two-element hf beams

Tune these medium-sized high performance antennas right from the shack

Physically small beam antennas that represent the least compromise in gain and directivity have been discussed in the literature.[1] Large antennas, for those for whom size is no problem, have received widespread coverage; the W2PV series of articles, for example, includes a wealth of material on large Yagis.[2] Yet the topic of medium-sized antennas — which includes the majority of Amateur beams — remains an area of uncertainty, about which many have sought without success for more information. The quad-versus-Yagi controversy continues unabated; conflicting claims are made for what might appear to be a bewildering variety of different beams, and an imperfect grasp of essentials has turned an inherently simple situation into one of needless complexity, with two-element beams deprived of their rightful status.

how small two-element beams work

In **fig. 1**, we have a bird's-eye view of two identical vertical elements carrying equal currents and spaced by a small fraction of a wavelength, with the plus and minus signs indicating that they are initially fed in opposite phase, thus tending to cancel each other.

At this point it's not important to know how the currents got there. Energy arriving at a specific point in space travels a different distance from each vertical element. This difference in path length and the opposite polarity of each vertical causes the maximum radiation from the array (two elements) to be along a line that goes through the elements. The two fields combine vectorially as shown in **fig. 1B**. For small angles, (ϕ_0) halving the angle halves the field. As one moves around the antenna, the difference is reduced; therefore the angle reduces, producing the directional pattern shown in **fig. 1C**, which is independent of spacing *as long as the angle remains small.* It follows that because energy remains similarly distributed throughout space, signal strengths must also be independent of spacing — provided there are no losses. Usually one introduces an electrical phase shift, ϕ. If this is equal to the spatial phase shift, ϕ_0, cancellation takes place in the reverse direction, producing the well-known cardioid pattern of **fig. 1D**. As the electrical phase shift, ϕ, is reduced, the null in the back direction splits into two. It gradually shifts around, with the back lobe increasing in strength until we arrive back at **fig. 1C**. However, for a given ratio of ϕ/ϕ_0, the pattern remains independent of spacing.

In the case of horizontal beams the directional pattern of the individual elements is superimposed on the beam patterns derived in accordance with **fig. 1**, but the principles are the same. Because no dimensions are mentioned, it follows that for two elements *the directive pattern — and therefore the gain — depend only upon the phase shift ratio, ϕ/ϕ_0, and are independent of the size, shape, or spacing of elements, provided the dimensions are not excessive.* This rule is reasonably accurate for element lengths up to about 0.7λ, and as shown in **fig. 2**, for spacings up to 0.2λ. It starts to break down if there are regions of high current that are separated by a substantial fraction of a wavelength, the λ/4 separation between top and bottom of a quad loop not being appreciable in this context. In this case, a directive pattern results through the addition of fields — a completely different mechanism that is the basic principle of large arrays. **Figure 2** shows the effect that element spacing has on gain and, from another source, a very similar curve for three elements.[3]

The basic statement emphasizing gain as primarily a function of the phase shift ratio rather than spacing — though it seems physically obvious — is in flat contradiction of widely published figures. These figures, derived mathematically for parasitic arrays, show gain and directivity to be *critically* dependent upon spacing and whether an element is tuned as a director or reflector. But although the calculations are indeed correct, they happen to be the wrong ones! More accurately stated, perhaps it's the designs to which they relate that are faulty, since I have assumed equal currents, whereas normally performance is sacrificed if the elements are straight.[5] As can be inferred from **fig. 3**, this is the worst possible shape because it minimizes coupling, consequently precluding the possibility of the presence of equal currents, except with very close

Les Moxon, G6XN, Gorse Hill, Tilford Road, Hindhead GU26 6SJ Surrey, England

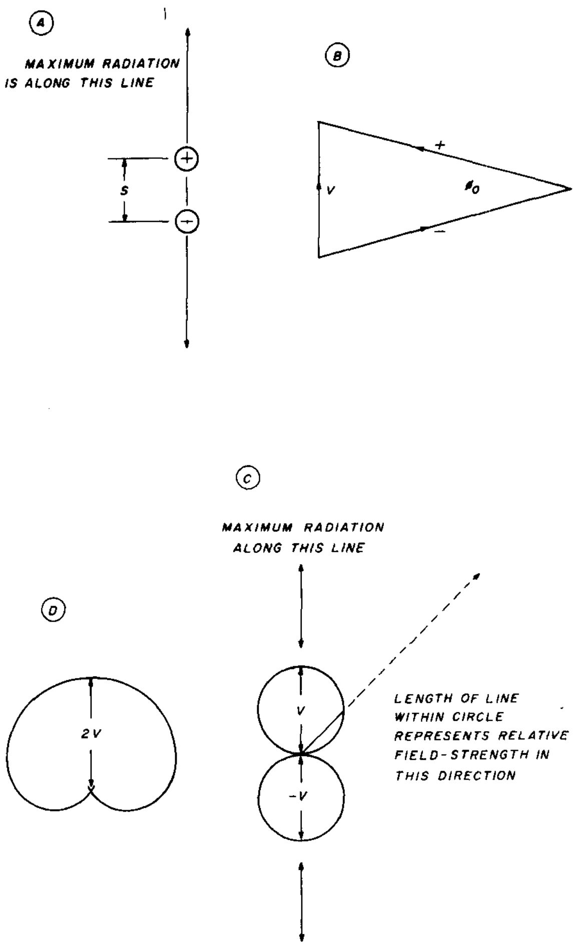


fig. 1. (A) Two closely-spaced sources separated by distance "S." (B) Vector addition results in voltage V that for small angles is proportional to ϕ_0. (C) Polar plot shows variation of field voltage as a function angle to observer. No additional (electrical) phase shift has been introduced — i.e., $\phi = 0$. (D) Field pattern if electrical phase shift, ϕ equals physical phase shift (ϕ_0).

spacing (0.05λ). Nulls in the directional pattern are filled in and gain is also adversely affected.

Coupling between straight elements can be increased by moving them closer together. But for this to be effective, spacing has to be reduced to about 0.05λ, which is normally regarded as unacceptable because of reduction in bandwidth and efficiency. For this spacing, my calculations and those of W2PV are in close agreement. In my case, however, due to the assumption of equal current amplitudes, dependence on physical dimensions has been eliminated.

Inductive coupling (**fig. 3A**) isn't advised because of the reduction in radiation resistance. More often, natural coupling tends to be capacitive and needs only to be supplemented. One advantage of the quad is that loops couple more tightly than straight dipoles; this probably accounts for its popularity despite a poor reputation for survival in high winds. On the other hand, bent elements (see **figs. 3B** and **3C**) lend themselves to the design of more compact but equally efficient antennas with overcoupling rather than undercoupling as the more common fault. This is easy to correct either by an alteration of spacing or, if necessary, neutralization[1], as shown in **fig. 3C.**

mutual impedance

From the narrowing of the radiation pattern, it is evident that there must be gain relative to a dipole. But how, one might ask, can this be if the elements are tending to cancel each other? The answer lies in the fact that the element currents rise to whatever value may be necessary to account for the observed gain, and they can do so only by virtue of the *mutual resistance* between the elements which subtracts from the self-resistances when closely spaced elements are excited in antiphase. Without mutual resistance, there can be no power gain; these quantities are inseparable, so that given equal currents, one follows from the other. On the other hand, mutual resistance alone cannot achieve the degree of current equality necessary for obtaining deep nulls. This requires *mutual reactance*, which exists in most cases but may need to be increased or decreased, with reflector operation requiring negative reactance. Mutual resistance, R_m and reactance, X_m data appears to be available only for straight λ/2 elements (**fig. 4**), but the "size rule" implies that mutual resistance bears a constant relationship to the single-element radiation resistance (R), and so the mutual resistance can, if necessary, be inferred from it. Likewise, if the elements are self-resonant, mutual reactance can in principle be determined from

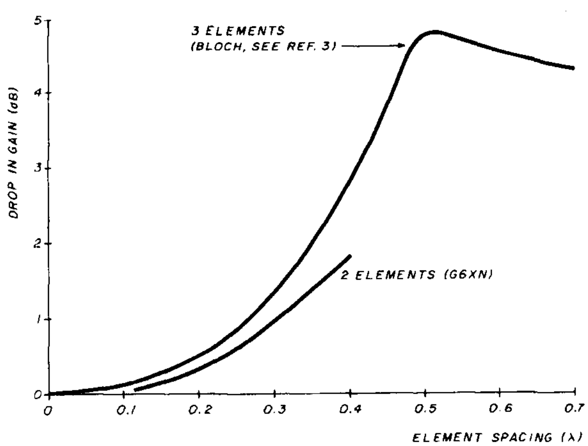


fig. 2. Decrease of gain with increase of spacing. Note that gain is a minimum for both two- and three-element beams as spacing diminishes though practical constraint impose a lower limit of about 0.1λ.

directions of minimum response by obtaining corresponding values of ϕ/ϕ_0 from **fig. 5**, since with no detuning the phase shift ϕ is determined solely by the phase angle of the mutual impedance — i.e. by X_m/R_m so that by knowing ϕ, ϕ_0 and R_m we can evaluate X_m. This method has not been evaluated in practice and its use is restricted by the fact that with large departures of X_m from its optimum value the nulls will not be deep enough for their direction to be determined. In general, however, with two elements there should be no need to know the actual value of X_m since constructions are available which allow it to be adjusted by trial and error. Additionally, by virtue of the "size rule" calculations or measurements for a set of dimensions for which X_m is known can be applied to any other, with due allowances for differences in R. Calculations, simplified by assuming equal currents,[5] have led to further results:

- Radiation resistance (R_B) for a parasitic array is given by

$$R_B = 2\,(R - R_M \cos \phi)$$

- For an array in which each element has its own feedline, the radiation resistances are:

$$R\ director = R - R_M \cos \phi + X_M \sin \phi$$

$$R\ reflector = R - R_M \cos \phi - X_M \sin \phi$$

With a resonant reflector and equal currents, the null directions are approximately 130 degrees relative to the beam heading in all cases. The total resistance is the same as before, but that of the reflector (or director if X_M is positive) can be zero or even negative, which bodes ill for the matching process. Until now, driven operation has been the usual method of trying to equalize currents, but the reason for some failures may now be apparent, particularly because the usual phasing lines must be matched for correct operation. Solutions to this problem have been discussed before.[1,5] Driven operation remains a possible solution to the problem of obtaining equal currents with straight elements, and a number of such antennas have been described. Most of these, intended to be reversible, specify λ/4 spacing, which is too wide, or λ/8 spacing, which is convenient and simplifies the mathematics to the extent that X_M disappears. Unfortunately, because of mutual reactances of opposite sign, which each element induces into the other, a very high VSWR exists in each of the individual feeders. This high VSWR may not be noticed because it doesn't appear in the common feed from the transmitter; if it were corrected, the beam couldn't be reversed because the correction would then be of the wrong sign and make matters worse. Open wire lines can be used, but result in excessively narrow bandwidth (though I've overcome this by using folded dipole elements).

