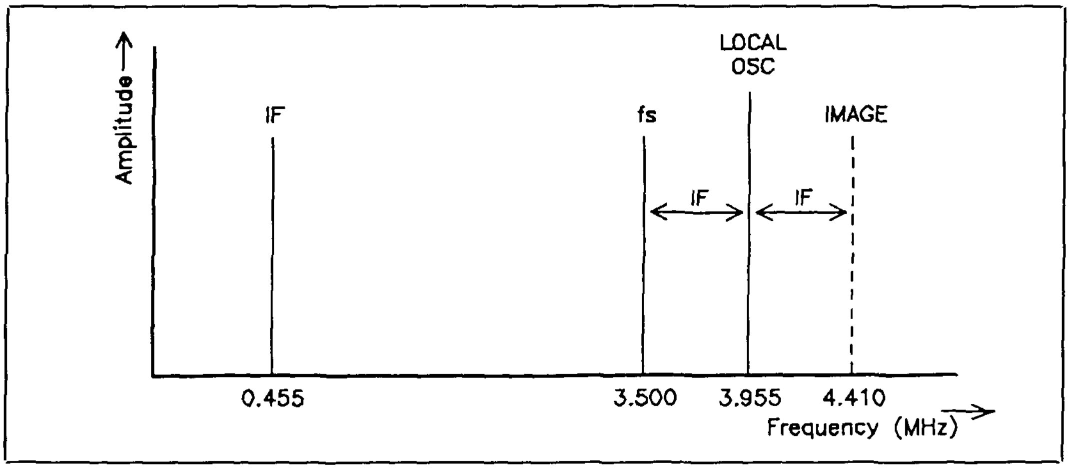

Figure 3. Spectrum of receiver signals.

low enough. Therefore the IF selected was always less than 1 MHz, and was commonly in the range 450 to 470 kHz. A frequency of 100 kHz (or even 85 kHz) was sometimes used to achieve extra selectivity in communications receivers—especially for the reception of CW signals.

In its simplest form, the frequency conversion is achieved using an oscillator and a mixer (see **Figure 2**). In order to "tune in" to the required station, the oscillator (usually known as the local oscillator) is adjusted (by turning the main tuning control) so its frequency (f_{OSC}) will mix (heterodyne/modulate) with the incoming frequency (f_s), producing a frequency product equal to the IF. (The mixer is basically a nonlinear device and, in fact, produces a large number of frequency products other than the required product—including $f_{OSC} - f_s$, $f_{OSC} + f_s$, $2.f_{OSC} - f_s$, etc. The selectivity of the IF amplifier will reject all the unwanted products, allowing only the required product to pass.)

This required product at the intermediate frequency is equal to the difference between the signal frequency (f_s) and the oscillator frequency (f_{OSC}). Putting this mathematically:

$$IF = f_{OSC} - f_s \textit{ or } f_s - f_{OSC} \quad \textbf{(1)}$$

$$\text{thus } f_{OSC} = f_s \pm IF \quad \textbf{(2)}$$

All this can best be seen using an example. Referring to the RF spectrum drawing in **Figure 3**, you'll note that to tune in a wanted signal (f_s) of 3.5 MHz for an IF of 455 kHz, the local oscillator frequency (f_{OSC}) must be set to 3.5 MHz plus or minus the IF of 455 kHz. In practice the plus side is normally used, thus the oscillator will be set to 3.955 MHz. This mixes (or beats/heterodynes) with the 3.5-MHz signal in the mixer, and one of the resulting products at 455 kHz ($f_{OSC} - f_s$)is picked out by the IF selectivity.

All this seems fairly satisfactory, however careful analysis of the **Figure 3** reveals a new problem. You can see that not only will the required signal at 3.500 MHz beat with

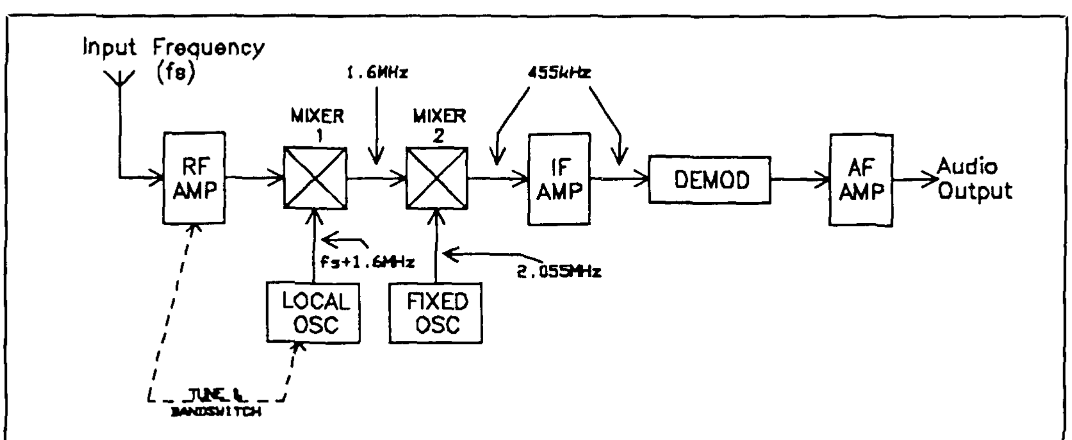

Figure 4. Double-conversion superhet, variable first LO.

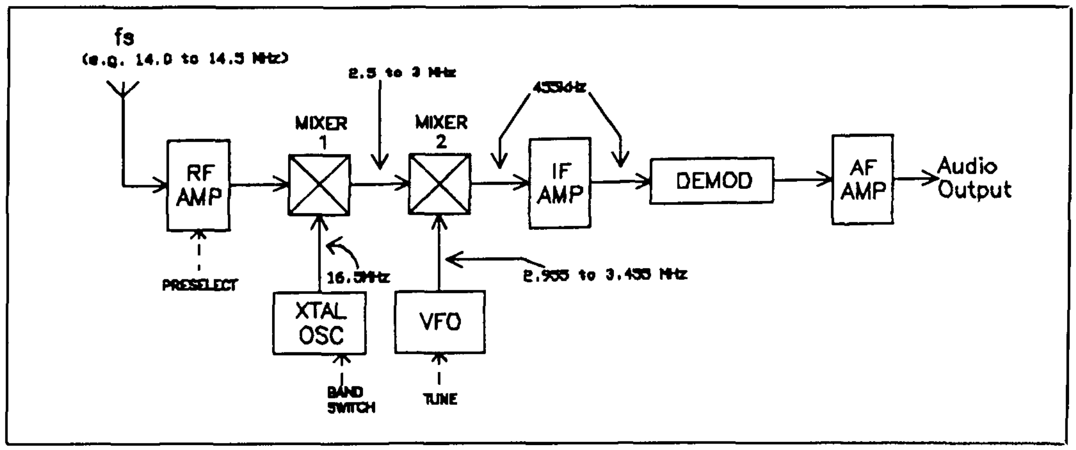

Figure 5. Double superhet receiver, variable second LO.

f_{osc} to produce the IF of 455 kHz, but another (unwanted) frequency at 4.410 MHz will also beat with f_{osc} to yield the IF, and would therefore produce interference. This 4.410 MHz frequency is called the image frequency (it's a mirror image of f_s around f_{osc}—see **Figure 3**), and is always removed from the wanted frequency by an amount equal to twice the IF, as shown in **Equation 3**:

$$f_{IMAGE} = f_s \pm 2\ IF \qquad (3)$$

(Generally, as in our case, $f_{IMAGE} = f_s + 2\ IF$)

The solution to this problem is theoretically fairly straightforward: the image frequency must be rejected, thus we are back to using a tuned RF amplifier. This tuning need not be nearly as tight as for the TRF receiver, as the main selectivity for the receiver will be provided by the IF amplifier. Nevertheless, adequate performance is not easy to achieve with the use of a low IF. This is, quite simply, because the higher the IF, the further away will be the image, and the easier it will be to reject. (Note that some degree of selectivity up front also has the important advantage of improving the receiver front-end performance.)

We now have conflicting requirements. We need a low IF for selectivity, and a high IF for rejection of the image frequency. The single conversion superhet receiver outlined here employs a compromise IF, usually of 455 kHz or 1.6 MHz, but the real solution lies in the double conversion receiver (more on this to come). Note also that for a tuneable receiver, the RF tuning must "track" the oscillator tuning even though the frequencies are different—hence the name "tracking superhet" sometimes used for this type of receiver.

The use of a local oscillator in the superhet introduces an extra problem of oscillator drift. For the reception of AM signals this is not normally significant, as the oscillator has to drift a long way to move the signal out of the passband. This is not the case for CW, where filters of 300 Hz or even 100 Hz may be used. Even less drift can be tolerated on SSB. An SSB signal must be "on tune" to within 80 Hz, or less, to be intelligible. In a tuneable single conversion receiver, the oscillator must cover a large range, and also must be bandswitched. This makes the required level of stability difficult, if not impossible, to achieve.

(As an aside, it's interesting to note that SSB was pioneered by amateur radio enthusiasts in the late 60s. However, due to problems obtaining the required stability, it was another decade before SSB became universal in the professional radio world—after the development of frequency synthesis.)

The Double Conversion Receiver. As was already mentioned, there is a conflicting requirement of a high IF for image frequency rejection, and a low IF for selectivity. This led to the development of the double conversion superhet receiver, which had two frequency conversions, and two intermediate frequencies. The first IF was at high frequency for good image rejection; the second IF was at a low frequency to provide good selectivity. Thus, we have the best of all worlds!

In a double conversion receiver, various local oscillator arrangements are possible. In one double superhet configuration the first oscillator was made variable and the second fixed, as shown in **Figure 4**. Note that this configuration inherits all the problems regarding frequency drift outlined for the single superhet.

The alternative configuration shown in

Figure 5 uses a fixed first local oscillator (usually crystal controlled) with a variable frequency oscillator (VFO) as the second oscillator. The first oscillator can have switchable crystals to cover different bands. In addition, the VFO can be made more stable than the variable oscillator in the first arrangement because it can have a limited range, with no bandswitching problems. This means that a separate control, usually known as the "preselector" control, must be used for RF tuning. It also means that the first IF is wideband, which could compromise certain aspects of receiver performance (more on this later).

Nevertheless, this arrangement has been the basis of many successful limited coverage receivers, like amateur band only or marine band only receivers. Note that the arrangement is equivalent to a single superhet with a fixed frequency converter up front. (In fact receivers of doubtful performance can often be given a new lease on life by literally connecting external frequency converter(s) up front to improve sensitivity or provide different frequency ranges.)

Premixing is a variation on this theme. With this technique, the output of a crystal oscillator is premixed with a VFO output before being applied to the signal mixer. The premixing technique can be used on a double superhet receiver, as shown in **Figure 6**, or can be used to produce a single superhet design with stability similar to the double superhet of **Figure 5**. Note that in both cases the wideband IF of **Figure 5** is not required. The bandpass filter is necessary to remove unwanted products of the premixing (Mixer 3) process.

However, suppose unbroken coverage of the complete HF spectrum is required. Or perhaps you need coverage of the VLF, LF, MF, *and* HF spectrum; that is, coverage from virtually DC to 30 MHz! This was once achieved by complex switching of a multi-conversion receiver. On some bands it would be single superhet, some double, and some even triple—with tuning scales for each band of different resolution. There was even a reversal of scale direction on some bands!

Nowadays, the solution is up conversion to a first IF higher than the highest frequency being received. Thus for an HF receiver, a VHF first IF is used (usually in the range of 40 to 90 MHz or more). This enables unbroken coverage and has the advantage not only of putting the image frequency very far away (for a 10-MHz signal, an IF of 48 MHz means an image is at 106 MHz!), but also ensures that it can be easily rejected by means of a 35-MHz low-pass filter (see **Figure 7**).

We are still left with a wideband first IF, and also with the problem of how to generate the first local oscillator signal. Banks of crystal oscillators were used originally. This is an expensive technique and involves a lot of switching. Also for an IF of 48 MHz and a signal of 29 MHz, the local oscillator will be running at 77 MHz, and will be none too stable.

This leads us to the idea of a device that can synthetically produce a number of frequencies (hence the name frequency synthesizer) with the same stability as that of a single crystal-controlled reference oscillator.

