Happiness is understanding trap-antenna design, and whether to make or buy.

The Multi-Band Trap Antenna—Part I

BY JOSEPH M. BOYER*, W6UYH

For one reason or another I have had occasion to mention to a number of amateur friends that the familiar multi-band "trap" antenna was invented by a clever radio engineer named Howard Morgan[1] just before the onset of WW II. Most of them were quite surprised to learn that the concept was that old. However, one friend burst out laughing and said, "Invented! What was there to invent?" I asked what he meant. "Why I didn't know you could get a patent on anything that simple," he said. It's just a colinear collection of half wave doublets or grounded monopole antennas, each cut to work in a different band and all *insulated* from one another by parallel resonant circuits!"

Is Morgan's electromagnetic brain child really all that simple in concept? Are each of its radiating sections just self resonant antennas, oscillating naturally on their respective frequency bands? Do those lumped LC parallel resonant traps actually *open* and *close* like frequency controlled *switches* along the antenna as you flip the rig from one band to another? The truth of the matter is that the Morgan trap antenna is better described by the words of the old song which tells how "the foot bone's connected to the ankle bone, and the ankle bone's connected to the leg bone . . .". If its design is approached by conventional methods of antenna analysis, it can give a headache even to a trained professional. This is unfortunate, because the trap multi-band antenna is extremely convenient to use in a cramped space QTH for all-band h.f. operation. Newcomers to amateur radio wish to understand at least its key features in order to intelligently compare the merits of several commercially available models on the market; experienced amateurs may want to apply certain of its basic theory principles to a special radiator of their own conception. Rather than let its true mode of operation remain a matter of doubt in amateur circles and its design a thing of exhausting cut-and-try while sweltering out in the hot sun, it might be interesting and useful to reduce the trap antenna design to clear cut simplicity.

Not only that, but while we are at it let's not make this a matter of following some other fellow's step-by-step blue print description of the DX Band Hopping Skyhook; instead let's use a general design approach so we can juggle our own available capacitor sizes, our own conductor stock to end up with a radiator which fits our own particular needs. Finally, let us do the job in such way that while its design is still on paper—long before we feed it r.f. power—we will know how the antenna is going to perform; what its inherent performance limitations are, and why things turned out that way.

Polarization—Horizontal Or Vertical

Right at the start we have a choice of using a horizontal trap doublet antenna or a trap vertical monopole working against its electrical image or ground. The choice we make really doesn't matter technically; any grounded monopole antenna of

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[1] H. K. Morgan, "A Multi-frequency Tuned Antenna System," *Electronics* vol. 13, page 42, August, 1940