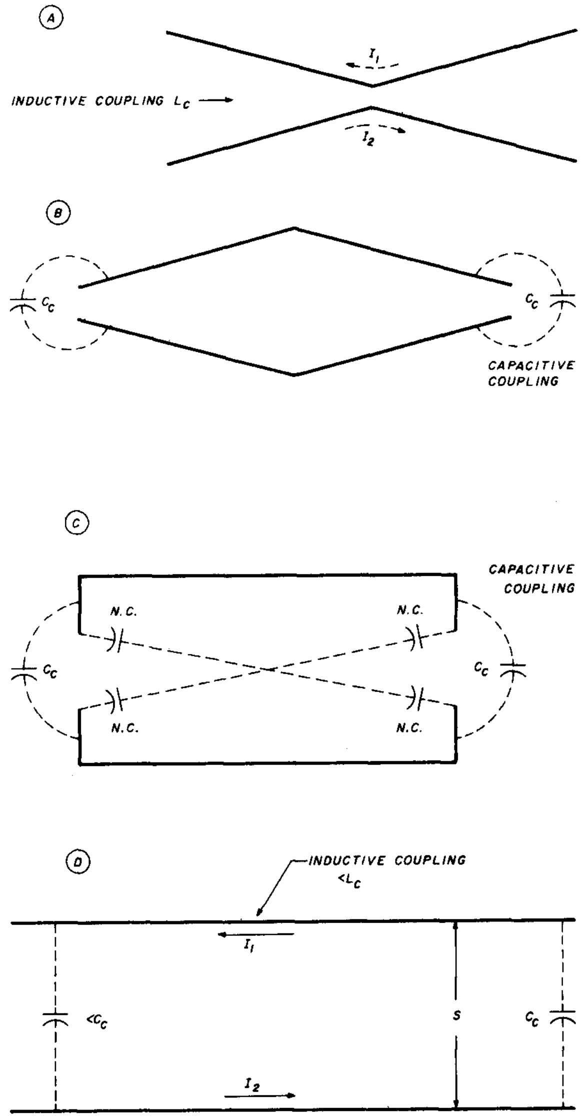


fig. 3. Inductive and capacitive coupling of two elements. (A) Elements bent inwards at their centers couple inductively. (B, C) Ends bent inwards couple capacitively. (D) Excessive coupling may be neutralized by capacitors as shown in (C). In (D), the two couplings cancel each other if S = λ/8, leaving only the resistive coupling, R. Straight elements minimize coupling.

Overcoupling is a problem likely to be experienced in the case of quads with less than 0.15λ spacing. The "Swiss Quad" (with 0.1λ spacing) gets around this by driven operation, but as we'll discuss later, there are many advantages to be derived from resonant feeders, including the possibility of increasing coupling or neutralizing excess coupling by means of capacitance between the lines. **Figure 5** shows the dependence of gain, radiation resistance, and null directions on the phase shift ratio.[1,5] If any one of these quantities is known, all the others except radiation resistance (R_B)

follow from it. For the last parameter, we also need to know the value for a single element (R); with a few exceptions, this can be obtained from **fig. 6**. The method of calculation is explained in reference 5. **Figure 7** is a further extension of **fig. 5** showing the direction and magnitude of the back lobes in the radiation pattern. They demonstrate the crucial importance of the ratio ϕ/ϕ_0 in determining all aspects of the performance of two-element beams (including bandwidth, since this is linked to radiation resistance, as discussed later).

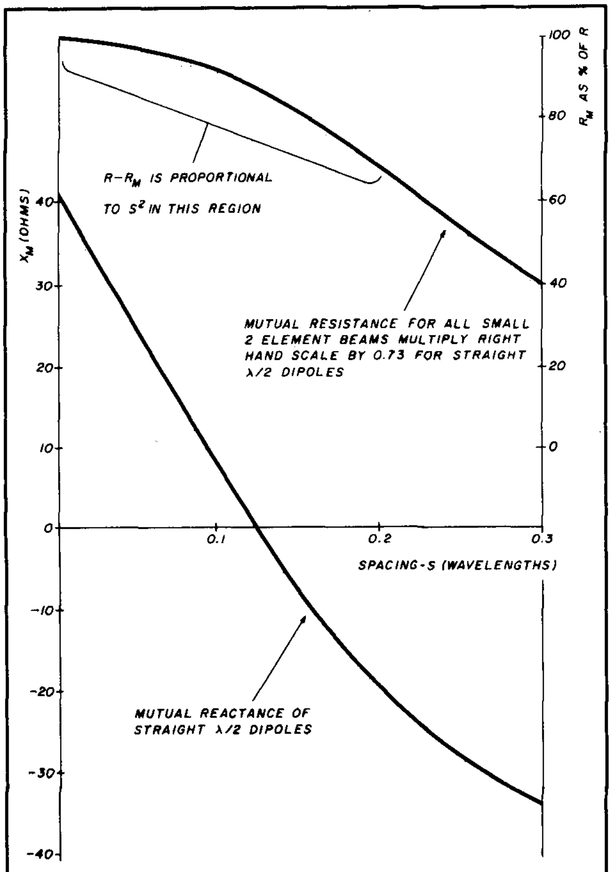


fig. 4. Mutual resistance R_m and mutual reactance X_m as a function of spacing for two-element beams.

directivity gain

So far we've assumed that there are no losses. Apart from feeder loss, other losses may occur because of proximity to nearby structures, use of very thin wire, or as the limiting factor when trying to make an antenna as small as possible. Besides radiation pattern distortion, currents induced in booms or supporting structures due to lack of symmetry may introduce additional losses. If equal currents and correct phasing are maintained, losses as such have no effect on directivity which, because of high external noise levels, is usually the sole requirement in the case of reception on the hf bands.

Losses can make it impossible to equalize currents by means of increased coupling, but there is then no longer any problem with driven operation since the mutual resistance, as a result of these losses, is no longer in control of the situation. Because of this, it's customary to distinguish between directivity gain and power gain, the two being equal when there are no losses.

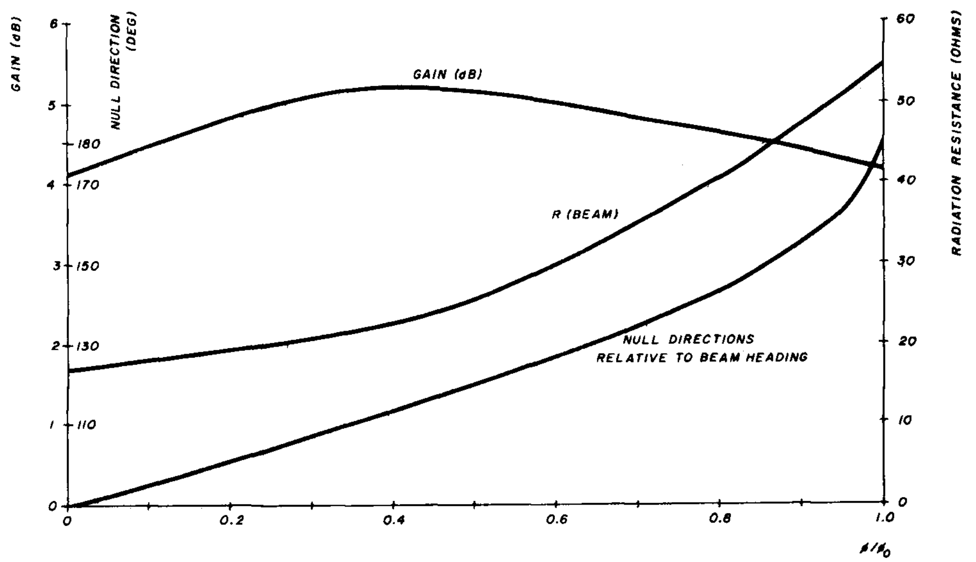


fig. 5. Relationship between phase angle, gain, null directions, and radiation resistance for all close-spaced, two-element beams with equal current drive. Note: losses not included; gain figures valid within 0.3 dB for null depths > 10 dB (approx.). Radiation resistance (in ohms) for half-wave elements spaced λ/8 (see text for other beams).

losses

Horizontal monoband hf beam losses rarely exceed a few ohms, and with radiation resistances of a few tens of ohms, can usually be neglected. In doubtful cases they can be roughly estimated from conductor sizes, assuming one has some idea of the current distribution.