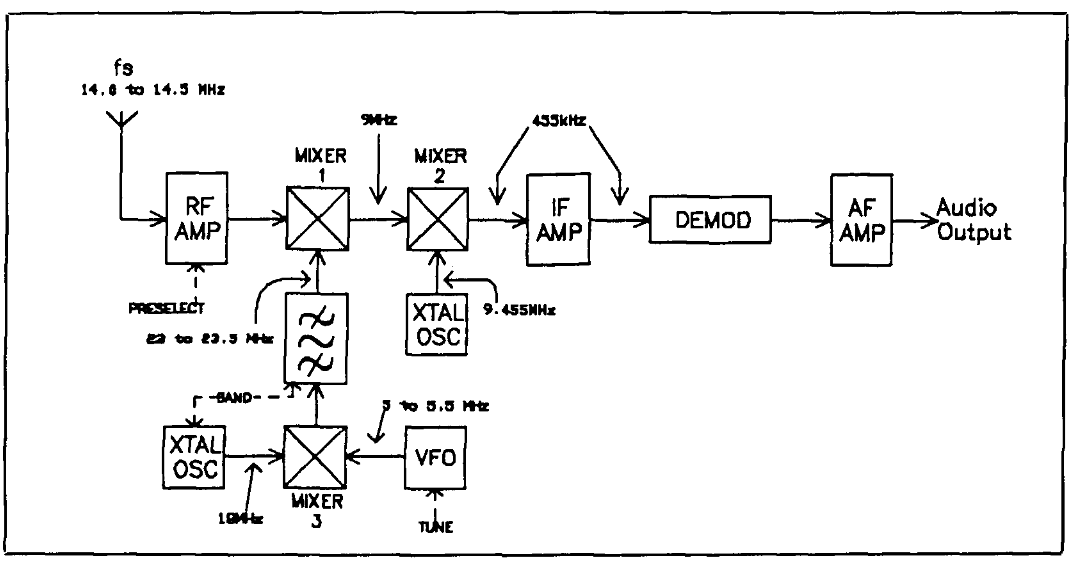

Figure 6. Premixer technique.

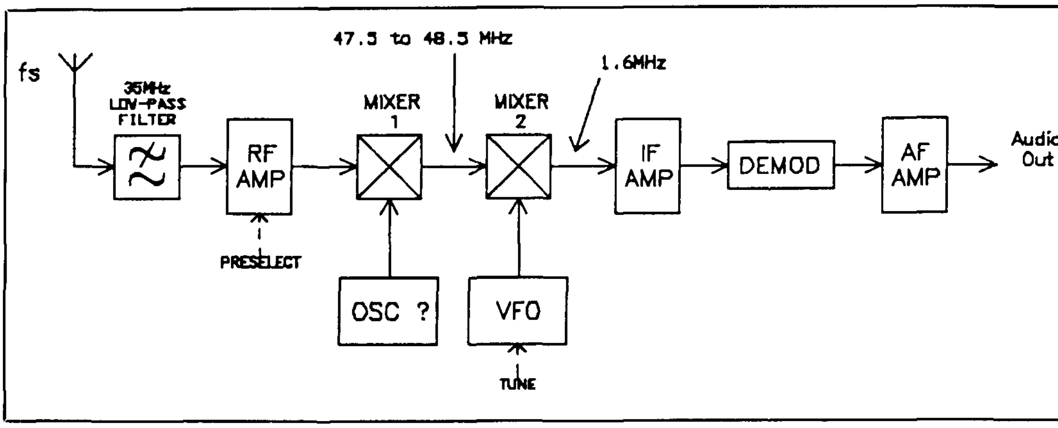

Figure 7. Basic up-conversion superhet.

The latter can be made very stable by suitable choice of crystal. Professional receivers use temperature-controlled crystal oscillators of 1 in $10^8/°C$ stability, giving 0.1 Hz drift at 10 MHz! Even an unovened crystal gives 1 in $10^6/°C$, which is 10 Hz at 10 MHz.

The synthesizer generates a discrete set of frequencies and doesn't tune continuously like a variable oscillator. Thus an HF receiver can either use a small number of big steps (e.g., 30 steps of 1 MHz used like a bandswitch, with VFO to tune across each 1 MHz band; that is, partial synthesis), or it can use a large number of small steps (e.g., 10 Hz, or 1 Hz)—and by dispensing with the VFO can achieve very high stability (full synthesis).

A form of partial synthesis was first employed on the Racal RA17 receiver using an error correcting loop technique called "Wadley-loop triple mix." The MHz oscillator has an output of 40.5 to 69.5 MHz in 1-MHz steps, and bandswitching is achieved by roughly setting the MHz oscillator using the bandswitch control. Any error or subsequent drift in the frequency of this oscillator will be reflected as a shift at the first IF. However, the oscillator output is also mixed with a comb of frequencies of 1-MHz spacing derived from an accurate 1-MHz reference source, and the relevant product at near 37.5 MHz is used as an injection frequency for the second mixer. Thus the frequency shift in the first IF is exactly canceled out in the second mixer, as described in the next paragraph. Tuning of the 1-MHz band is by the VFO feeding into Mixer 3 (see **Figure 8**).

As an example, consider an input signal (f_s) of 15.260 MHz. The MHz oscillator would be set to about 55.5 MHz to set the receiver to the band 15 to 16 MHz. With our 15.260-MHz signal, the first IF will be (about) $55.5 - 15.260 = 40.240$ MHz. The 55.5-MHz oscillator signal will also mix in Mixer 4 with our 1-MHz comb of frequen-

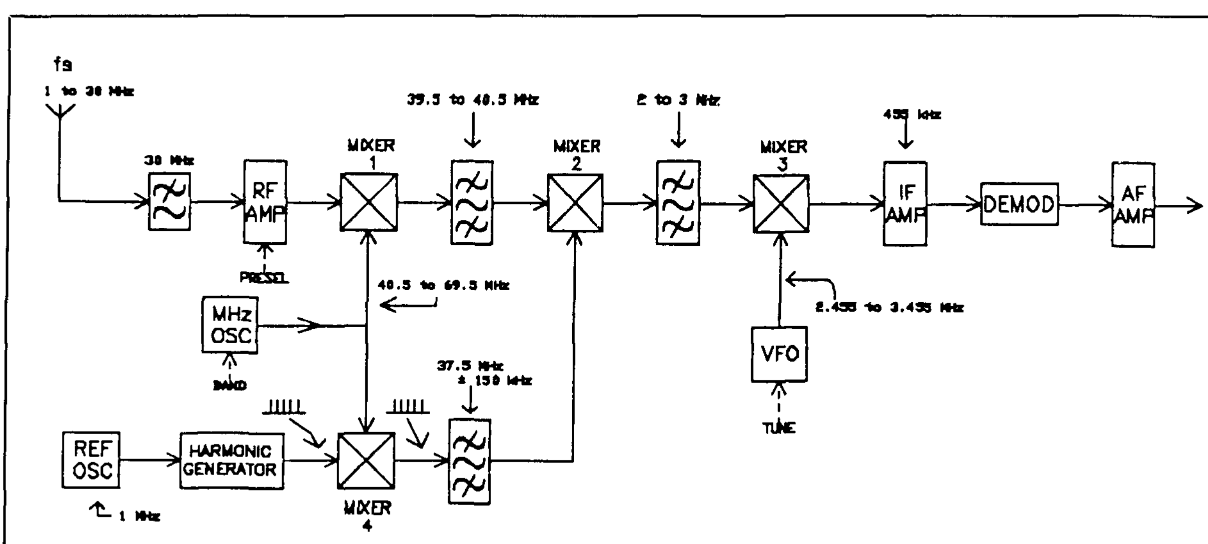

Figure 8. Wadley loop receiver—Racal RA17.

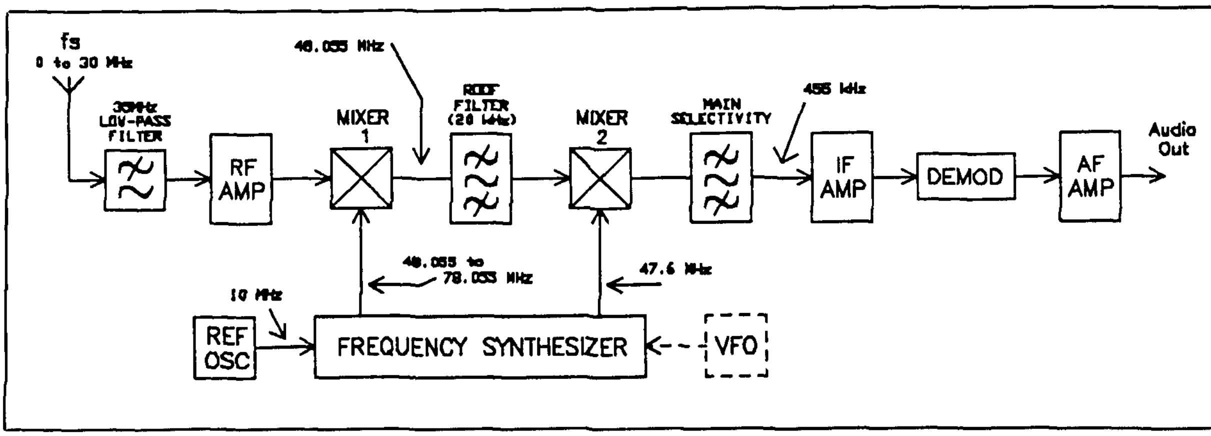

Figure 9. Up-conversion receiver, preferred configuration.

cies. In particular, the 18-MHz "tooth" of the comb will mix with the 55.5 MHz to produce a 37.5-MHz product (55.5 – 18.0) that will pass through the 37.5-MHz bandpass filter to form the injection frequency for Mixer 2. This mixer will thus translate the first IF signal (40.240 MHz) into a second IF of 40.240 – 37.500 = 2.740 MHz. (The VFO will be at 2.740 + 0.455 = 3.195 MHz to yield a third IF of 455 kHz.)

Now suppose the MHz oscillator were set, or were to subsequently drift, say, 20 kHz high. This would cause the first IF to be at 55.52 – 15.260 = 40.26 MHz; that is 20 kHz high. However, the output from Mixer 4 will also be 20 kHz high (55.52 – 18.0 = 37.52 MHz) and the second IF will be 40.26 – 37.52 = 2.740 MHz, which is exactly as it was before! The drift of the MHz oscillator has indeed been exactly canceled out in the second mixer. The 1-MHz reference oscillator is a high-stability temperature-controlled crystal oscillator to maintain accuracy and stability.

The RA17 represented a great step forward in its time (the early 50s), with a tuning scale that was five feet wide for each of its 1 MHz wide bands, achieved by spooling a 35-mm film-strip across the tuning window! The receiver had near constant tuning on each band, and rock-like stability. However, the first two IFs were both wideband (1 MHz wide), and this had an effect on cross modulation and intermodulation performance. This was especially true in later versions of the receiver (RA217, RA1217) that used bipolar transistors in place of vacuum tubes (valves for U.K. readers). (Note that all these performance problems were fully resolved by Racal during the development of the RA1771/1772 receivers. See **References 1** and **2** for full details.)

Cross modulation and intermodulation

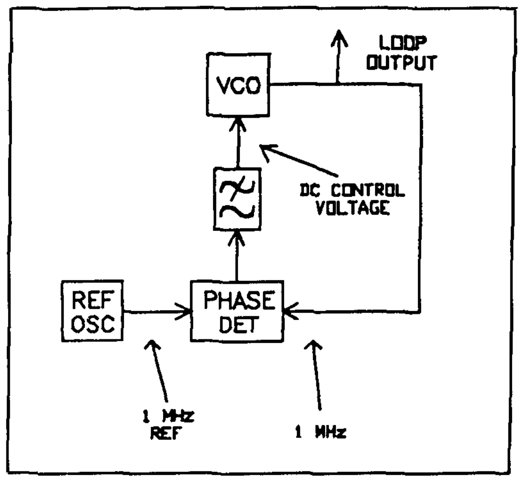

Figure 10. Basic PLL.

occur due to nonlinearities in any of the receiver stages (especially mixers) up to the first really selective stage. Thus for maximum performance, we want a first conversion stage to feed a narrow-band first IF; that is, to introduce a fairly sharp filter (called a roofing filter) at the first (VHF) IF. To achieve this we need either a full frequency synthesizer; or a partial frequency synthesizer, with a VFO feeding into the synthesizer and not the second mixer. The synthesis process must be completely separate from the signal path—not intertwined as in the Wadley loop.