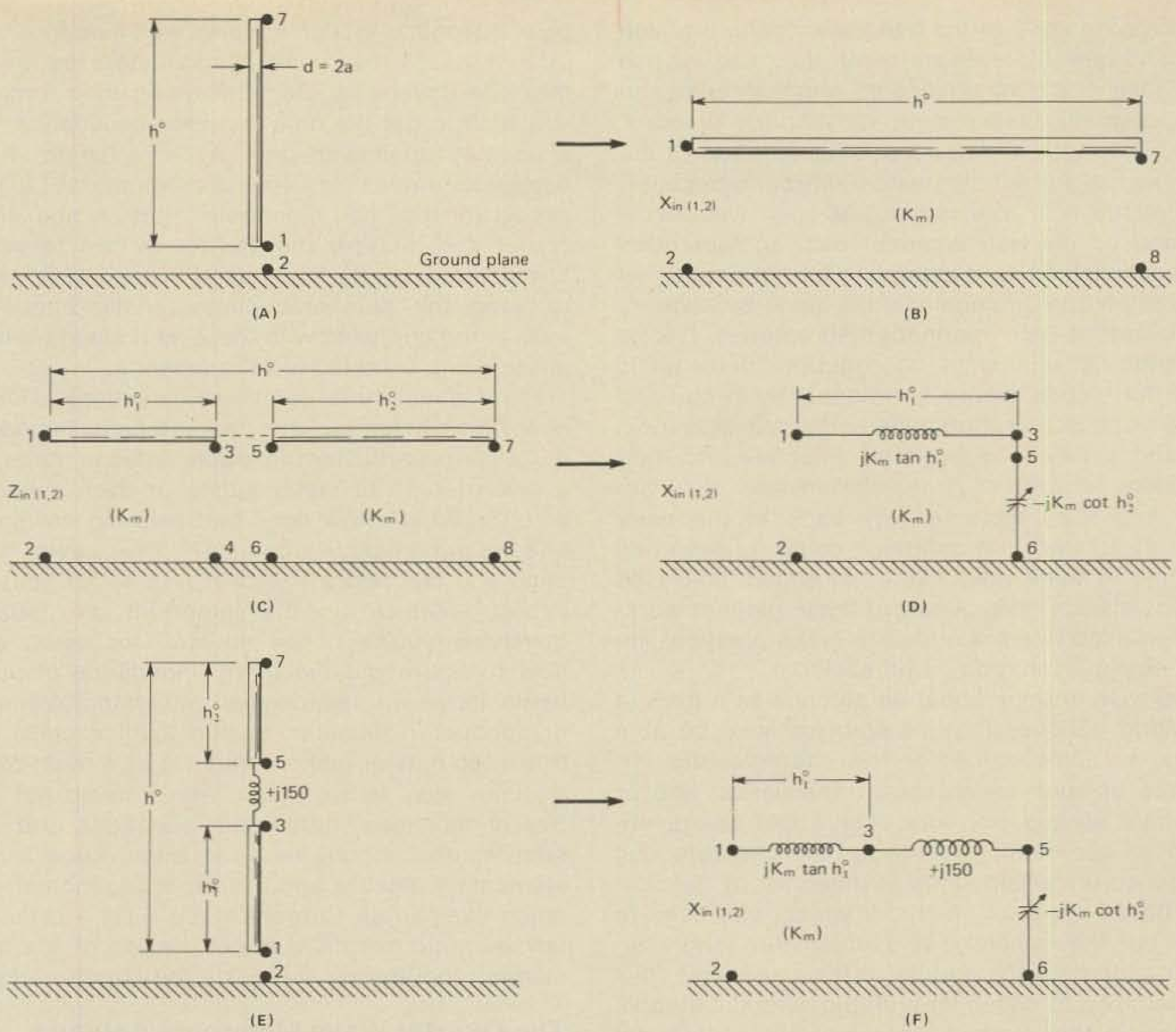

Fig. 1—(A) Naturally resonant λ/4 monopole antenna and (B) its analogue r.f. transmission line. (C) R.f. transmission line stub broken into two sections and (D) its equivalent "circuit". (E) Morgan two-band trap monopole series loaded with out-of-resonance trap operating on a lower frequency band. (F) Analogue r.f. transmission line representation of same Morgan trap antenna.

length h is merely one half of the equivalent doublet form having a total length 2h and located high above the earth. Therefore, when we go through the process of finishing the design of a multi-band trap monopole of length h and conductor radius "a", all we need do to convert it into its equivalent free space doublet is to build another duplicate monopole—complete with identical traps—and connect the two "half elements" on each side of the center insulator. Here, we will use the vertical monopole only because it makes our discussion and the drawings related to it easier to follow.

Now, of course, if we contemplate going the whole route and including coverage of all h.f. amateur bands from eighty to ten meters, well then we might have good reason to think about polarization choice for a moment. Any good antenna handbook has diagrams showing the radiation patterns in the elevation plane of horizontal doublet antennas at various heights above ground. Inspection of such radiation patterns quickly shows that amateurs at least face a real world problem in using horizontal h.f. antennas. Only at heights of about 0.5 λ or more above ground does the maximum amplitude response point on the major radiation pattern lobes of the horizontal doublet get near enough to the horizon to consistently produce decent DX performance. At eighty meters that 0.5λ height means about 130 feet. Few urban amateurs have available the space, facilities, and freedom from local height ordinances to erect an eighty meter doublet that high in the air. Conversely, a grounded vertical λ/4 monopole on any ham band produces (a) omnidirectional response in the horizon plane and (b) its elevation plane pattern yields maximum gain quite close to (but never right on) the horizon. Now let's turn to the receiving noise problem in terms of antenna polarization... oops, sorry! That is an entirely different subject; another time, perhaps.

Pulling The Fangs Of The Beast

Our first move in the design game we are to play is to remove the *teeth* of the general antenna problem so it can't bite. By *teeth* we mean the need for use of the advanced mathematics on which antenna and electromagnetic theory are quite properly founded. Luckily, we can make a great simplification in the math used in a practical design approach because, at about the time that Howard Morgan was nicely wrapping up his trap antenna idea, another good man was finishing an outstanding job on an elegant theory related to *all* antennas; his name is Sergei A. Schelkunoff of Bell Telephone Laboratories. Doctor Schelkunoff's significant contribution to us all is called the "mode theory of antennas"[2]. In carrying out his work, Schelkunoff pulled off a neat side trick: he found a way to convert *all* antennas into their equivalent form of *r.f. transmission line*. Now antenna engineers working way back in the early 1900's suspected that antennas could be regarded as acting in some ways like open ended, stub type r.f. transmission lines. A few of these pioneer workers even used this basic idea to make practical antenna design in that day a bit easier.