Resistance data in graph form can be found in textbooks, but a handy figure to remember is one ohm per half-wavelength for 3-mm diameter copper wire (approximately No.10 AWG) at 14 MHz. The loss is inversely proportional to diameter and to the square root of frequency. However, divide by 2 for long conductors that have a sine wave current distribution, such as antenna wires or resonant feeders. This gives the resistance referred to a point of maximum current, which is standard practice also in the case of the radiation resistance with which it must be compared. For parallel wires, divide by the number of wires; for aluminum alloy, multiply by 1.6, and don't use iron or steel!

bandwidth

Bandwidth is roughly proportional to radiation resistance but also varies inversely with the length of the resonant system (which includes the antenna up to the point of matching). We are interested in two kinds of bandwidth:

- **SWR bandwidth** — i.e., the frequency range over which the SWR is less than 2.0. As might be expected by analogy with coupled circuits, bandwidth in this sense is improved by tighter coupling between elements. The better the SWR bandwidth, the less frequently the antenna tuner has to be readjusted, or the better the chance of being able to dispense with it. In general, SWR tends to rise steeply at the low frequency end of the tuning range, since tuning low is equivalent to tuning the reflector higher. This reduces ϕ, causing a shift to the left on the curves shown in **fig. 5**, with R_B dropping to a relatively low value.

- **Pattern bandwidth**, or the frequency range in which a specified null depth such as 10 dB is exceeded or the forward gain remains within 1 dB of maximum. Despite their relatively small size the antennas to be described here come close to meeting the above specifications on most bands without retuning the reflector. This is consistent with a reasonable degree of operating convenience, but to take advantage of very deep nulls it's essential for the reflector to be connected through its own feeder to a tuning device at the operating position. In this case, pattern bandwidth is less important but makes for added convenience. Use of two feeders provides an additional bonus: the ability to reverse beam direction.

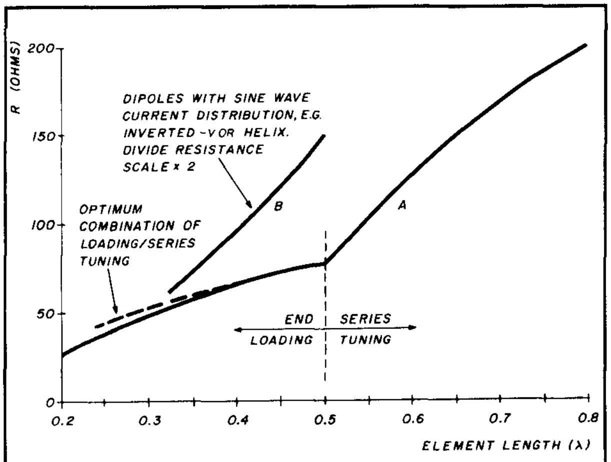


fig. 6. Radiation resistance of (A) "best possible" dipole elements (includes fig. 8C) and (B) resonant λ/2 dipoles fitting into restricted space (e.g., fig. 8D) — in this case, R is proportional to (1/length)^2.

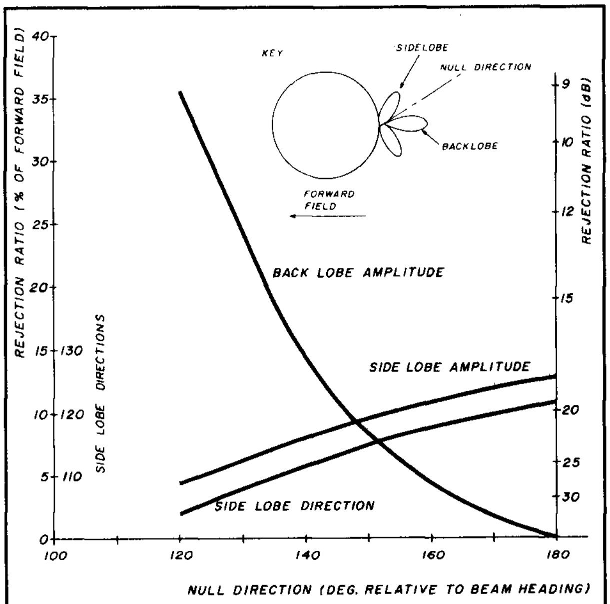


fig. 7. Size and direction of back lobes (see key, above). For corresponding gains, see *fig. 5*.

design of elements

Although elements don't have to be identical, it usually helps — and is essential if one wants to be able to reverse beam direction without having to retune. **Figure 6** shows that half-wave elements can be reduced by 30 percent in length for only a trivial reduction (17 percent) in radiation resistance, provided capacitive end-loading (or its equivalent) is used. **Figure 8** shows three practical ways of achieving this. In the case of **fig. 8C**, *AB* and *CD* act solely as end loading because radiation from them is cancelled by that from *AE* and *DE*.

Most of the designs to be described are based on the small delta loop (**fig. 8C**), which can be suspended between the tips of fiberglass fishing rods angled upwards, thus achieving an effective height considerably in excess of the mast height, since there's little or no radiation from the sides of the loop. In this form it has become known as "The Claw" for reasons obvious from the photograph (**fig. 15**). In most cases there seems to be little point in exceeding an element span of about 35 percent of the longest wavelength to be used; further size reduction is governed by three main constraints: bandwidth, losses, and difficulty of folding enough loading wire into the space available. For a reasonable approximation to "full size" performance, the above length can be more than halved if some form of remote tuning is provided.[1] Practical difficulties escalate rapidly as the span length drops below 0.25λ.

The arrangement shown in **fig. 8D** can be erected as an inverted V and is important as an alternative option though the sinusoidal current distribution halves the radiation resistance for a given span.

coupling and null depth

All four arrangements shown in **fig. 8** provide increased coupling. In cases A and D, this is readily adjustable by altering the spacing between ends; 30 inches for a span of 20 to 24 feet at 14 MHz has been found suitable, but some experiments may be advisable. Ajustment of coupling isn't critical. I've found that "design by guesswork" frequently produces null depths in excess of 20 dB. It's best for errors to be corrected in the antenna itself, but for fine tuning, placing capacitors between convenient points on feeders has been found satisfactory. These can be connected either in phase to increase coupling or out of phase to neutralize excessive coupling. I then find it possible to get null depths usually in excess of 20 dB, and often much greater, for all back directions and in-band (14 MHz) frequencies with single-knob tuning of

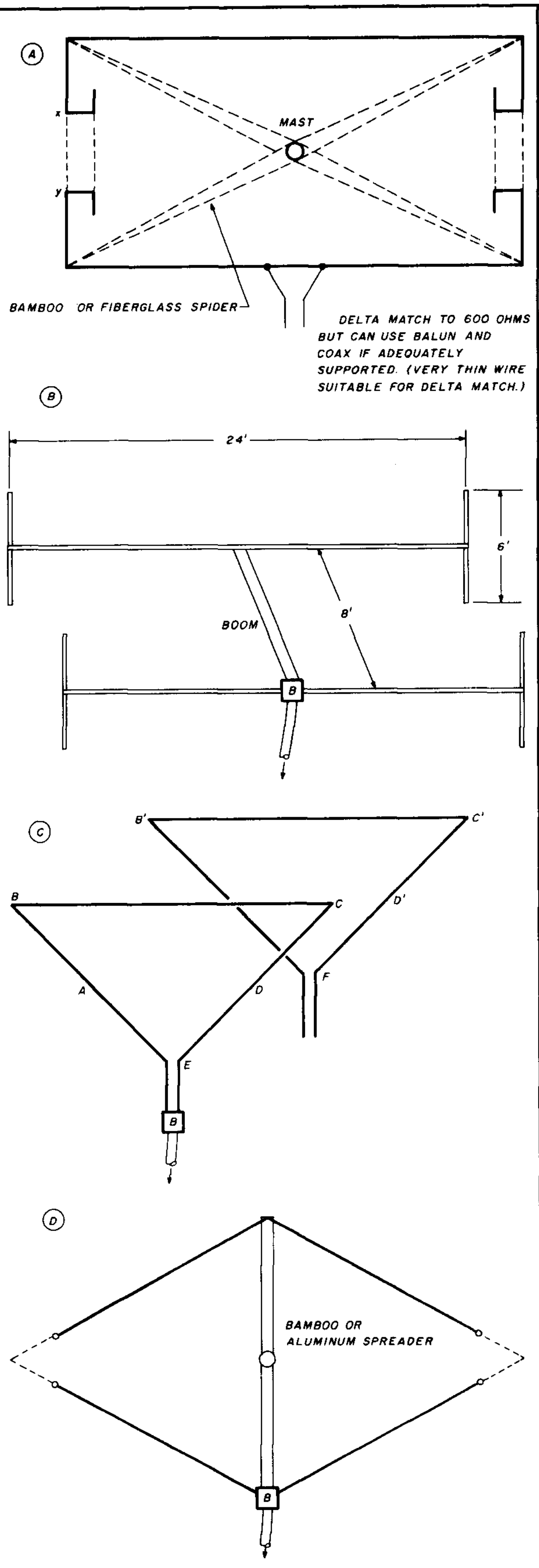


fig. 8. Two-element horizontal beams with reduced length and enhanced coupling. Reflector should preferably be a duplicate of the driven element (see text). Otherwise, if currents equal (i.e. if deep nulls are obtainable), tune reflector to the low edge of desired band. (A) Bent dipole elements (20 x 10 feet suggested, though dimensions are not critical) xy (nylon fishing line) ≈ 30 inches. (B) End-loading by vertical rods. Single elements have been used successfully with the dimensions shown. Coupling may need augmenting. (C) Small delta loops (ABCD) should be just over λ/2. Size can be reduced by small loading stubs. BB′ may be 0.12-0.2λ; EF<BB′/2. (D) Erect between posts or as inverted V. Spreader (or boom) may be 9-12 feet for 14 MHz.

the reflector. "Deepest possible" nulls aren't thought to be worth additional effort, given the unstable nature of the ionosphere and sensitivity to local field disturbances — for example, trees blowing in the wind, the presence of other antennas, even aircraft reflections. Up to 30 dB can be useful, but deeper nulls require precise adjustments of phase and amplitude (two knobs). By the time these adjustments are made, the interference one is trying to remove has probably disappeared anyway! With deep nulls a tendency has been observed for non-reciprocity between transmission and reception, due probably to pickup on wiring in the shack. If a linear is used, this won't be identical for the two cases.