These requirements set the stage for what has become the definitive configuration for the modern communications receiver. It consists of a synthesized dual-conversion superhet employing up conversion to a VHF first IF, with a roofing filter immediately following the first mixer, and with the main selectivity filters at the second IF (see **Figure 9**). The remainder of the article examines this "preferred" configuration in greater detail, and describes the radical

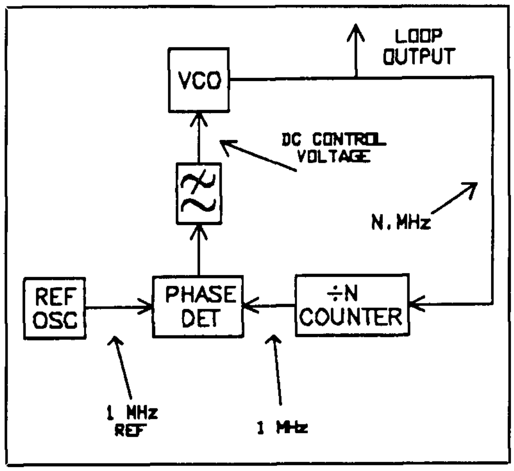

Figure 11. PLL with programmable divider.

changes brought about by the advent of the digital revolution in electronics.

The digital revolution

Partial synthesis. Initially, all sorts of techniques were used for frequency synthesis. The premixer and Wadley loop partial synthesizers are just two examples of the wide proliferation of synthesis methods—some of which used complex analog circuitry (filters, mixers, etc.) that were very expensive and difficult to align. After many years, one method began to reign supreme as the universal choice for receiver designs—the phase lock loop.

A phase lock loop (PLL) works as follows. A voltage controlled oscillator (VCO) provides the output frequency from the loop. This frequency is also fed back to a phase detector where it is compared with an accurately controlled reference frequency. The phase detector output is a DC voltage proportional to the frequency difference (or actually the phase difference) of the two inputs. This DC control voltage is fed back to the VCO through a low-pass filter. Any error in the VCO output will result in the generation of a DC control voltage that corrects the error, thus (given enough loop gain) the loop will be locked (in frequency and phase) to the input reference (see **Figure 10**).

To summarize, the VCO is forced by its DC control voltage input to run in such a way that frequency fed back to a phase detector is the same as the accurate reference frequency also fed to the phase detector. At this point one might ask, "What's the point, why not use the reference output direct and dispense with the phase lock loop?" Well, what really makes the PLL useful in this application is

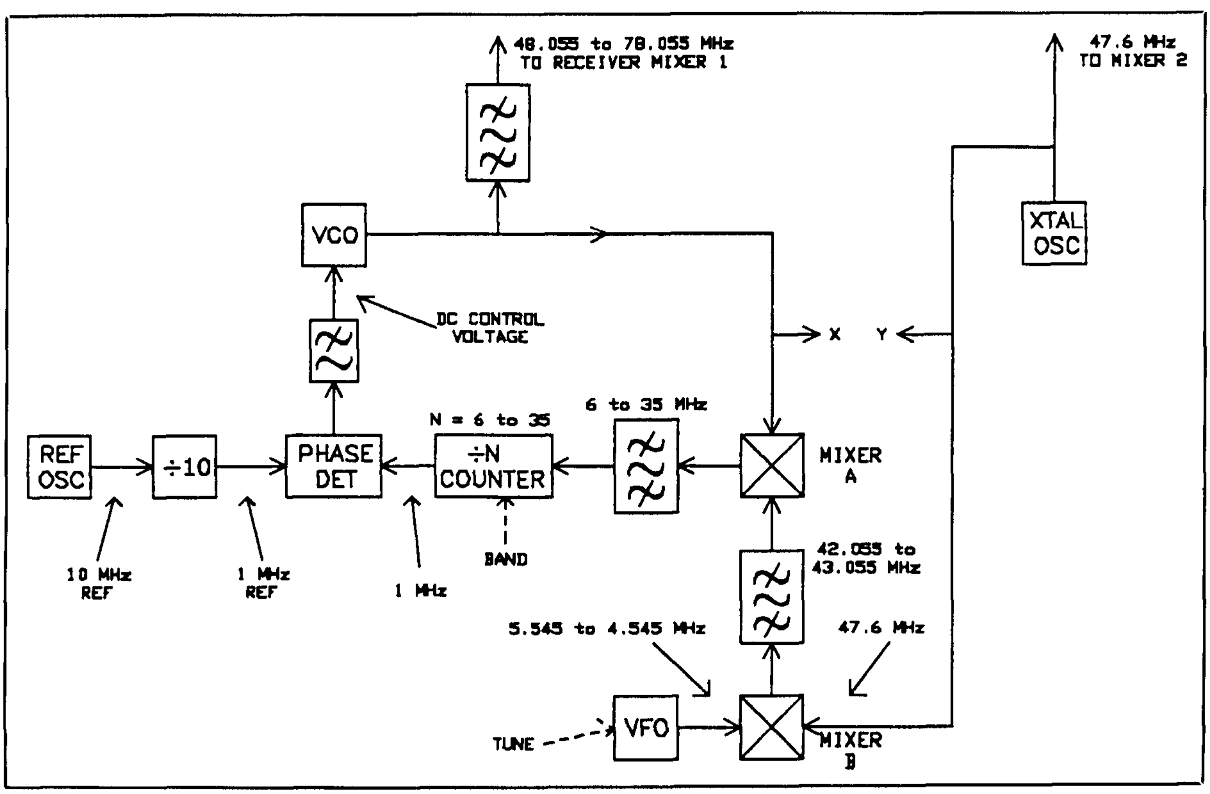

Figure 12. Phase locked partial frequency synthesizer.

the inclusion of a programmable frequency divider into the loop (see **Figure 11**). By varying the division ratio (N) of this frequency divider, we can vary the loop output frequency. For example, if we set N in **Figure 11** to be 20, then the output frequency will be 20 MHz. If we change N to 21, we get 21 MHz! Thus by incorporating a programmable frequency divider into the loop, a discrete set of frequencies can be generated by altering the division ratio of the divider (in this case 1 to 30 MHz), all of which are frequency locked to the high stability frequency reference.

What's more, if a mixer is also included in the loop, a continuously variable frequency—like that from a VFO, or other discrete frequencies (of smaller step size) from other phase lock loops—can also be fed into the loop.

For an example of a partial synthesizer for a low-cost general-coverage double-conversion receiver introduced in the early 80s (the Kenwood Trio R1000) see **Figure 12** (and also **Figure 9**). This synthesizer uses a single PLL in a very simple and straightforward way, and is typical of a single loop synthesizer with a "mixed-in" VFO. Thirty 1-MHz wide bands are provided by switching the division ratio of the ÷N programmable counter. Tuning across each band is accomplished by feeding the output of a VFO via Mixer B into the loop mixer, Mixer A. (In practice the single VCO shown does not cover the range 48.055 to 78.055 MHz, so a bank of four VCOs are used—automatically switched in as appropriate.)

As a further refinement in this design, the receiver second mixer injection frequency of 47.6 MHz, which is generated by a crystal oscillator (and which may thus be prone to significant drift), is also fed into the loop via Mixer B. Any drift of, say, 1 kHz high will cause the output of Mixer B to go up by 1 kHz. This will cause the VCO to increase by 1 kHz to keep the output of Mixer A the same. (Don't forget the loop forces the output of Mixer A—divided by N—to be exactly equal to the 1-MHz reference.) If the VCO goes up 1 kHz, the receiver first IF will also go up by 1 kHz; thus the original 1-kHz rise of the 47.6-MHz oscillator will be exactly canceled out in the receiver second mixer. That is, as well as the main (phase locked) loop, we also have a secondary error-correcting loop involving both mixers and the first IF (in a similar way as the Wadley loop), which corrects for any drift in the 47.6-MHz local oscillator. (The points marked X and Y are discussed later—see **Figure 15**.)

The first PLL receivers used all analog circuitry, but very soon the advantages of using digital techniques for at least the (quite complex) frequency dividers became apparent. The resultant circuits are smaller, cheaper, give higher performance, and are easier to align. Thus the digital revolution had started, and with the availability of TTL and ECL (later CMOS) integrated circuits, became a fact of life. Soon other parts of the circuit were transformed from analog to digital, for example, the phase detector; and eventually single chip PLLs were produced.

Full synthesis. Partial synthesis goes a long way towards providing a stable, accurate, and simple-to-use receiver; but we still have a VFO—albeit a reasonably stable one. To improve stability still further (needed for some exotic transmission modes like Piccolo and Kineplex), and to make possible full remote control of receivers in "point-to-point" applications, methods of full synthesis were developed. Again, many methods were tried initially, including direct synthesis using banks of oscillators and mixers, but soon the ubiquitous PLL became the method of choice. In a typical fully synthesized receiver, the VFO was replaced by another phase locked loop (or a number of loops) in which synthesis down to steps of 100 Hz, 10 Hz, 1 Hz, or even 0.1 Hz, could be achieved. The entire frequency generation system in the receiver could now be locked to the single frequency reference oscillator, which was normally oven controlled. Stabilities of better than 0.1 Hz could thus be achieved over the entire frequency range (and temperature range) of the receiver. Alternatively, many receivers in a communications station could use a single very high accuracy frequency reference (such as a rubidium or atomic standard) piped to all receivers.

The first generation of full-synthesis receivers used a row of decade switches to set the desired frequency. While this may be suitable for point-to-point type applications, it is very inconvenient for a general-purpose "search" type receiver. As a result, the "digital VFO" concept was developed in which the operator can tune the receiver just as if he were tuning a VFO. In fact, however, the tuning knob is connected to an optical shaft-encoder, which generates pulses to drive a digital counter up or down. It's the output of this counter (usually in BCD form) that programs the ÷N counters, and thus sets the receiver frequency. Provided the circuit is fast enough and steps are small enough (10 Hz or less), a good approximation to true continuous tuning is achieved.

The arrangement shown in **Figure 13** is a hypothetical extension of the R1000 circuit to incorporate full synthesis, and is typical of what was done in many "free-tune" receivers. A total of four phase lock loops are used—three to generate the main (stepped) 48.055 to 78.05499-MHz output, with a fourth around the 47.6-MHz variable crystal oscillator (VXO). The 1-MHz wide bandswitching is done as before in the top loop, with two more loops providing coverage in 10-Hz steps. The ÷10 stage in the output of the bottom loop enables a reference frequency of 100 Hz rather than 10 Hz

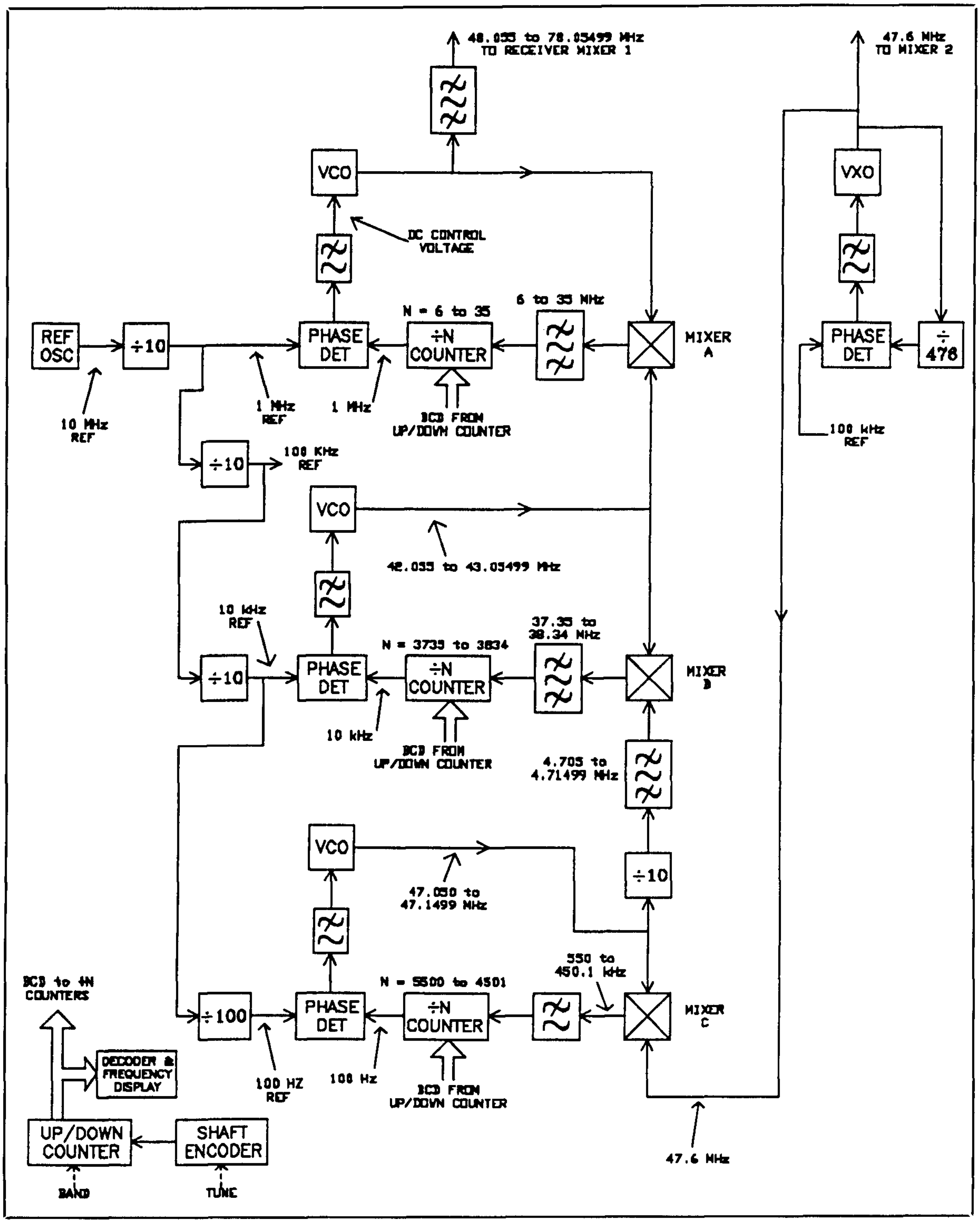

Figure 13. Full synthesizer.