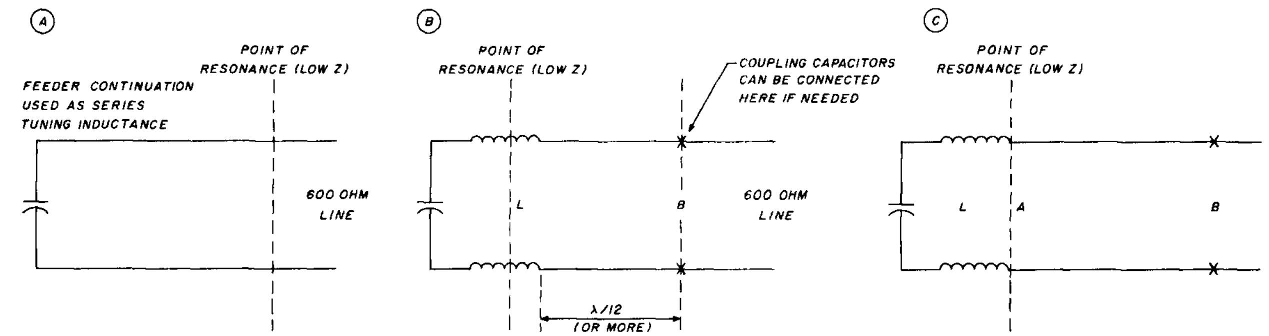


fig. 9. Reflector tuning of "Claw" antenna: (A) Long feedline; (B) short feedline; (C) resonant feedline. This method, especially as illustrated in (B), is convenient, since point B is more likely to be accessible. Total L = 3.5 μH. Terminations can be band-switched, but often cover two or three bands.

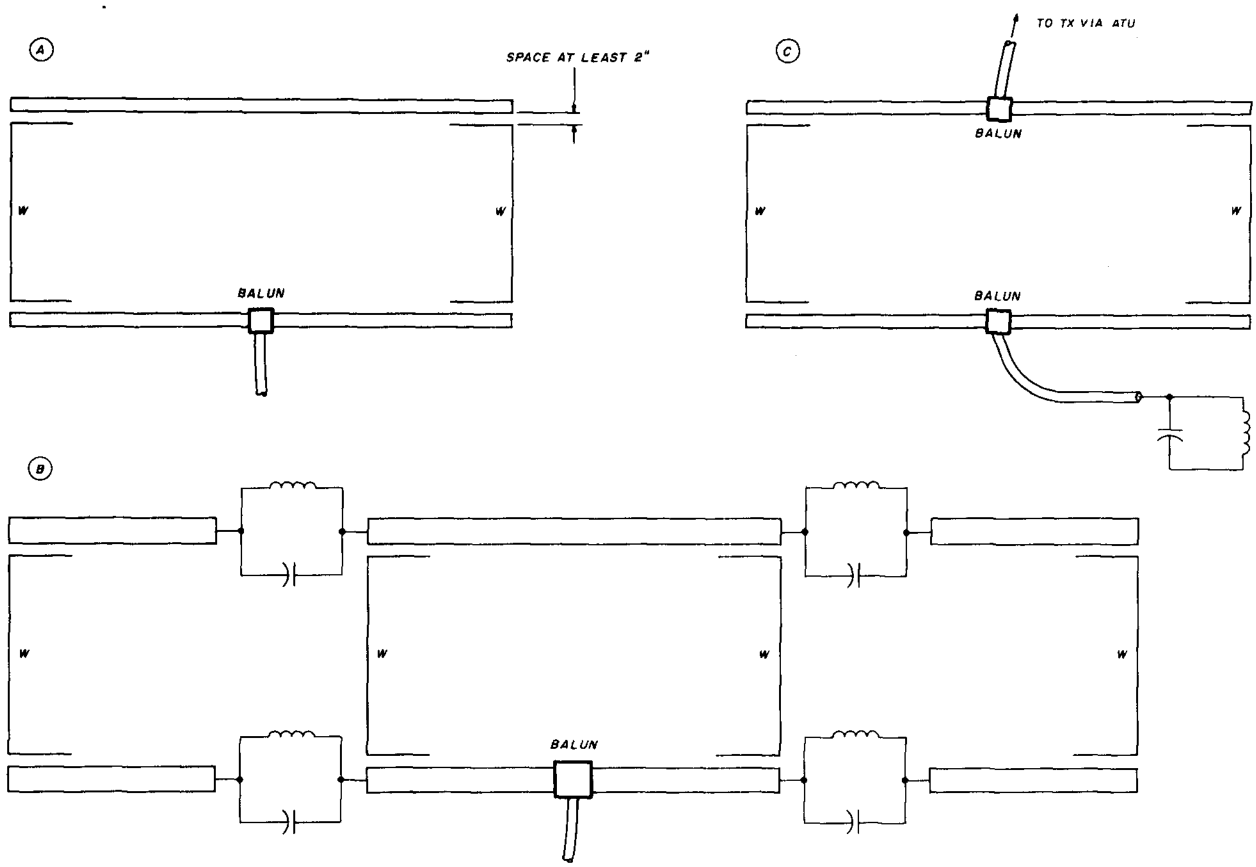


fig. 10. Application of enhanced coupling to conventional beams. (A) monobander — coupling wires (w) suggested as alternatives to bending the elements. (B) Beam with traps. (C) is similar to A, except in that the reflector is tuned via the feeder.

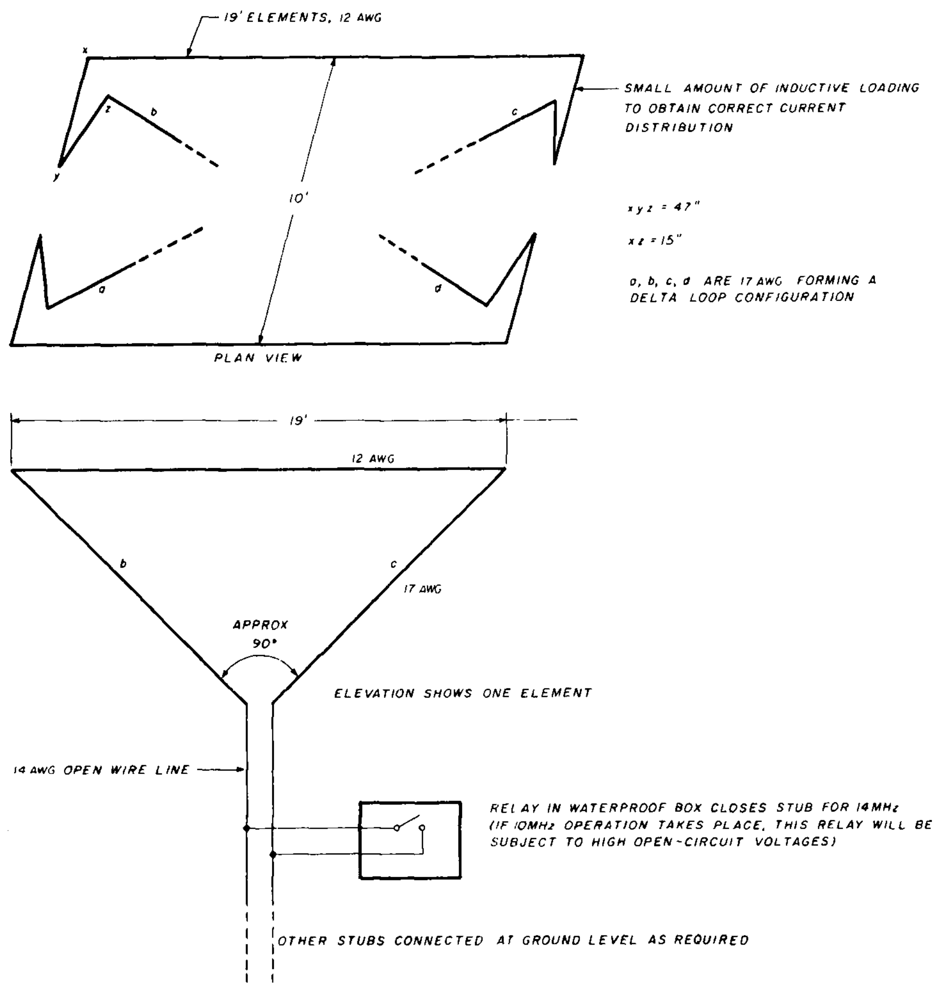


fig. 11. Wire beam based on the small delta loop. In plan view, this may be recognized also as a derivative of the VK2ABQ antenna. Wires should not be attached directly to bamboo supporting arms. Lengths of nylon fishing line should be used as insulators. (Reproduced with permission from RSGB.)

reflections from other antennas

These reflections can be large enough to seriously degrade forward gain at distances of over 30 meters at 14 MHz. (In one case I observed a loss of 2 S-units due to screening by another antenna at about 25 meters.) Their effect on front/back ratio is much greater. From mutual resistance data, it appears that another antenna 7 wavelengths away might be expected to degrade a 30-dB null by about 6 dB. Such effects depend, of course, on the extent to which the interfering antenna is in the beam path, so the effective radiation pattern varies with beam heading. The interfering antenna may be "removed" by detuning or rotating it to an end-on position, but any front/back ratio *it* may possess applies *only to its own* input terminals and does nothing to help. Unless such effects can be eliminated, the possibility for serious errors remains. This further emphasizes the importance of being able to make adjustments at the operating position to suit the needs of the moment. However good its f/b ratio, a large Yagi will still be wide open to reflected signals from other antennas in its beam path.

reflector tuners and t/r switching

Separate optimization of transmitting and receiving characteristics by switching between two different reflector tuners is a further important option. Tuning of the reflector to null out an interfering station can have a large effect on the SWR unless the antenna tuner is readjusted. Using an FS710H automatic in-line SWR meter, I haven't found this to be a problem, but without such a device and with modern solid-state amplifiers it can be an embarrassment. Because transmitting and receiving requirements aren't identical, it makes sense, in any case, to use separate tuners switched by a relay, not forgetting to use the "transmit" (tuner) condition when checking for prior channel occupancy. While any feeder length can be used, it simplifies matters to choose one that allows series tuning as shown in **fig. 9**, preferably as shown in **fig. 9B**, which increases the chance that points such as *xx*, where the rf voltage is suitable for coupling adjustment, will still be available in the shack if required. With feeder impedances of both 50 and 600 ohms at 14 MHz, I have found a capacitance (C) of 40 to 250 pF and an inductance (L) of 3 to 4 μH to be suitable, but this would depend on feeder length. The trick is to pick a likely-looking capacitor out of the junk box and, with a grid dip oscillator, find an inductance that allows the reflector to be tuned through the band and down to about 2 percent lower.

beam reversal

This reduces the average time required for changing beam heading since rotation can be limited to about 140 degrees. Also, in many cases an Armstrong method of rotation can be employed, with two ropes substituted for the usual heavy, expensive, and sometimes unreliable beam rotator. It can be particularly useful in multi-way contacts, and if there's uncertainty about whether propagation is short or long path. It also provides confirmation that the antenna is working correctly. **Figure 10** includes suggestions for the addition of controlled coupling to conventional beams which, though as yet untried, may be useful to experimenters. In (A) and (C), it may be easier to pull the ends in toward each other with nylon fishing line. To permit beam reversal and null-steering, the reflector must be replaced by a replica of the driven element.

comparison between two and three elements

Two-element beams can be expected to produce deeper and more controllable nulls, but this may not help if more than one interference source is present. In this respect, three-element beams are superior, assuming fixed tuning of the reflector in both cases. With two elements, a rejection ratio better than 18 dB can be obtained for all back angles if the nulls are placed at 150 degrees to the beam heading; this is for a spot frequency, but can be obtained throughout the band if reflector tuning is synchronized with the main tuning. With three elements, W2PV[2] found null depths in excess of 20 dB over 2 percent bandwidth in the best case, but this was only for the 180 degree direction. It has also been shown that three elements can provide rejection in excess of 28 dB at all back angles simultaneously on a spot frequency.[4] From tuning data given in this reference, I estimate a minimum rejection of around 18 to 20 dB over a 2 percent band, which is little better than the two-element result and requires very precise adjustment of the kind difficult to obtain in practice because one is working with too many variables. The two-element beam, moreover, provides the option of deeper nulls in specific directions, though the rejection of QRM from several sources simultaneously may then be adversely affected. In practice, not more than 1 to 1-1/2 dB extra gain can be expected from the third element, and in terms of the low-angle radiation required for chordal-hop propagation (which is probably responsible for most long-haul DX[5]), this is equivalent to an antenna height increase of only 15 to 20 percent. The small size, light weight, and construction of antennas described in the next section should, in many cases, lead to height increases of this order. It is not suggested that these small beams can compete with a six-element monobander on a 100-foot tower, but they can hold their own in more ordinary circumstances, requiring only a modest means of support, indifferent results quoted by earlier authors being attributable entirely to the assumption of straight elements and the resulting current inequalities. Much of this ground has been covered in earlier publications; a complete bibliography can be found in reference 5.

multiband beams

Multiband beams have previously consisted mainly of tribanders involving some degree of compromise. Instead of using the whole of a 14-MHz $\lambda/2$ element on 28 MHz, it is cut down to size by traps, or 28-MHz loops are stacked inside 14-MHz loops which could be used to provide extra gain on 28 MHz. This sacrifices 2 to 4 dB of gain at 28 MHz[1], as well as incurring losses due to circulating currents in traps or "wrong way" currents induced in unused elements, which can reduce bandwidth and affect coupling.

Trapped beams rate highly in terms of convenience as well as off-the-shelf availability, and their popularity could no doubt be further enhanced by design improvements — for example, along lines suggested by W0JF.[11] At the same time, the fact that now we have, within 1-1/2 octaves, six bands instead of three, presents an exciting challenge unlikely to be met by traps or stacking without further compromises. The log periodic antenna, though simple in use, is large, heavy, expensive, and inferior in performance to a Yagi of the same size.

the "poor man's log periodic" (PMLP)

When short of new ideas, it often helps to take a fresh look at old ones, which is what I've been doing in the case of resonant feeders. Though these have been *blamed* for TVI, radiation from balanced lines is very small, and they've consequently also been *recommended* as cures for TVI. It's true that they can get themselves twisted around the mast or entangled in guy ropes, but there's no excuse for this if beams are reversible as previously recommended, since rotation can then be restricted to less than 180 degrees. A more serious objection is the restriction of bandwidth. But with remote tuning, this becomes an inconvenience rather than a basic limitation. Now, through a fortunate accident, a solution to the bandwidth problem has also emerged and with it a family of small, lightweight antennas that provide the performance and other characteristics of the antennas described earlier, yet are tunable over the frequency range from 10 to 30 MHz. This largely achieves the object of the log periodic antenna, though the principle at work is entirely different.

the ideal antenna?

To establish the respectability of resonant feeders and provide a useful perspective, the design of a "best possible" multiband beam will be attempted.

Consider three identical straight tubing elements 44 feet long, spaced 8 feet apart, fed with about 46 feet of open wire line, and tuned to 14 MHz. The radiation resistance of a single element is 150 ohms and the bandwidth of antenna plus feeder for an SWR of 2:1 is about 8 percent, differing only slightly from that of a normal half-wave element. This should provide gain and f/b ratio in line with the previous three-element example. The larger value of radiation resistance, however, makes it much easier to use resonant feeders for remote tuning and beam reversal; even if this is not required, it helps to ensure reasonable bandwidth. The main benefit occurs at other frequencies, since on 28 MHz the elements are "extended double Zepp," which was the reason for the choice of length. The boom length is nearly optimum,[2] bringing the total gain to about 10.5 dBd. The spacing is too wide for good f/b ratio on 28 MHz, but this could be improved by additional coupling. On 10 MHz, omitting the center element provides for satisfactory spacing and achievable gain is in accordance with numbers indicated in **fig. 5**. Losses are negligible. The lower end of the feeders is assumed to be accessible near the antenna and can be lengthened or matched into, if necessary, utilizing band switching, depending either on local requirements or bandwidth considerations. With two elements, the boom length would have to be reduced, with 0.1λ the minimum acceptable length at 10 MHz. The gain at 28 MHz is reduced to 7 dB, but remains high for two elements. For comparison, the gain of a log periodic will be in the region of 6 dBd, assuming a boom length of 50 feet, making the PMLP far superior except for the inconvenience of having to tune it. This can be done in the shack if the feeders aren't too long. Nevertheless even the PMLP in this ideal form is in the "monster" class, and much effort has been devoted to applying the same principles to smaller antennas.

the small delta loop

The antenna shown in **fig. 8C**, a fixed pair of loops supported by a tree, was selected for this purpose after being used initially as the quickest and easiest way of getting back on the air from a new location. Though it wasn't possible to make a direct comparison, the performance of a full-size quad — in the same tree at the same height — was later judged to be about the same. The first multiband version of this antenna[6], shown in **fig. 11**, was used initially without the switched stubs, with correct coupling established by capacitance between the corner stubs (which also provided a small amount of loading) and by bringing the lower corners in towards the mast. The electrical lengths from top center to a shorting bar at ground level were arranged to be λ on 14 MHz, 1-1/2λ on 21 MHz, and 2λ on 28 MHz, providing for 600-ohm feedline matching as well. Current nulls in the center of the sides on 14 MHz ensured that all the radiation was from the top of the system, while at higher frequencies the loops radiated as ordinary loops, with effective spacing reduced by bringing the lower corners together. The matched lines (120 feet) were taken respectively to an antenna tuner and a series-tuned circuit for reflector tuning via a beam-reversing switch which interchanged the feeders in the shack. The elements were supported by four 8-foot bamboo garden canes radiating outwards and upwards from an aluminum hub with four 4-foot spokes, for a total radial length of 12 feet. The SWR bandwidth for given settings of the antenna tuner and reflector tuning at 14 MHz was only 100 kHz which generated a tendency to stay in one part of the band. Differences in tuning between wet and dry weather were a nuisance. These were aggravated by short lengths of 300-ohm line initially used for bypassing the rotating joint. Checking the feasibility of operation on 10 MHz was difficult because it required the attachment of matching stubs at half the mast height, but it was achieved on one occasion, which resulted in an S8 report from VK with reasonable f/b ratio.

Despite the narrow bandwidth, such an arrangement may well be acceptable on the grounds that it is simple, cheap, versatile, and efficient, especially if the shack is located at the base of the mast. For me, however, it has one fatal flaw: the difficulty of living with the knowledge that identical performance could be obtained from a smaller antenna! To improve bandwidth on 14 MHz, the stubs were moved a half wavelength closer to the loops, with relays to disconnect them as shown in **figs. 11** and **12**. This enabled tuning for the 21- and 28-MHz band, where the radiation resistance is much higher, to be carried out at ground level as before.

The small delta loop is mechanically superior to a center-fed bent dipole (**fig. 8A**), since the center of the element doesn't have to support the weight of the feeder. Other features include higher radiation resistance and reduced effective spacing on 21 and 28 MHz, but on 14 MHz an adverse impedance transformation takes place at the bottom corner of the loops so that an estimated radiation resistance of 45 ohms referred to the top center is stepped down to only 20 ohms in the feeder. Using the relay as shown, this isn't important, since only a short length of line is affected. But with no relay and No. 14 AWG feeders, there's an estimated loss of 1 dB, as well as the narrow bandwidth already noted. With tubing elements as shown in **fig. 12**, the situation is more favorable, with an estimated impedance reduction of only 30 percent. There's also more flexibility because the impedance discontinuity at the ends of the tubing tends to offset the step-down at the bottom corner and the boom can, if necessary, be used to support a heavier feeder system.