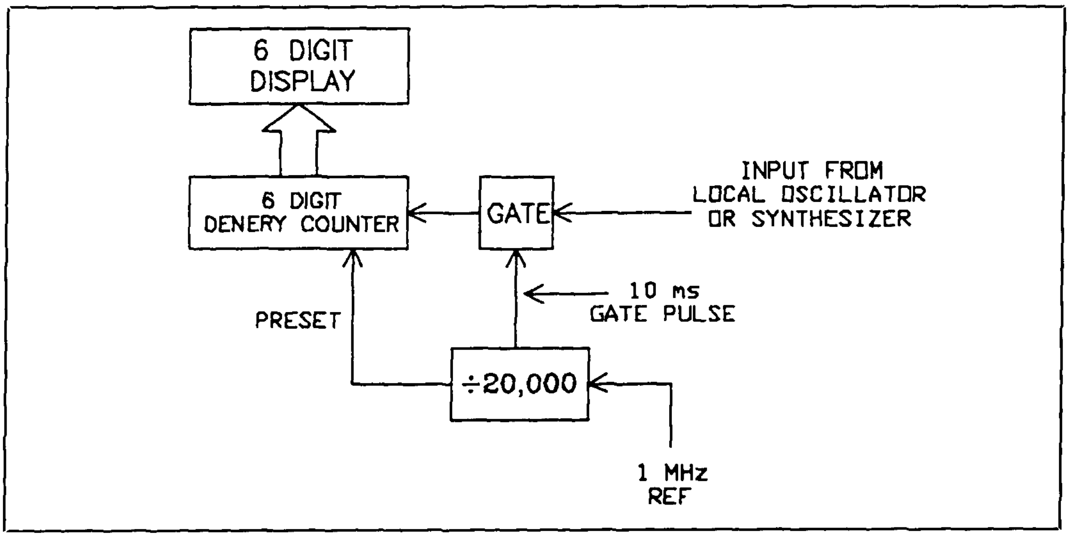

Figure 14. Basic frequency counter.

to be used at the phase detector, thus avoiding problems with loop gain/jitter, etc.

It should be noted that with a simple PLL (for example, **Figure 11**), a drift of, say, 1 in 10^7 in the reference oscillator will simply result in a similar 1 in 10^7 drift at the PLL output. In the more complex arrangement of **Figures 12** and **13** this no longer holds true, and calculation of receiver drift is a slightly more difficult task. For the circuit shown in **Figure 12**, a drift of 1 Hz (1 in 10^7) in the reference oscillator causes a 0.1-Hz drift in the 1-MHz reference, and receiver drift will be about 0.6 Hz close to zero frequency (where $N = 6$). At near 30 MHz (where $N = 35$), receiver drift will increase to approximately 3.5 Hz.

An analysis of the even more complex circuit of **Figure 13** reveals some very interesting facts. By careful design, the idea of the secondary error-correcting loop can be extended (rather magically) to even provide partial correction of frequency drift of the reference oscillator. In the design shown, correction near zero signal frequency is quite good; 1 in 10^7 drift will cause only 0.046-Hz receiver drift, while receiver drift is 2.9 Hz at 30 MHz. Compare this with the drift due to the reference oscillator of **Figure 12**. (I leave it to you to figure out exactly how this correction occurs. Just plug in some numbers and see what happens!)

Various "nice" features can be incorporated into fully synthesized receivers, like auto-bandswitching with continuous tuning right through the spectrum (no need to wind the VFO back and forth when changing bands), direct frequency setting via numerical keypad, fast and slow tune rates, memory units in which frequencies can be stored in channels for instant recall, frequency and channel scan, computer control by external data bus, and so on. In some receivers, "almost-but-not-quite" full synthesis is employed in which a variable crystal oscillator (VXO) is mixed into the loop to interpolate the final 100 Hz (usually called the "clarifier" or "Δf" control). Receiver independent tuning (RIT) can also be implemented in a similar way, but using a range of, say, 8 kHz.

A large proportion of the circuitry of the full synthesizer is digital, including all frequency dividers, the phase detectors, the shaft encoder, up/down counter, frequency display circuits, and all the "optional extras" mentioned previously. This digital circuitry is invariably implemented by the extensive use of integrated circuits, in some cases using LSI chips. Spectral purity of frequency synthesizers is of great importance, and close-in phase noise leads to a little known phenomenon called reciprocal mixing that reduces the dynamic range of the receiver (see **Reference 3**).

Direct Digital Synthesis. Direct Digital Frequency Synthesis (DDFS) has been developed as an alternative to PLL for low-frequency applications ($\leq$100 kHz). The technique employs a microprocessor to generate a "mathematical waveform" in software (using a recursion equation or a look-up table in ROM), which is then output as digital data, and finally converted to a "real" analog waveform using a D/A converter. The microprocessor is clocked by a high stability reference and the system can be very accurate, with fine frequency resolution and very fast frequency switching. The method was only made possible with the advent of high-speed microprocessors and D/A converters. Advances in these areas are pushing the useful frequency range up towards 100 kHz, although even higher frequencies have been achieved with custom-designed logic circuits. Spectral purity is a problem with DDFS, with quantization noise coming from the D/A converter. This is improved by increasing the number of bits in the data word. Adequate performance can be achieved with a good low-pass filter in the output.

Frequency counters. A further well-known example of the application of digital electronics to "steam" radio, is the use of a frequency counter (FC) and digital display to replace (or supplement) the conventional tuning scale. The FC actually measures the frequency being received by looking at the mixer injection frequency and, applying a suitable IF sidestep, it counts the pulses over an accurate gate period and displays the result. The IF sidestep is required as the mixer injection frequency is not equal to the signal frequency, but rather equal to the signal frequency plus or minus the IF frequency. **Figure 14** shows a much simplified frequency counter suitable for many applications, including the single superhet receiver of **Figure 2**.

For those interested in the details, this FC displays the received frequency using six digits of accuracy, and thus has a resolution of 100 Hz. The output of the ÷20,000 reference counter will be 50 Hz, the period will be 20 ms, and each (positive) gate pulse will be 10 ms. Normally an FC counter is reset to zero after each count period, ready for the next count. For an input frequency of 12.567,800 MHz, for example, it would count 125,678 pulses during each 10 ms gate period, to display 12.567,8 (MHz). However in our application (let's say the superhet of **Figure 2**) a sidestep of "minus" 455 kHz is required as the local oscillator is equal to f_s + 455 kHz. This is achieved by presetting the counter to the 9's complement of 455 kHz (equivalent to subtracting 455 kHz from zero!), thus we preset the counter to 995,450. For a signal frequency of, say, 15.678,900 MHz, the local oscillator will be at 15.678,900 MHz + 455 kHz = 16.133,900 MHz. Now during the 10-ms gate pulse, the counter counts 995,450 + 161,339 = 156,789 pulses as required. (The MSB carry produced is ignored.)

Accuracy can be very good, and basically depends on the accuracy of the reference frequency used to generate the gate period. This reference source can be the same as that for the synthesizer; thus, a single high-accuracy (low drift) reference source can do both jobs. Nowadays, all the complex digital FC circuitry is available on a single chip that can drive a display direct (LED, fluorescent or liquid crystal) and can be programmed for a number of standard IFs. The R1000 uses such a chip, and the various frequencies in the receiver have been carefully chosen so that by mixing the two main synthesizer output frequencies together in Mixer C an output is obtained that is equal to the received frequency plus the standard IF of 455 kHz (see **Figure 15**). This, after a ÷10 prescale, is suitable for feeding the FC chip.

Note that some fully synthesized receivers don't use a true frequency counter but simply decode and display the BCD data used to program the ÷N loop counters. However, this relies on all the loops being in lock (which they should be), and doesn't show the effect of the clarifier or RIT controls (see **Figure 13**).

The modern HF receiver—synthesized and wideband

Figure 16 depicts a possible configuration for a general-coverage HF communications receiver with a first IF of 48 MHz and a second IF of 1.4 MHz. As you can see, the use of frequency synthesis and up-conversion techniques aren't the only changes that have taken place with HF receiver design. Early synthesized receivers retain the tuned RF stage, with separate preselector bandswitch and preselector tune control (e.g., the Racal RA17/217/1217 range of receivers). These are somewhat inconvenient in use, and in the interest of simplicity of operation (considered to be of great importance in the professional field), the tuned RF amplifier has been replaced by a wideband RF stage preceded by a number of bandpass filters. These are usually octave or sub-octave filters; that is, they cover a frequency range where the highest frequency is twice the lowest (octave) or less than twice the lowest (sub-octave). They are normally automatically switched using electronic (diode) switches, by decoding the BCD bandswitch data from the synthesizer. See **Figure 16** for a typical array. (This configuration uses a low-pass filter below 500 kHz.)

Linearity in RF amplifiers and mixers is, as already mentioned, of great importance. (That is, mixers must be linear for two signals arriving at the signal port, or input. However they must, of course, be nonlinear in the case of one signal on each port, or

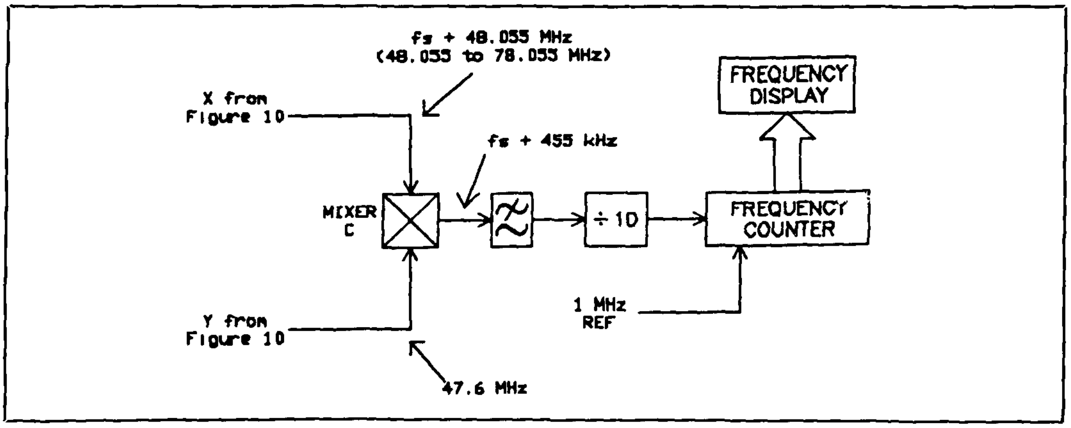

Figure 15. FC arrangement for partial synthesizer of *Figure 10.*

they wouldn't be mixers!) FETs are now universally used as RF amplifiers because they are more linear than bipolar transistors, and have good noise performance. (The dual-gate MOSFET has the additional advantage of a good AGC characteristic.) Linearity can be further improved by using high-voltage supply rails. In some cases, power transistors and FETs have been used with claims that 140 dB of dynamic range have been achieved. Mixers are usually balanced or double-balanced (for good cross-modulation performance) using FETs or diode ring modulators, and use high injection levels to improve single-port linearity.

The use of a front-end attenuator to improve dynamic signal performance (intermodulation, cross modulation, and blocking) has been commonplace for some time, but the use of such a device can reduce sensitivity—often when good sensitivity is most needed. An example of such a situation is when one is trying to receive a small signal in the presence of many very strong in-band and out-of-band signals. With a carefully designed receiver (good dynamic signal performance) this attenuator can be completely dispensed with.

Traditionally, the whole IF amplifier ("IF strip") had a "distributed selectivity." Each amplifier stage of the IF strip contained a tuned circuit that contributed to the overall selectivity of the strip. Again, a change has taken place. Now the IF strip is either wideband, or fairly broadly tuned, and preceded by a block filter that alone determines the selectivity of the receiver. In practice, various filters are provided and can either be selected manually or are switched automatically by the mode switch (almost always using electronic switching). Good quality filters can be made with the nearly ideal response of a flat top (low ripple) with very steep sides. Filters with this characteristic are referred to as having good "shape factor." This is usually defined as the ratio of the filter bandwidth at 60 dB to its bandwidth at 6 dB. A shape factor of 2.0 or better is considered good. The filter's ultimate off-frequency rejection ("stop-band"), including spurious responses, is often better than 80 dB.

A typical array of five IF filters for a communications receiver is shown in **Figure 16**. The 8-kHz filter would be used for AM, the 2.7 kHz filter for SSB, and the other three filters for CW, RTTY, and so on. In practice, a professional quality receiver would probably use two separate asymmetric filters in place of the single 2.7-kHz SSB filter shown—one for LSB and one for USB. If reception of independent sideband (ISB) is required, this duplication would be extended to the whole IF strip and all audio sections. (ISB consists of two completely different audio signals each modulated onto a single suppressed carrier, hence two completely separate IF filters, IF amplifiers, product detectors, audio amplifiers, and audio outputs are required.)

The IF filters may be crystal, mechanical, or (cheaper) ceramic, and the use of such filters means that the old constraint of needing a low second IF no longer applies. Indeed, it's much easier to make crystal filters at IFs above 1 MHz (1.4, 1.6, 9, and 10.7 MHz are common values). For the traditional IF of 455 kHz still in widespread use today, ceramic filters are normally used. Although less costly than crystal filters, they can almost equal them in performance, and can be very small in size.

Note that the availability of an SSB filter in a receiver is very useful for the reception of AM under conditions of deep fading (which seems to be all the time on HF at my location!). The AM is simply received as if it were an SSB signal. Only one sideband is received and the carrier is partly suppressed, so the main cause of fading (constructive and destructive interference between the sidebands) is eliminated. The operator can also switch from one sideband to the other to minimize adjacent channel interference. This technique is sometimes called "exalted carrier" reception.

"Variable IF bandwidth," using pairs of filters and at least two extra conversion

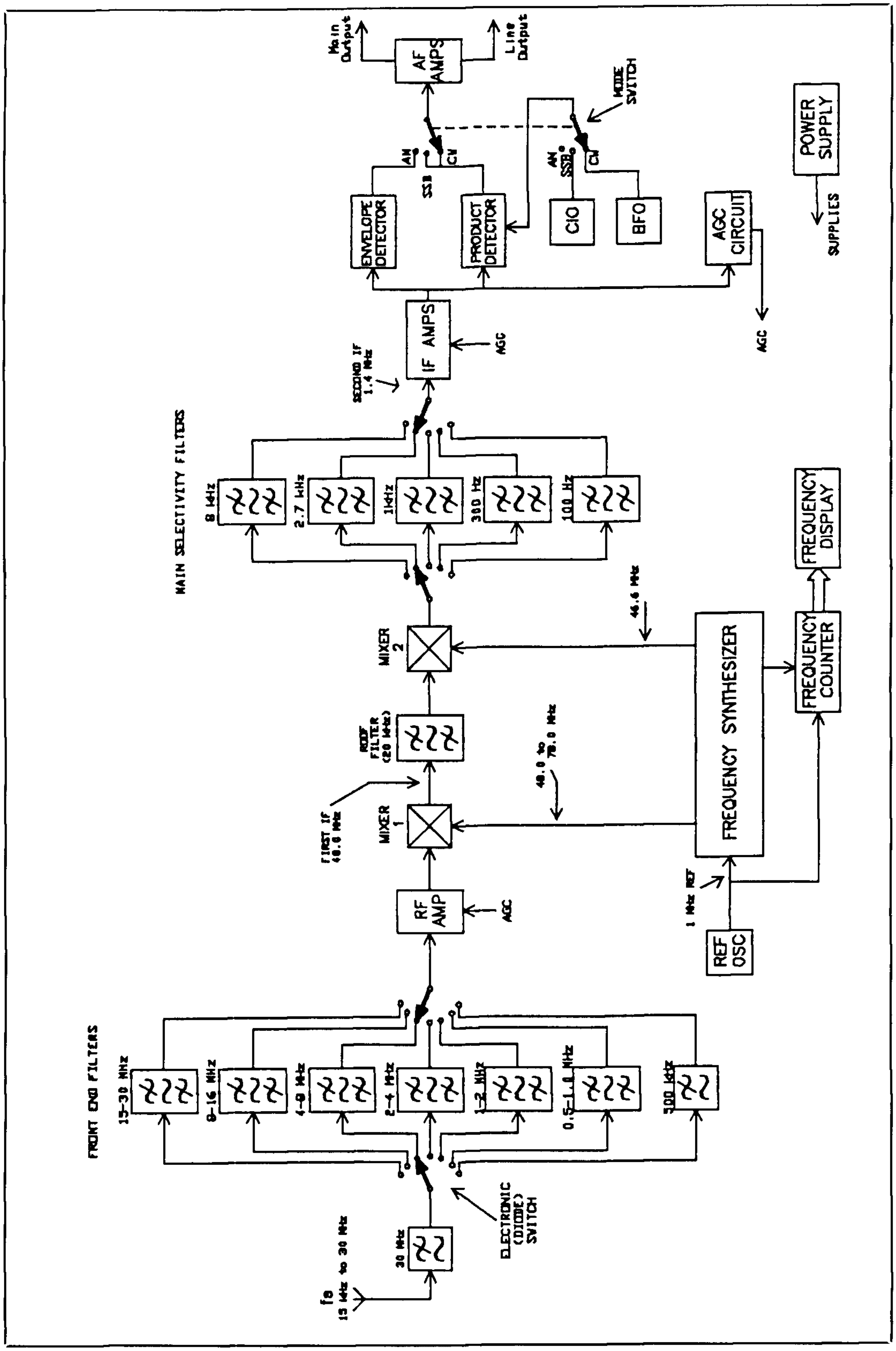

Figure 16. Typical modern HF receiver.

stages, is a technique often employed in imported amateur radio transceivers. **Figure 17** shows a possible scheme to illustrate the principle. With the variable crystal oscillator (VXO) set to 9.455 MHz, the response curves of the two filters coincide. If the VXO frequency is varied slightly, the response curves no longer coincide and the effective bandwidth is reduced (see **Figure 18**). The second mixer returns the IF to 9 MHz and compensates for a signal shift caused by the first mixer. If high quality (8 pole) crystal filters are fitted, good results can be obtained. With two 2.4-kHz SSB filters, bandwidth can be varied from about 600 kHz to 2.4 kHz. For CW with 600-Hz filters, bandwidth can be varied down to 150 Hz or less.

IF shift is another similar technique. Here the received signals can be (apparently) shifted within the IF passband without a change in CW pitch or SSB intelligibility. Thus, an interfering signal can be pushed out of the passband. **Figure 19** shows a simplified block diagram of the conversion and filter stages of a triple-conversion superhet (Trio R820), which employs independent variable IF bandwidth control, IF shift control, and a variable IF notch filter, using the same basic principle of juggling the IFs.

Automatic gain control (AGC) systems have become fairly complex, using considerable amplification to achieve figures of only 2 dB change in output for a 110 dB change in RF input! On SSB and CW, where there is no carrier, the AGC must have fast attack (10 ms typically) and a slow decay (2s), or even better—fast attack—a "hang" of 2s where the output doesn't change, followed by a fairly fast decay (200 ms). For AM, a fast attack and fast decay is required (e.g. 10 ms) to follow fast fading. Often "fast/slow" switches are fitted to give a choice of decay times. The demodulator stage invariably consists of an envelope detector for AM, and a product detector for SSB and CW. Some form of FM discriminator is sometimes now fitted as NBFM has made an appearance on HF.

On SSB the missing carrier has to be reinserted, so either a crystal controlled carrier insertion oscillator (CIO) or the necessary frequencies generated by the synthesizer are used. If the receiver uses a single symmetrical SSB filter, the CIO normally provides injection at the IF minus 1.5 kHz for USB, and plus 1.5 kHz for LSB. The rather superior alternative mentioned, which uses two separate asymmetrical SSB filters, simply requires injection at exactly

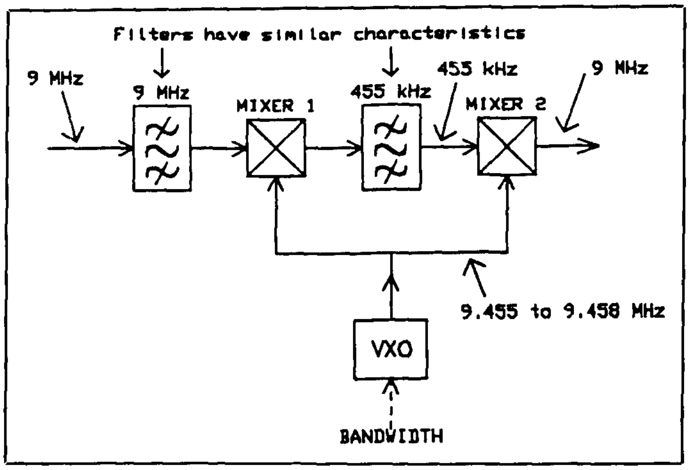

Figure 17. Variable bandwidth IF.

the IF. For CW, a variable "beat frequency oscillator" (BFO) should always be provided.

Discrete circuitry (FETs and transistors) is still often used for the IF amplifier, but complete "IF system" chips are available with a multi-stage IF strip, AM and SSB demodulator, comprehensive AGC system including hang AGC, and an audio preamplifier. These large scale integration (LSI) devices will undoubtedly be increasingly used, but one possible snag that needs careful watching is the introduction of wideband noise after the main selectivity filters.

The noise blanker is another example of the single chip. The chip basically detects noise pulses and gates them out, leaving very brief "holes" of silence. It consists of a separate high gain IF amplifier, noise detector or discriminator, and a gating circuit located in the main IF path to remove the noise pulses. The noise blanker can be very effective in removing regular impulse-type noise (such as noise from electric motors, and the dreaded over-the-horizon-radar "woodpecker"), so it is more useful for QRM (manmade noise) rather than QRN (atmospheric noise). The noise blanker sometimes comes with a threshold control and/or fast/slow switch to cater to different pulse lengths.

The audio amplifier is fairly conventional and is usually implemented these days with a conventional audio power chip of 2 to 5 watts output. A separate audio "line amplifier" is often provided with its output level unaffected by the audio gain control.

In many amateur radio applications, sharp audio filters have been used in lieu of a good IF filter. This has been especially true in the case of CW, as most "ham" transceivers have good SSB filters, but CW filters are often an expensive optional extra.

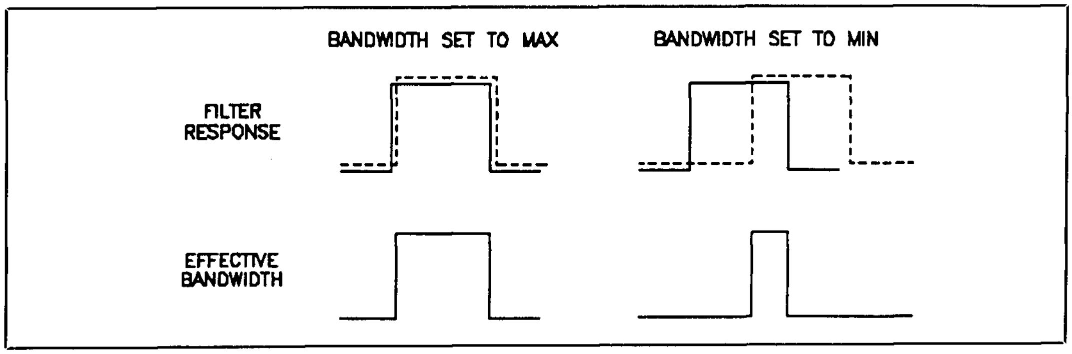

Figure 18. Variable bandwidth response.

These audio filters can be of very high performance, and are sometimes very effective, with (variable) bandwidths down to 100 Hz and good shape factors. In two respects, they are inferior to good IF filters. The IF bandwidth is still wide, so the AGC circuit can respond to adjacent large signals thus reducing sensitivity, and a filter at the audio stage cannot reject what is usually referred to as the audio image frequency. For instance, if receiving a CW signal at the center of the 455-kHz IF, a BFO frequency of 455.6 kHz (offset 600 Hz from signal) will result in a 600-Hz tone. However, another signal in the IF at 456.2 kHz will also produce a 600-Hz tone, and no amount of filtering at audio frequencies can possibly reject this signal! (Does all this sound familiar? It should. Remember the earlier discussion on the superhet and its image frequency problem? Well this image is just the same sort of thing, but instead of occurring at the first mixer stage, it occurs at the demodulator stage which, after all, is just another mixer!)