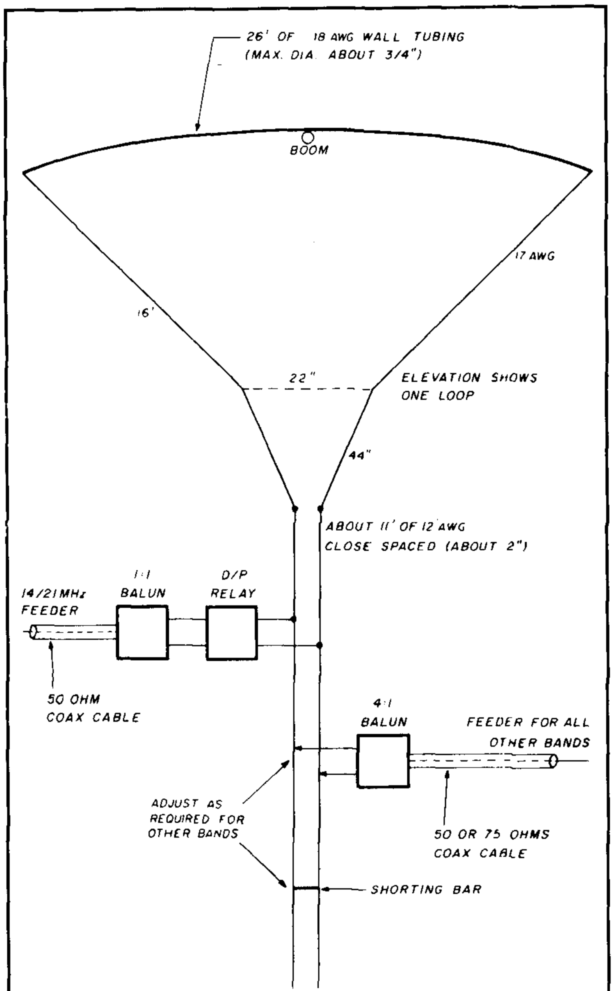


fig. 12. One element of small delta loop array using tubing end-fed with wires. These wires can be thin because current zeros occur at the crosspoints on 14 MHz, and on higher frequency bands Rg is relatively large. Element spacing is with the lower corners 4 feet apart. (Reproduced with permission of RSGB.)

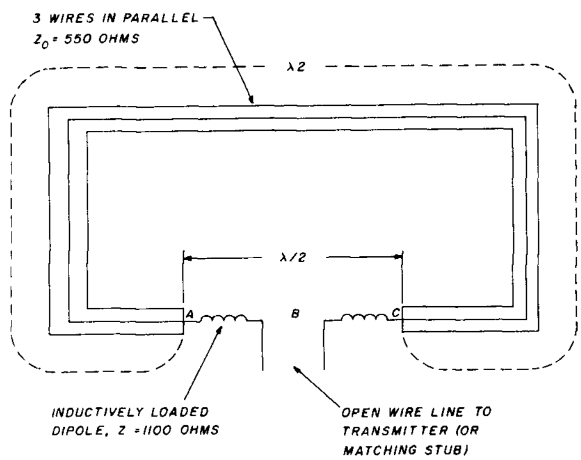


fig. 13A. Example of the impedance transforming loop showing principle of operation. (A) Loop acts as pair of λ/4 transformers so that a radiation resistance R appears (in this case) as 4R at the feedpoint.

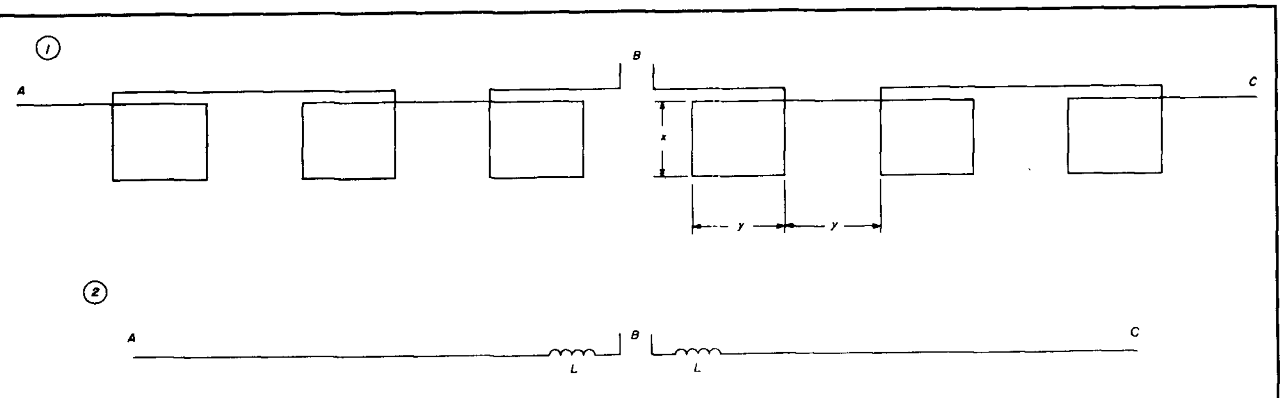
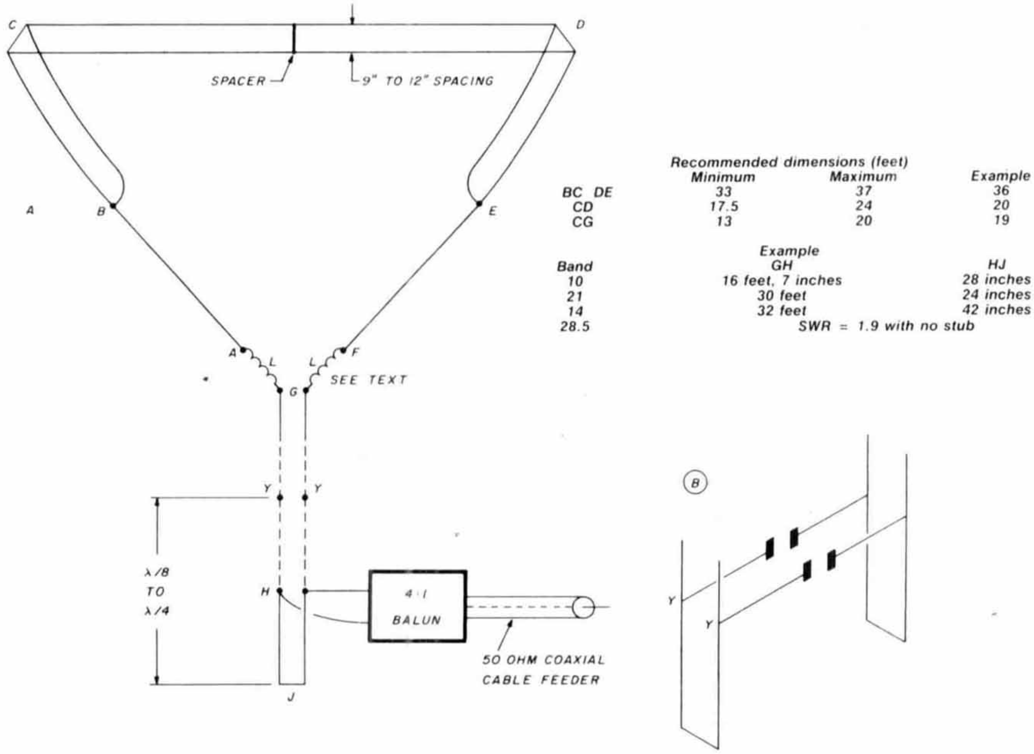


fig. 13B. Alternative forms of inductively loaded dipole. Note: in (1), x = 10 inches, y = 15 inches (at 14 MHz). In (2), AB = BC = 9 feet; each L is a 40-turn, 1-inch diameter, 20-inch coil.



	Recommended dimensions (feet) Minimum	Maximum	Example
BC DE	33	37	36
CD	17.5	24	20
CG	13	20	19

Band	Example GH	HJ
10	16 feet, 7 inches	28 inches
21	30 feet	24 inches
14	32 feet	42 inches
28.5	SWR = 1.9 with no stub	

fig. 14. Claw antenna design: (A) exact shape isn't critical (see fig. 15); (B) small capacitors are used to adjust coupling if necessary. (Reproduced with permission of RSGB.)

Figures 11 and **12** represent two extremes of design in which "anything goes." Typical observed SWR and f/b ratios are included in **table 1**. No additional coupling or neutralization was needed to obtain the results shown. In the case of the **fig. 12** configuration, it must be assumed that although on 14 MHz the radiation was coming from straight elements, the "quad loop effect" was operative with respect to coupling. In the antenna shown in **fig. 12**, it was found that for two-band operation (14 and 21 MHz), the relay could be omitted because of a chance combination of impedance transformations which caused the second harmonic resonance to occur at only 1-1/2 times the signal frequency. In general, I've not found it difficult to obtain efficient two-band operation of antennas without switching, but three bands are much more difficult.

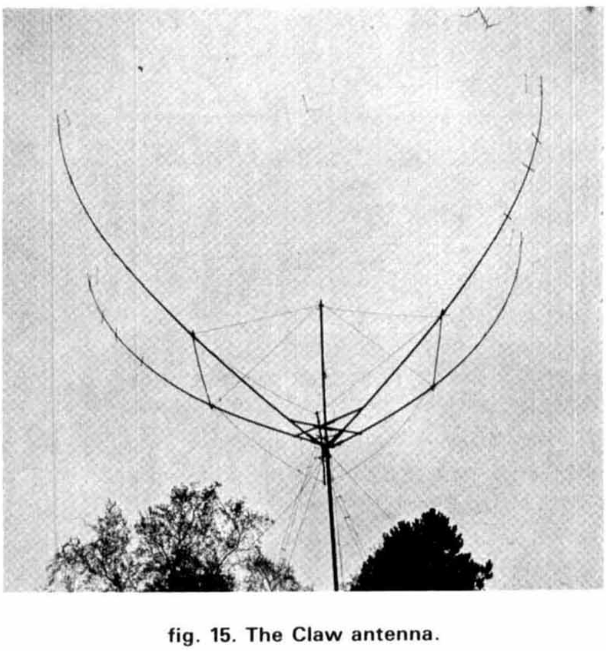


fig. 15. The Claw antenna.

the impedance transforming loop(ITL)

Disadvantages of the systems illustrated in **figs. 11** and **12** include the need for switches or relays in positions that are usually inaccessible. Even if this is acceptable, frequency coverage is restricted because as R decreases, switching devices have to meet increasingly stringent requirements with respect to capacitance and rf voltages.

Figures 13, 14, and **15** show a means of dispensing with relays and, to a large extent, the need for matching stubs. This was the outcome of an unsuccessful attempt to develop a small (i.e., 18-foot, 14 MHz) broadband folded dipole by slowing down the wave velocity.[8] The idea was to trick the wave into thinking

Table 1. Comparison of various antennas described in this article.

BAND (MHz)	Antenna	Bandwidth for F/B ratio >20dB	>15dB	>10dB	Bandwidth for SWR <1.25	<1.5	<2.0
14	S.D.L. (wire)		140	280	110	190	300
	S.D.L. (tube)	150	220	385	225	390	
	Claw No. 1	40	120	220	160		330
	Claw No. 2		145	300			230
	Folded Dipole			230	100	260	350
21	S.D.L. (wire)			320	110	205	370
	S.D.L. (tube)			550	360		
	Claw No. 1	85	210	350	330	420	
	Claw No. 2			400		320	380
	Folded Dipole			260			450
28	S.D.L. (wire)				190		500
	S.D.L. (tube)			800	550	800	
	Claw No. 1		240-300	600	180	370	680
	Claw No. 2	170	840	1000		1000	
	Folded Dipole					280	1000

the element was larger, but this was unsuccessful. Being difficult to draw and of limited practical interest, its somewhat fearsome appearance will not be inflicted on the reader, though some performance figures are included in **table 1**. The surprise came in the form of a chance discovery that long feeders could be connected without degrading the bandwidth; the explanation, though elusive, led eventually to the design of a number of antennas bearing little resemblance to the original dipoles.[7,9]

principles of operation

Figure 13A illustrates a small loop element which could be any shape. Two or more half-wave wires are used in parallel for the top part of the loop, resulting in a low value of characteristic impedance Z_{OT}.

The remainder of the loop consists of a second $\lambda/2$ dipole with a high value of characteristic impedance, Z_{OB}; this can be a helix as shown, or inductively loaded in other ways. Each dipole functions as a $\lambda/4$ transformer so that the radiation resistance, R, after being stepped up to the value

$$Z_{OT}^2 / R$$

at the ends of the top dipole, is then stepped down to

$$(Z_{OB} / Z_{OT})^2 \cdot R$$

at the feedpoint. As illustrated, a typical R value of 50 ohms is stepped up to 200 ohms at the feedpoint, which is high enough to ensure that bandwidth remains an intrinsic property of the antenna and is free from serious degradation due to the feeder. A selection of Z values is given in **table 2**. Depending on size and construction, the lower dipole may be a thin wire unloaded V or one of the alternatives shown in **fig. 13B**. All of these arrangements have been used successfully. The usual objections to inductive loading don't apply because the radiation is mainly from the top part of the loop. This comes about because the current is stepped down in the ratio of the impedances; because the lower dipole is shorter; and because the current distribution in it is sinusoidal or triangular, in contrast to the almost uniform current in the top dipole. This constitutes a major advantage over the quad loop, in which the mean height is dragged down by radiation from the lower side. On the other hand, there is some radiation from the sides; it can be cancelled

Table 2. Design data for ITL antennas.

Characteristic impedance for various wire sizes and combinations.

Number of conductors	Diameter (inches)	Spacing (inches)	Z_0(ohms)
1	0.04		1000
1	0.8		650
2	0.04	4	690
2	0.04	6	640
2	0.04	12	550
3	0.04	4	550
4	0.04	4	490

Equivalent Z_O of lower dipole (wire diameter, 0.04 inches) with inductive loading.

Physical length of dipole (wavelengths)	Equivalent Z_0 (ohms)
0.375	1200
0.3	1500
0.25	2000
0.2	29000

Note: Z_O values are calculated for 14 MHz. However, because of some length dependance, they will be slightly different for other bands. No data is available for helical windings.

by reverting to a more or less triangular shape as shown in **fig. 14**. Assuming an ITL to be designed for 14 MHz, operation on the higher frequency bands differs little from that of the small delta loops described earlier; at 21 MHz there tends to be some "wrong-way" impedance transformation, suggesting the desirability of matching stubs at ground level in the case of long feeders. Concentrated loading, as shown

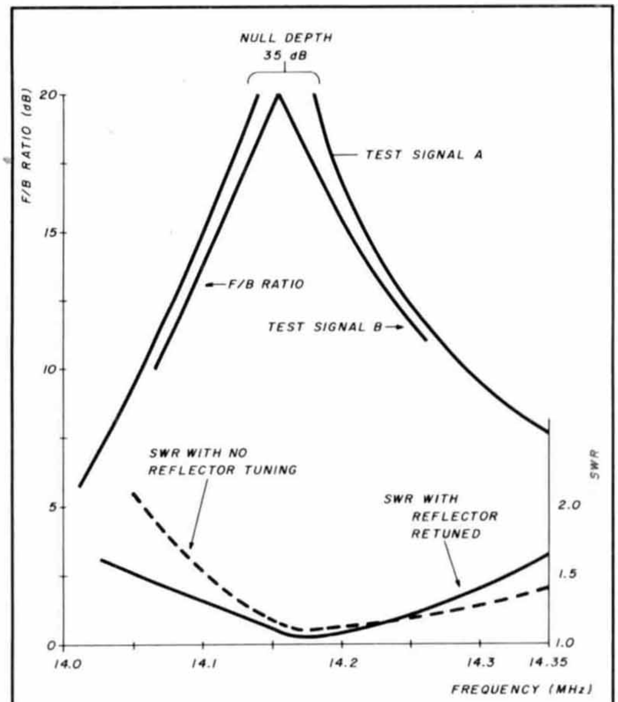


fig. 16. Typical performance of Claw No. 1 on 14 MHz. F/B ratio curves demonstrate null-filling due to any slight error in adjustment. In this case, one test signal was slightly too close. SWR rapidly increases as the π approaches zero. The reflector was adjusted for nulls at 140-150 degrees, but curves were roughly repeatable over a range of 110-180 inches without readjustment of coupling.

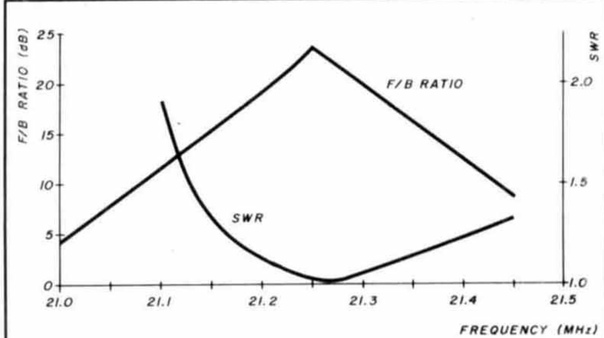


fig. 17. Typical performance of Claw No. 1 on 21 MHz. Note steep rise in SWR at low frequencies — i.e., as ϕ decreases.

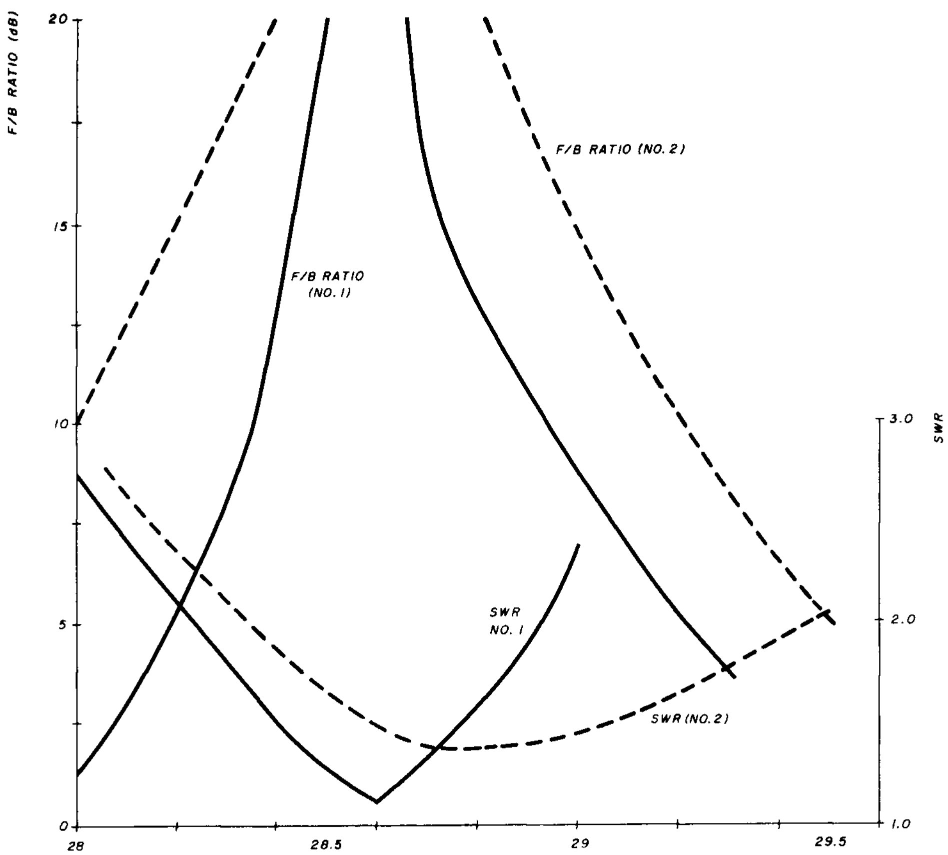


fig. 18. Typical performance of both Claws on 28 MHz. Curves demonstrate adverse effects on bandwidth of very long unmatched feeder.