Digital filtering is the latest digital technique to invade traditional radio. Digital filtering is based on discrete waveform sampling using high speed A/D converters to digitize the samples, and microprocessors to process the data—the filter is all in the mathematics! The advantages are small size, low cost, and the ease of changing filter characteristics under software control! First used at audio frequencies, digital filtering has spread to IF filters, and experimentally, to other signal processing stages. We may soon see receivers that consist of an up converter, roofing filter, and second mixer, with all the rest digital; that is, an A/D converter, all signal processing done in the computer, with a final D/A converter if we want to listen to the output!

Another development is the use of

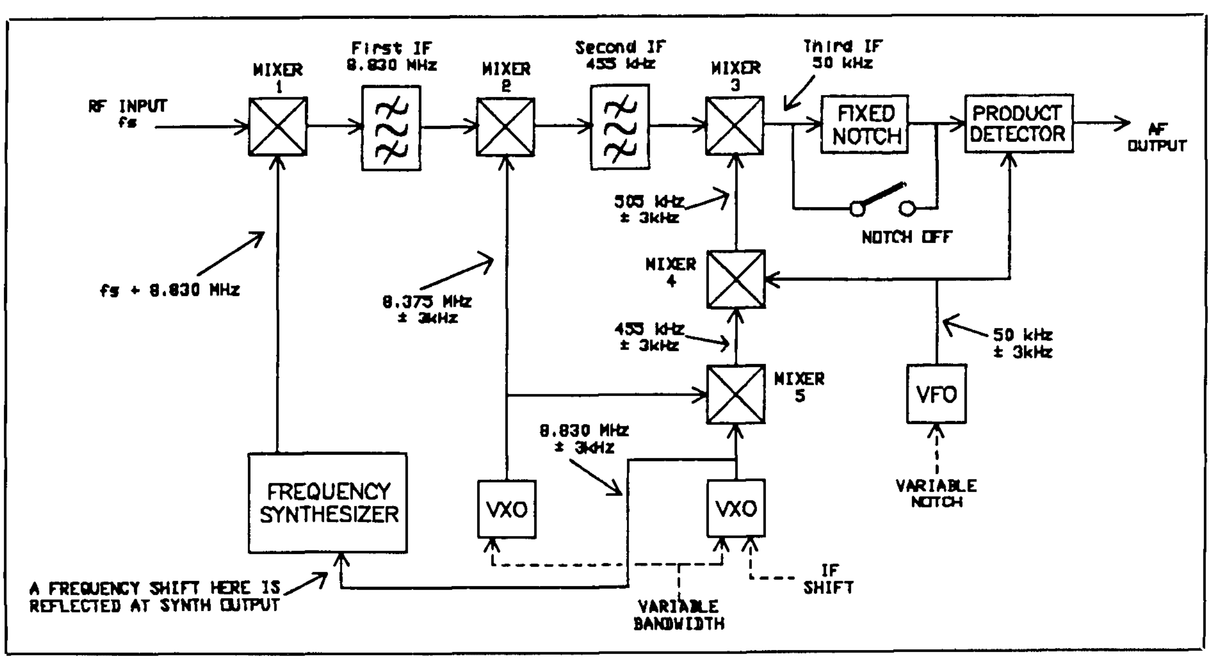

Figure 19. Variable bandwidth, IF shift, and tuneable notch.

microprocessor control to provide an "intelligent" HF receiver! Such a receiver could, for example, have a 1000-channel memory with user-programmed frequencies, modes, filter settings, and so on, all saved in memory. Once a particular channel is recalled using the built-in numerical keypad, the frequency, modes, and filter will be set automatically. The numerical keypad could also be used for direct frequency selection by simply keying in the required frequency, and advanced channel scan or frequency scan features can be incorporated. The microprocessor could also be used for decoding packet radio, RTTY, and other data transmissions. Such a receiver might contain 50,000 or more active devices (if you add up all the devices in every chip), thus receiver complexity will have increased by roughly 5000 times in the space of 50 years!

Finally, consider the man-machine interface. Considerable effort is put into designing front panels that are logical and easy to use (ergonomic). This is made possible by the current trend of making virtually all controls semi-remote. Thus front panel switches actually switch DC control voltages used to operate the "actual" switch, which is an electronic device on the circuit board itself (usually diodes, transistors, CMOS analog switches, etc.). Similarly a BFO knob, instead of being connected by a shaft to a variable capacitor in the BFO, now operates a variable resistor, and a DC control voltage is fed to a varicap diode in the BFO. This gives the freedom to place controls anywhere on the front panel, and also eases the problem of internal layout design considerably. With the replacement of the tuning scale by a digital frequency display, there may now be virtually no controls in the receiver that are physically connected to the circuits they control. Full remote control is therefore possible, with a receiver at a remote aerial site (free from local man-made interference, or QRM) and an operator and his dummy front panel twenty miles away down a microwave link! We've come a long way from the traditional 10-tube "tracking" superhet. ■

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THE ELUSIVE NUMBERS RELATING TO RECEIVER PERFORMANCE

Defining and demystifying four receiver parameters

How many times have you studied the factory specification sheet for a ham receiver and tried to ferret out information on how the unit really performs? It can be quite confusing. The array of numbers do seem to cover the page, but there's not much that tells how the unit performs in either an urban or rural environment.

What exactly are these important measurements that are overlooked in the specification sheets? How would one determine their meaning even if they were given? It's time that this knowledge be made available to a wider population of the ham society. The real research into this area has been performed by such noted technical DXers as ON5DO, G3SJX, DL1BU, WB0JGP, DJ7VD, and N6ND. John Devoldere, ON4UN, assembled all of the data, which covers more than 150 models of receivers and transceivers. The information gathered independently by these hams has real value in that the results on similar models correlates so closely. I'd like to examine four of these parameters, and define and demystify some familiar terms that have been shrouded in mystery.

The receiver noise floor

The receiver noise floor is the minimum level required for a discernable signal response (MDS). Measured data is taken at the 3-kHz point of the receiver's bandwidth. A readable single sideband signal requiring a 10-dB signal-to-noise ratio would be above the noise floor by 10 dB. For example: a modern receiver with a noise floor of − 130 dBm will produce a readable SSB signal of − 120 dBm. For a receiver with an S meter calibrated at 50 μV for S-9, the indication will be S-1. This is a signal of about 0.25 μV into a 50-ohm input on a quiet day.

Two-tone dynamic range

The most important specification is quite possibly the two-tone dynamic range of the receiver as measured at the 20-kHz point of the bandpass. The dynamic range (DR) allows one to compare directly the strong signal handling performance of a receiver before any third order intermodulation distortion products rise above the noise floor. The lower limit of the DR would be the power level of the weakest detectable signal (the noise floor). The upper limit is the point at which intermodulation products are at the level of the noise floor. This is called the two-tone IMD point—the level at which strong signals start to generate audible intermodulation products. IMD above this point increases at a rate of 3:1 with that of the parent signal. Expressed in another way, any increase in the level of the parent signal will produce IMD three times faster than the desired signal.

*We regret that W6YUY became a Silent Key before this article went to press. We wish to thank his wife, Florence, for giving us the opportunity to publish Mr. Bloom's work posthumously. Ed.

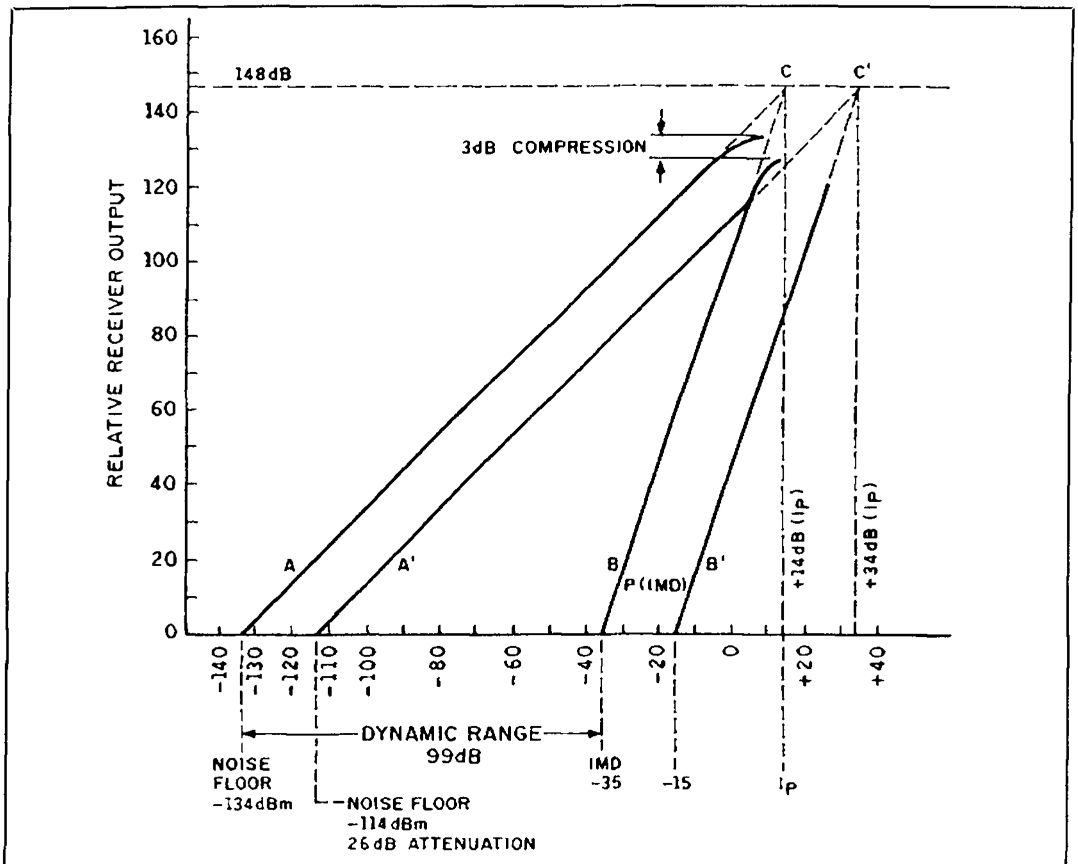

Figure 1. Kenwood TS-940S IMD spectrum.

When expressed graphically, as shown in **Figure 1**, the vertical axis represents the receiver's front-end relative signal level beginning at an arbitrary zero level in dB. This scale also serves to show the relative dynamic range of the IMD (dB). The horizontal axis represents signal levels expressed in dBm. Starting at the receiver's noise floor, we see two plots—the second of which is attenuated by 20 dB. The plots begin at the receiver's noise floor and extend on a 1:1 ratio to a point beyond that where 3 dB of compression for both signal and IMD intersect. This point is called the intermod "intercept point."

The intercept point

The third order IMD intercept point P(IMD) is determined by the ratio of the minimum discernable signal level and the third order intermodulation.

The graph of **Figure 1** is a characteristic plot of a transceiver with qualities very high up on the merit graph of ON4UN's "Dream Receiver"—that for the Kenwood TS-940S. Unfortunately, specifications for other units such as the Kenwood 950, ICOM 761, ICOM 781, Yaesu 767, and FT-1000 were not available at the time of this writing, so I couldn't compare them against the 940S.

The following figures are given in this plot of the Kenwood TS-940S:

Noise floor (N_f): – 134 dBm
Dynamic range (DR): 99 dB at 20 kHz
IMD intercept point (I_p): + 14 dB at 20 kHz

When not given, the intercept points can be determined by a formula where both the noise floor (N_f) and dynamic range (DR) are known (see **Equation 1**):

$$I_p = \frac{2(N_f) + 3(DR)}{2} = \frac{2(-134\ dBm) + 3(99dB)}{2} = \frac{-268 + 297}{2} = 14.5\ dBm \quad (1)$$

and this whole number correlates with that given in the specification sheets.

Similarly, the dynamic range can be determined mathematically when both the noise floor and intercept point is known.

As given: DR = – 134 dBm to – 35 dBm = 99 dB.

$$DR = \tfrac{2}{3}(I_p - N_f) = \tfrac{2}{3}[14 - (-134)] = \tfrac{2}{3}(148) = 98.66\ dB \quad (2)$$

This also correlates with the 99 dB whole number given.

The + 14 dBm I_p point is the important factor needed to construct the graph of the 940S.

Extending the two-tone parent and IMD signals at a ratio of 1:1, 134 + 14 = 148 dB, and projecting both the +14 dB point on the horizontal axis and the 148 dB point on the vertical, we find the I_p (Point C). Now, draw a set of parallel lines for signal and IMD 20 dB up on the base line.

Examining the receiver's +14 dBm intercept point and the −134 dBm noise floor by formula:

$$P(IMD) = \tfrac{1}{3}[2(N_f) + (DR)] = \tfrac{1}{3}[(2 \times 14) + (-134\,\mathrm{dBm})] = \frac{-106}{3} = 35.6\,\mathrm{dBm} \tag{3}$$

This signifies that signals up to −35.6 dBm will be free of any noticeable intermod products, but signals above this level will contain IMD. A signal of −35.6 dBm would indicate S-9 + 37 dB (on an S meter where S-9 is equal to 50 μV). This brings emphasis to the importance of the variable attenuator incorporated in many modern transceivers. An attenuator is included in receiver circuits that have an excess of front-end sensitivity. The TS-940S, for example, has a selection on 10, 20, or 30 dB.

Notice that **Figure 1** shows two sets of projections 20 dB apart. By inserting 20 dB of attenuation to the incoming signal, we bring the noise floor down from −134 dB to −114 dB. This not only moves the signal down by 20 dB but also improves the IMD by 20 dB. However, the dynamic range of the receiver remains the same. We can illustrate what is shown graphically in a mathematical sense.

$$P(IMD) = \tfrac{1}{3}[(2 \times 34) + (-114)] = -15\,\mathrm{dBm} \tag{4}$$

A −15 dB signal will indicate S-9 + 57 dB on the S meter. By inserting 20 dB of front-end attenuation, we can now receive a 20-dB stronger signal level before any third order intermod begins to overload the TS-940S.

This is a pretty useful piece of information that would provide a lot of data on a specification sheet. It wouldn't take a ham long to decide whether to purchase a new piece of equipment!

Noise level

One important piece of data not generally found on sales specification sheets remains. This is the level of the noise spectrum generated by the receiver's local oscillator. These noise sidebands are caused by the phase locked loops of voltage controlled oscillators (VCO) used in modern receivers. Yes, these oscillators generate sidebands that enter the mixer stage to combine with the input signals and the IF. Reciprocal mixing is the result of strong signals joining with the local oscillator sidebands. It introduces off channel signals into the receiver's IF that effectively reduce the receiver's sensitivity. The local oscillator's level at some specific point in the receiver's bandwidth becomes an important specification. This specification identifies the local oscillator's level in terms of −dBc at various bandwidths between 2 and 20 kHz. The level identified with the TS-940S is given as −127 dBc at 20 kHz.

There have been quite a number of technical articles written on "phase noise" that address the ill effects caused by this noise modulation and its generation of unwanted signals. John DeVoldere, ON4UN, one of the world's most proficient DXers, goes into considerable detail on the subjects of cross modulation and reciprocal mixing. His book *Low Band DXing* published by the American Radio Relay League was a must for my library.

I drew curves for several renowned transceivers. The oft mentioned "Signal One"—the DXer's choice in past years—took on a new perspective once I knew the numbers.

The difference between the two models, the CX-7 and CX-11, soon became apparent. The CX-7, with a noise floor of −126 dBm, dynamic range of 62 dB at 20 kHz, and intercept point of −33 dB equate for an IMD point of only −64 dBm. This indicates that high levels of intermod products will be generated with signals above S-9 + 10 dB—not very good.

The CX-11, on the other hand, has a slightly higher noise floor of −124 dBm. The dynamic range of 99 dBm was taken at 200 kHz. This would be about 85 dBm at 20 kHz. The intercept point is given as +24 dBm at 200 kHz. A 20-kHz dynamic range of 85 dBm would shift the I_p to +7 dBm at 20 kHz. Then:

$$P(IMD) = \tfrac{1}{3}[(2 \times I_p) + N_f] = \frac{-106}{3} = -35.6\,\mathrm{dBm} \tag{5}$$

A −36 dBm signal is equivalent to an S level of S-9 + 37 dB. The CX-11 can receive an S-9 + 37 dB signal without increased IMD at this level—a considerable improvement over that of the CX-7.

I, for one, hope that the receiver and transceiver manufacturing sales departments will soon include these four important quantities on their specification sheets. It might be prudent for those who would like to know more about the product they now own to drop a line to the manufacturer of the unit and ask for these numbers. ■

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SPECIAL COLUMN

THE SOLAR SPECTRUM

Understanding the Solar Wind

Several decades ago, the astronomical community was confronted by a perplexing problem concerning the extended atmosphere of the Sun: the solar corona **(Photo A)**. The influence of solar gravity falls off rapidly as distance from the Sun increases, so a static corona could only be sustained by some type of restraining force.

At the time, this constraint was thought to be provided by the inward pressure of interstellar gas. However, theoretical calculations showed that the extraordinary high temperature of the corona—upwards of one million degrees Celsius*—and correspondingly high conductivity, would overwhelm such a weak influence and allow the farthest reaches of the Sun's atmosphere to continuously expand into space.

In unrelated research at about this same time, Ludwig Biermann[1] challenged the notion that the pressure of sunlight caused the tail of the comet to perpetually point away

**Flares can raise local temperatures by a factor of 10 to 30.*

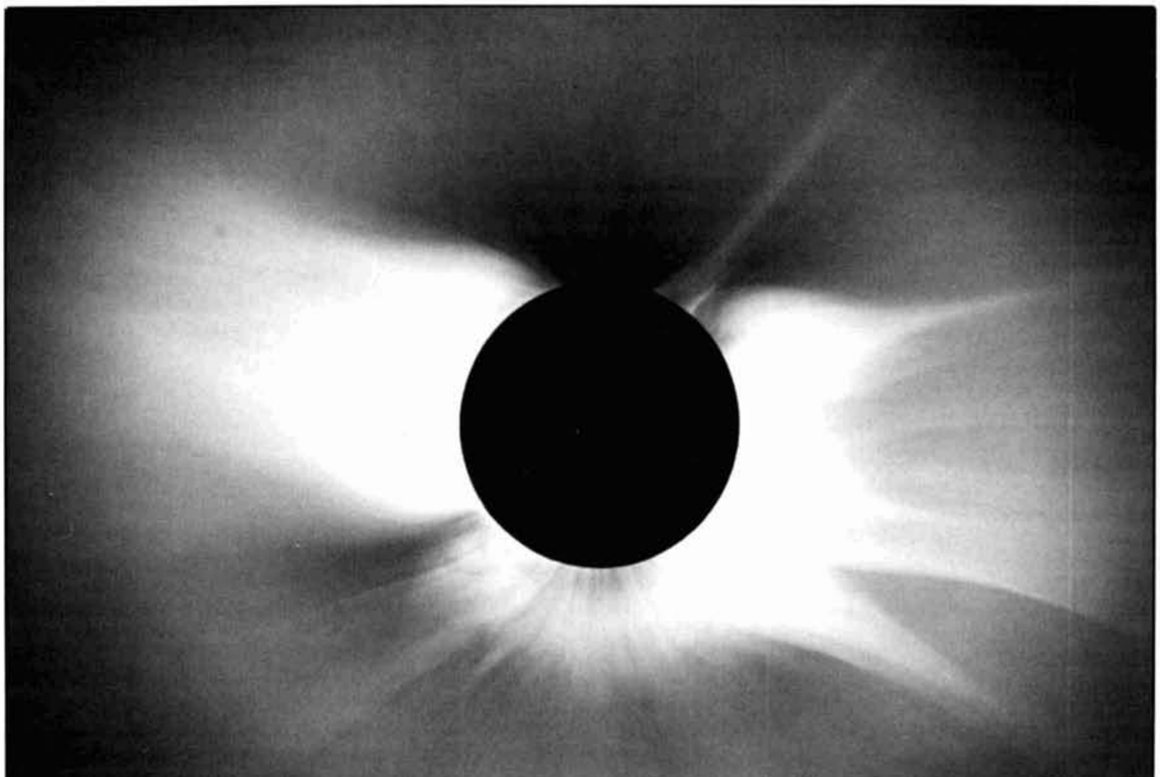

Photo A. This spectacular photograph of the June 1973 solar eclipse was taken by a five-man team from the *High Altitude Observatory - National Center for Atmospheric Research*. The group employed a camera devised by G.A. Newkirk, equipped with a special radial-density filter that progressively dims the corona's brightness as distance from the Sun decreases.

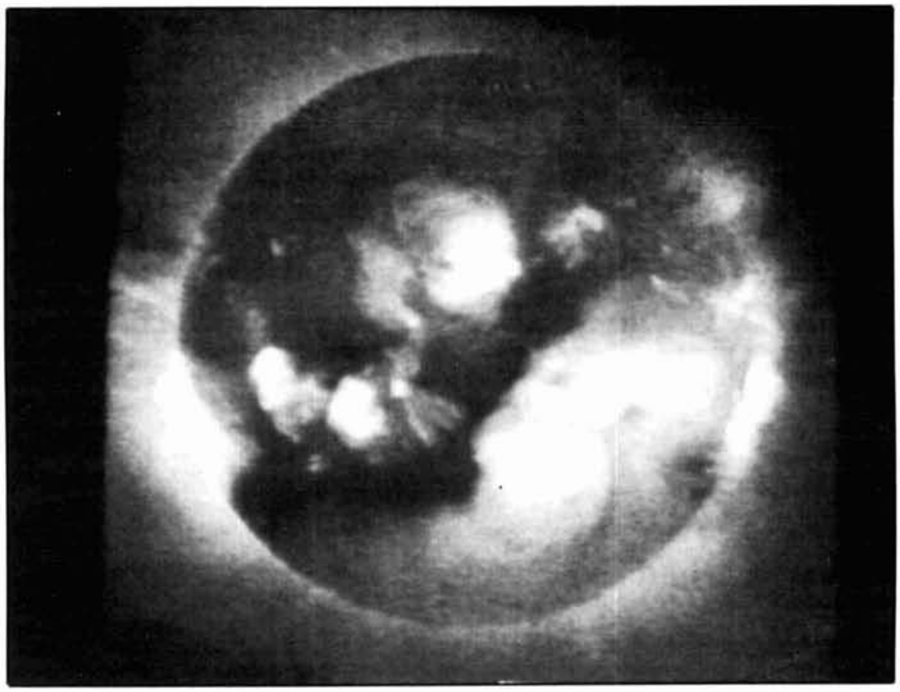

Photo B. The X-ray corona as it appeared near the end of November 1991. A long series of extraordinary images such as this has been obtained by the Soft X-ray Telescope aboard the Japanese Yohkoh satellite. The instrument was designed and built as a collaboration between the Lockheed Palo Alto Research Laboratory, the National Astronomical Observatory in Japan and University of Tokyo, with support from NASA and Institute for Space and Astronautical Science (Japan).

from the Sun. Biermann found that this phenomenon was far too weak to produce this effect, and suggested that a more likely cause was a steady outflow of particles streaming away from the Sun.

We now realize that Biermann was essentially correct, at least qualitatively. Incredible as it may seem, the Sun loses a staggering, several million tons of solar material in this process every second! (No need to worry, however. At this rate, it would take over 150 billion years for the Sun to lose just one percent of its present mass.[2])

Biermann's concept was subsequently confirmed by University of Chicago physicist Eugene Parker,[3] who argued that only two explanations could describe this phenomenon. According to one portrayal, the "solar breeze," outflow rate near the Earth would be low, perhaps only a few kilometers per second. But Parker felt that the velocity was actually considerably higher, a "solar wind" that rushes by the Earth at a speed hundreds of times greater.

Unfortunately, the flow of material away from the Sun could not be measured from the Earth since it would be deflected by the geomagnetic field. Therefore, the first hints of a solution came from instruments aboard early Soviet and American satellites which indicated that the rush of particles in the terrestrial environment exceeded the speed of sound in the corona ($\sim$200 km/sec). The issue was concluded in 1962 after additional measurements from the Mariner II spacecraft showed a velocity that varied between 300 and 800 km/sec—an astonishing 650,000 to 1,700,000 miles per hour!

As outlined in previous columns, flares on the Sun are not the only solar phenomenon traditionally thought to initiate geomagnetic and ionospheric storms. We discussed another such source, disappearing solar filaments, in the Winter 1992 issue of *Communications Quarterly*. However in the final analysis, a third origin—high-speed solar wind streams associated with low-density regions of the corona—may play the greatest role.

For nearly a century, astronomers suspected that some type of localized source on the Sun was responsible for certain disturbances in the geomagnetic field; storm conditions which seemed to fluctuate regularly according to a period of just under four weeks. The first indication that such formations existed came from a pre-war investigation by Chapman and Bartels[4] who suggested that unseen, "M-regions," were the origin of recurrent storms.