in **fig. 13B2**, improves matters on 14 MHz by increasing Z_{OB} (see **table 2**), but reference to a Smith Chart suggests that at 21 MHz this could lead to an increase of SWR in the open-wire line of 2 or more. On the other hand, with long coils of small diameter (20 inches x 1 inch), as shown in **fig. 14**, much better agreement between theory and practice resulted from regarding them as a harmless continuation of the traveling wave system.

construction

Figures 14 and **15** illustrate the latest version of the Claw antenna, which uses two pairs of fiberglass fishing rods 13 feet long, extended at their lower ends by an additional 6 feet of 1-inch diameter fiberglass tubing. These are plugged into alloy sockets which radiate outwards from the top of the mast. They are braced back with further lengths of fiberglass tubing to a short mast extension. The elements are held apart by 6-foot spacing rods of 1/2-inch diameter alloy tubing. Plastic rod end-pieces are used to keep the rods a few inches clear of the elements; even so, these may be responsible for some of the coupling. The tips of the fishing rods are pulled in by nylon fishing line to give an element spacing of 12 feet. Points on the rods are guyed back to crosspieces at the top of the mast. The top wires are held 11 inches apart in the horizontal plane by fiberglass spacers cut from the discarded tips of the fishing rods. Additional spacers on the rods themselves (with fishing-line ties) are used to maintain even tension in the wires to avoid flexing and breakage. Earlier versions used three copper wires spaced 4 inches apart, but the benefit from the extra wire hardly justifies the added complication (see **table 2**). The latest version uses No. 16 AWG aluminum alloy wire, which reduces weight for a given rf resistance, but increases windage. Since the rf resistance is only half that of a single wire, mechanical considerations are more likely to be the deciding factor. Wires break if not kept under tension, but with two antennas over a period of three years — which has included periods of heavy winds — no fishing rods have broken, and there has been no other damage to the main structures.

The same wire gauge is used throughout. The loading helices are each wound with 40 turns over a total

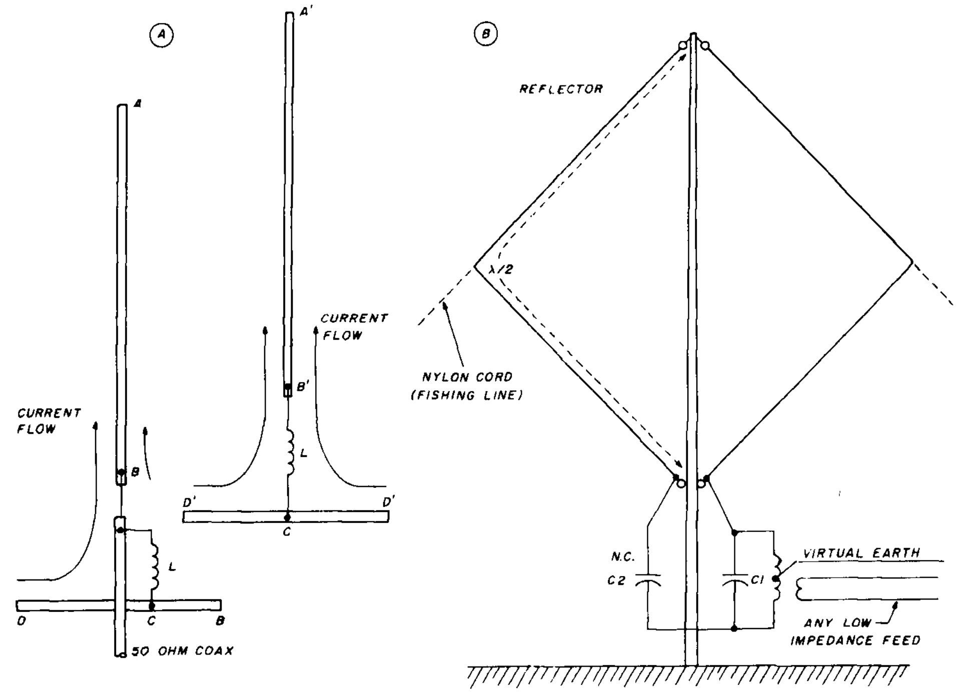


fig. 19. Vertical arrays. (A) Asymetrical dipolr array can be mounted on fence posts. To adjust coupling, swivel around to alter spacing DD'. (B) Maypole array.

length of 20 inches, partly near the lower end of the rod extensions and partly on the bracing struts. Because of the low radiation resistance on 10 MHz, it's advisable to provide matching as close as possible to the antenna. In one case, 28-inch stubs were placed 16 feet, 7 inches from the element. Later, for greater convenience I used a pair of series-connected 10-pF capacitors near ground level. The location could be determined by finding points of "zero" current and then moving 3 feet closer to the antenna. Matching on the other bands was used initially, but discarded because it made no difference in signal strengths, though **table 1** suggests the loss of some bandwidth. Even on 10 MHz, despite the additional 120 feet of open wire line (No. 19 AWG), the loss without matching was less than one S-unit. Measured performance data for both Claw antennas is included in **table 1.** The plots of f/b ratio and SWR shown in **figs. 16, 17** and **18** are typical of results obtained with Claw No. 1.

alternative designs

The loops can also be suspended from spreaders between two supports. In this case planar loading (an idea suggested to me by Steve Hart, VK5HA), as shown in **fig. 13B**, is suggested. The assembly can be supported by three lengths of nylon fishing line with small ring insulators cut from fiberglass tubing to keep adjacent edges apart. I find that nearly a 3:1 reduction in length can be achieved this way; because it uses less wire, its efficiency is greater than with helical loading. For this reason, it was used in the first two versions of the Claw,[7] but helical windings in this case are easier and neater, with losses insignificant. Similar loops can be suspended from their centers in inverted V fashion. I've also built a rotary version of such an antenna modelled after the one shown in **fig. 3B**. This used a lightweight mast extension surmounted by a 1/2-inch diameter aluminum boom. Two fiberglass radial arms were used to hold up the dipole ends. Apart from the neutralization problem mentioned earlier, this worked well on 14 MHz. A three-element version of the Claw was also constructed; since only triband operation was required, I was able to use a coaxial feeder for the center element and a relay to switch in an additional length of helix on 21 MHz. The third element was effective on 28 MHz and was indirectly useful for 10 MHz because, though not in use, it allowed wider spacing between the other two elements without degradation of performance on 28 MHz. On 14 and 21 MHz, there was no improvement compared to using any one of the three possible pairs on its own. SWR on 14 MHz could be varied between 1.0 and at least 5.0 by tuning the parasitic elements! The problem was basically one of "too many variables," and it was concluded that for three elements to be viable, they would need to be spread out along a boom; the "Claw" concept would then lose most of its attraction in return for a relatively small improvement in performance.

vertical arrays

Vertical beams can also be constructed using controlled coupling.[10] **Figure 19** shows two examples. The first is an "asymmetrical dipole" array that uses inductively-loaded counterpoises to form the lower half of the dipoles. The inductances can take the form of linear loading.[5] The elements are movable and can be plugged into sockets on fence posts. Coupling is varied by rotating the counterpoises towards each other and all adjustments are conveniently accessible. Each half of the dipole should be resonated separately against a $\lambda/4$ wire. The second can be regarded as a "vertical VK2ABQ" antenna. It's best to use four wires at right angles; adjacent pairs may be connected in parallel, though they can also be used in a three-element configuration. In the two-element case, which is recommended, overcoupling was experienced, requiring neutralization as shown. ("Zepp feed" can also be used, provided the open end of the feeder is closed with a $\lambda/4$ stub[5,10] as recommended by G6CJ.)

conclusion

My intention has been to provide guidelines, rather than blueprints, for the construction of antennas tailored to suit individual needs. The Claw designs will be useful even if the best mast available is only a garden post, and I hope that some who have decided regretfully that beams are "not for them" will have second thoughts. The null-steering and beam-reversal capabilities are particularly useful. In addition to coverage of six bands — with "monoband" performance on several — Claw elements are particularly suitable for use as top-loaded verticals for the lower-frequency bands.

references

1. L. A. Moxon, G6XN, "Small Beams with Big Performance," *ham radio*, March, 1979, page 12.
2. J. L. Lawson, W2PV, "Yagi Antenna Design," *ham radio*, May 1980, page 18.
3. Bloch, Medhurst and Pool, "A New Approach to The Design of Super-directive Aerial Arrays," *Proceedings of the IEE*, Part III, September, 1953.
4. W. Walkinshaw, "Theoretical Treatment of Short Yagi Aerials," *Journal of the IEEE*, Part IIIA, March-May, 1948, page 598.
5. L. A. Moxon, G6XN, "HF Antennas for All Locations," *Radio Society of Great Britain*, 1982.
6. Pat Hawker, G3VA, "Technical Topics: Multiband Arrays with 'KISS,'" *Radio Communication*, August, 1983, page 705.
7. Pat Hawker, G3VA, "Technical Topics," *Radio Communication*, March, 1985.
8. Pat Hawker, G3VA, "Technical Topics: Antennas That Slow The Wave," *Radio Communication*, April 1984, page 316.
9. Pat Hawker, G3VA, "The 'Claw Mark 4' Antenna," *Radio Communication*, January 1987, page 28.
10. F. Charman, G6CJ, "Aerial Reflection," *R.S.G.B. Bulletin*, December, 1955.
11. Private communication.

G6XN's book, *HF Antennas for All Locations* is available from Ham Radio's Bookstore for $11.95 plus $3.50 S&H.

ham radio