The high temperature of the corona causes it to radiate strongly in UV and soft (low-energy) X-ray wavelengths. Therefore these long sought features, now called "coronal holes" (**Photo B**), are especially visible in this range. Unfortunately, extra-terrestrial X-radiation is difficult to detect at the Earth, so detailed information about these unique structures necessarily awaited the X-ray pictures taken from the Skylab spacecraft. (Nowadays their presence can also be inferred from solar images in HeI 10830, principally at Kitt Peak National Solar Observatory. See **Figure 1.**)

Subsequent investigations[5] have shown that the more persistent of these phenomena produce high velocity (>500 km/sec) wind "gusts" that are responsible for geomagnetic disturbances which reoccur at intervals of about 27 days (the period of apparent solar rotation), as a hole repeatedly returns to its original position relative to the Earth.

Coronal holes are associated with unipolar regions on the Sun which have an "open" magnetic geometry. That is, their magnetic field lines extend far into interplanetary space rather than loop back to the photosphere, as do those associated with closed regions. The hottest portions of the corona are constrained by the Sun's magnetic field, so this factor aids the mechanism which causes high-speed solar wind conditions to occur; ionized material flows easily along such a path.

Long-lived holes are primarily a property of the declining solar cycle, a situation we will encounter during the next several years. They form as low-latitude holes expand upward to link with those above the Sun's poles. The circumstances are much different during a cycle's ascent to maximum. At this initial phase, low-latitude holes have been found to be unstable, short-lived features which are characterized by sudden eruptive outflows.[6] Interestingly, the wind gusts spawned by these events generate compression zones which could be mistaken for the blast effects from flares.

Even though its density near the Earth is several million times less than the inner corona—only a few particles per cubic centimeter—the solar wind provides the link in the solar-terrestrial relationship. The charged wind, actually a flowing plasma of hot electrons and protons (hydrogen ions), cannot easily pass across the Earth's magnetic field except near the poles where the field lines are more nearly vertical, and so blows mainly around the Earth.

This process forms the magnetosphere into a teardrop-shaped feature which normally extends sun-ward for about ten Earth radii, and to more than 100 radii away from the Sun. As this occurs, the boundary between the wind and magnetosphere is turned into a powerful natural generator.[7] Strong currents are created in this manner; perhaps as high as one million megawatts.

When particles carried in the wind are able to penetrate to a sufficiently dense portion of the atmosphere, they give rise to aurorae through a process known as collisional ionization. As a consequence, radio communications may be disrupted, particularly HF propagation over the poles.

Around the times of vernal and autumnal equinox, the direction of the Earth's magnetic field axis is more nearly at right angles to the solar wind flow than at other times. Therefore, incidents which occur during the spring and fall are more likely to affect our environment than similar events at the solstices.

New information concerning these occurrences suggests that some of the conditions historically attributed to solar flares may actually mask short-term effects from low-latitude coronal holes. If this analysis[6] proves

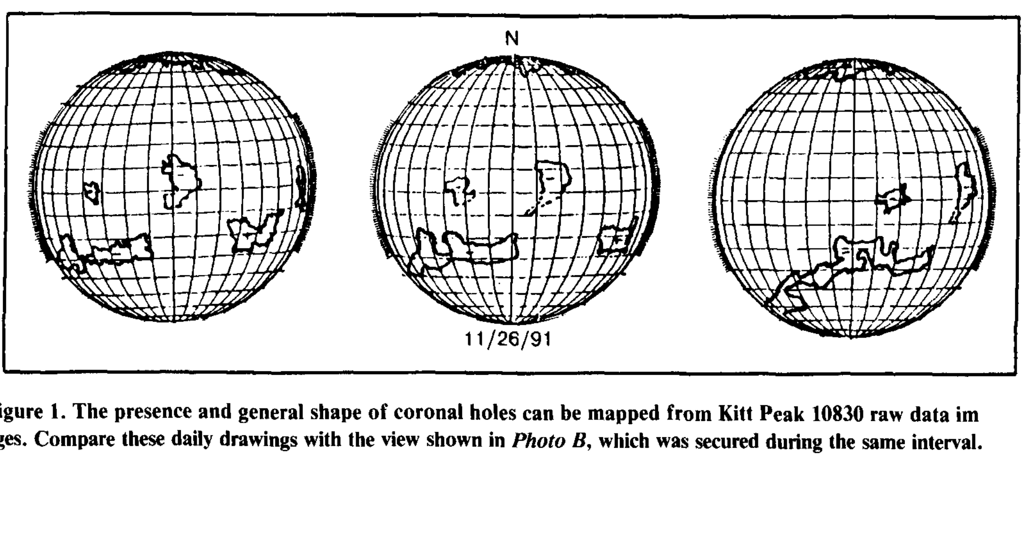

Figure 1. The presence and general shape of coronal holes can be mapped from Kitt Peak 10830 raw data im ages. Compare these daily drawings with the view shown in *Photo B*, which was secured during the same interval.

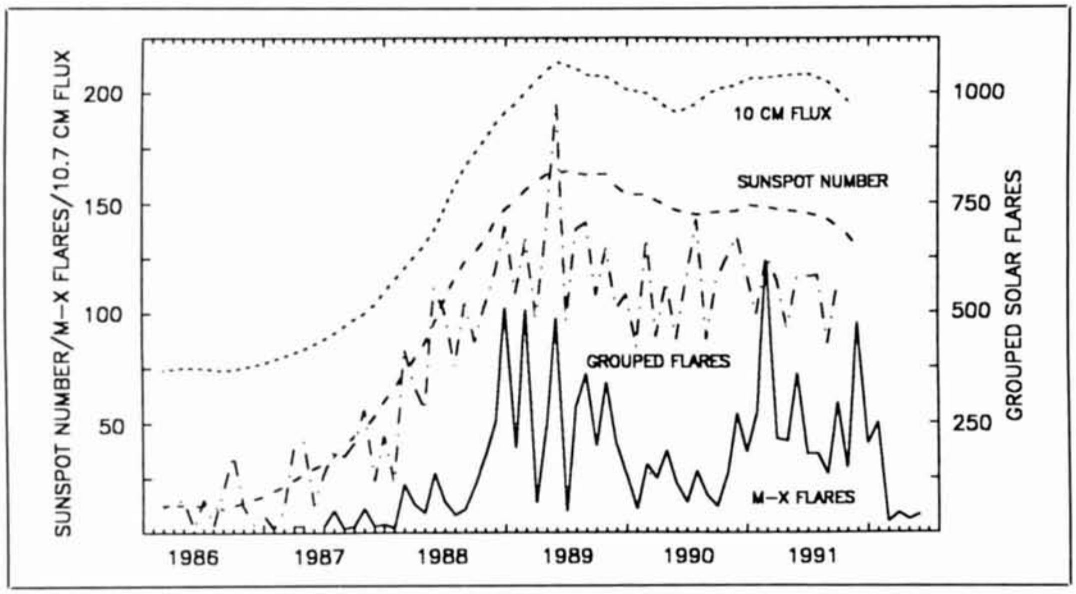

Figure 2. The trend of several indices of solar activity during solar cycle 22. (Sunspot and flux indices have been smoothed.)

to be correct, flares and some other active solar phenomena would play only peripheral roles in the onset of magnetic storms. At the least, some connection between coronal holes and these events is virtually certain, although the scenario is complicated when the Sun is very active.

Recent activity and short-term outlook*

Solar activity declined precipitously during March, a trend which continued until the last week of June. It appears that prediction models failed when they suggested that a substantial portion of this period would be one of relatively enhanced activity.

The strongest X-ray flare to be recorded since October 1991 occurred on June 25th, disabling communications facilities in polar and auroral regions for the remainder of June. These conditions were bolstered by a second class X flare which erupted from behind the Sun's west limb three days later. Additional fading and absorption was noted at mid-latitudes during the peak of this major geomagnetic storm. The Sun was also active for a brief period in May, but the impact on the terrestrial environment was limited to aurorae and a short disruption of high-latitude communications.

With these exceptions, solar activity at the beginning of July is near the level experienced in early 1988 (**Figure 2**). At this time, activity is expected to hover around current levels for a short period and then decline to a lower plateau. ■

**A portion of this information was taken from the SELDADS* database.

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4. S. Chapman and J. Bartels, *Geomagnetism*, Oxford University Press, 1940.
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7. S. Akasofu, *Sky & Telescope*, 534, *December 1982*.

LETTERS (from page 8)

Early superhets used IFs in the 50 to 85-kHz range (5000 meters). The IF transformers were both fixed tuned and broad, and resembled early audio transformers. The assumption made in the article concerning the ability of these sets to work with "sum frequency," or high side LO, is probably wrong considering the relatively low IF used. While RCA held the rights to the superheterodyne design and had the commercial market cornered, it was still possible to homebrew a working superhet. Circuits for superhets appeared in such diverse publications as the *Citizens Radio Call Book Magazine*, the *Los Angeles Eveing Express* and *Radio News*. Sampson Electric Company of Canton, Massachusetts, offered a transformer super-kit for aspiring superhet builders: three long-wave transformers for the 5000-meter wavelength, a filter transformer, and an oscillator coupler—all for $22.50.

Peter J. Bertini, K1ZJH
Somers, Connecticut

Thanks, gentlemen, for bringing this information to our attention. Ed.

A critique

Thanks for the early "author's" edition of *Communications Quarterly*. I read it all before the regular subscription copy arrived. A collage of thoughts that came to me while I read it might be of interest to you.

Technical Conversations is a good section to keep as a regular. Maybe you could view it as a "Letters to the Editor" column presenting just the technical side of correspondent's contributions. I would much prefer pages spent on this type of information than an opinion column like that presented on page 99 of the Winter '92 issue. I count on *Communications Quarterly* for technical excellence and details that prod my creativity. Less expensive publications cover the raw opinion sides of issues.

The "Radio Receivers of the Past" was good in this respect. It took control for Joe Carr to take a normally flowery issue and turn it into a foundational and educational presentation. There are a lot of counselors, facilitators, etc. Real teachers are becoming rare. He did a good job.

Communications Quarterly is presenting an image. Roy Lewallen (page 92, Spring issue) aggressively called you on that point. The sharp black and white graphics you publish contribute significantly to the image of *Communications Quarterly* (don't go to color). Considering the variance of ways the same information could be presented, I am again reminded of *The Visual Display of Quantitative Information* referenced on page 62 of the Spring issue. You really should consider picking up a copy.

Thanks again.

Brian Mork, KA9SNF
Oscoda, Michigan

Thanks for the comments, Brian. Constructive criticism from both readers and authors is always welcome. Brian's article "A User's View of Charge Coupled Device Imaging" begins on page 49 of the Spring 1992 issue of Communications Quarterly. Ed.

ADVERTISER'S INDEX

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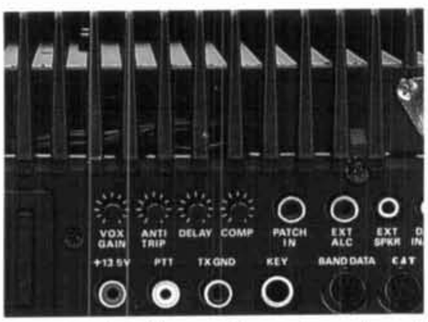
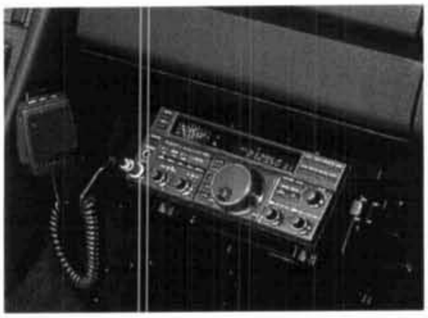


Rear Panel: Seldom used controls are located on rear panel allowing clean uncluttered front panel design.



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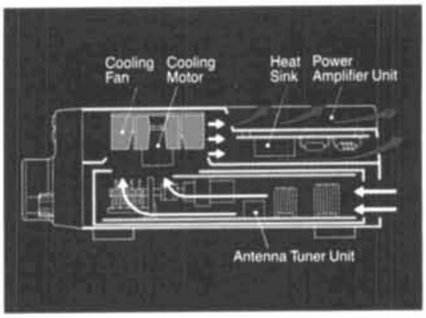

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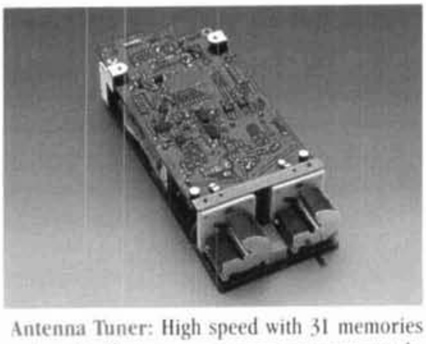

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