If we wish to think about an antenna as a form of oscillating transmission line stub we must be able to plug in some value for the *characteristic impedance* of such an antenna/transmission line or the whole idea is not very useful. Old timers obtained an approximate value for the characteristic impedance of certain kinds of antennas by calculations based on a d.c. method, which was used to figure out the capacity of a one-meter long conductor section in the middle of their antennas; this d.c. method was called logarithmic potential theory[3]. Not only is this method laborious, but it does not account for all types of r.f. waves which really exist on antennas, just the d.c. or static mode. Still, the idea was very helpful.

Schelkunoff, however, found a way to get an *average* value of characteristic impedance representing *all* the wave modes existing on various types of antennas. Not only that, but the formulas derived by this most eminent theoretician are expressed in ordinary, every day mathematics familiar to us all, not just people with a Ph.D in physics. Although Schelkunoff worked out formulas for the characteristic impedance of antennas having many different kinds of cross sectional conductor geometry, his formula for a monopole antenna made from a conductor having a uniform radius "a" and length h is just,

$$K_m = 60 \left[2.306 \log_{10} \frac{2\,(h)}{(a)} - 1 \right] \text{ OHMS} \quad (1.0\text{-}1.)$$

The above formula gives the average characteristic impedance K_m for a grounded cylindrical monopole antenna, although it works well also for square or triangular cross section conductors as long as they are *uniform* in diameter along their length. If you wish to get the characteristic impedance K_a of a doublet antenna in free space of length 2h and conductor radius "a" you use equation (1.0-1.) to get K_m for the "half monopole" part on one side of center and multiply the answer by two to get K_a. The notation K_m (or K_a) is used instead of Z_o merely to keep the antenna's characteristic impedance from being confused with the Z_o of a standard transmission line used to feed the antenna.

Right about here someone may think, "Ok, so now I can calculate the characteristic impedance K_m of some particular monopole antenna made from a given length of metal tubing or even a hunk of wire. So what? How does that help me design and understand amateur antennas?" The answer is that once you can find a K_m (or K_a) value for your *particular* antenna, all the important and puzzling questions you've asked yourself for years about how to figure out the input impedance of an antenna, its v.s.w.r. response as you change frequency or conductor diameter, or add loading coils, band traps, top hats or make it perform as a multi-bander become easy to work out. And you do not need Bessel functions, differential equations and other super-math cannons to do it, either. Just a bit of elementary algebra and a dash of trigonometry. It's much like "Brain Surgery Made Easy." Let's try a few *warm up* exercises to get the feel of this handy antenna tool before we tackle the Morgan antenna.

The Quarter Wave Monopole Antenna As An R.F. Transmission Line

The naturally resonant quarter wave vertical monopole antenna working against ground forms an ideal *launching pad* for our design attack on the multi-band trap antenna. A monopole which is naturally resonant attains this freely oscillating state solely by means of its shunt capacity distributed along its conductor to the ground plane and the distributed series inductance of the conductor. (In doublets, this distributed shunt capacity is from one doublet conductor half to the other.) A typical amateur problem related to the $\lambda/4$ monopole can be stated in two related parts as follows:

(1) We wish to operate a $\lambda/4$ vertical monopole across the entire frequency width of the eighty meter band, which means an upper frequency band limit f_{high} of 4.000 MHz, a band center frequency f_o of 3.750 MHz, and a low band limit of 3.500 MHz. If such monopole is naturally resonant at f_o, what v.s.w.r. will the monopole's input impedance produce in a fifty-ohm coaxial cable at f_o, f_{low} and f_{high}?

(2) How will the cross section diameter $d=2a$ of the conductor used in such monopole influence v.s.w.r. at the stated frequencies?

We can start by choosing monopole conductors

[2] S. A. Schelkunoff, "Theory of Antennas of Arbitrary Size and Shape," IRE Proc., *29*, pp. 493-521, September, 1941

[3] Edmund A. Laport, "Radio Antenna Engineering," page 28 and chapter 6, McGraw-Hill Book Company, Inc. First Edition.

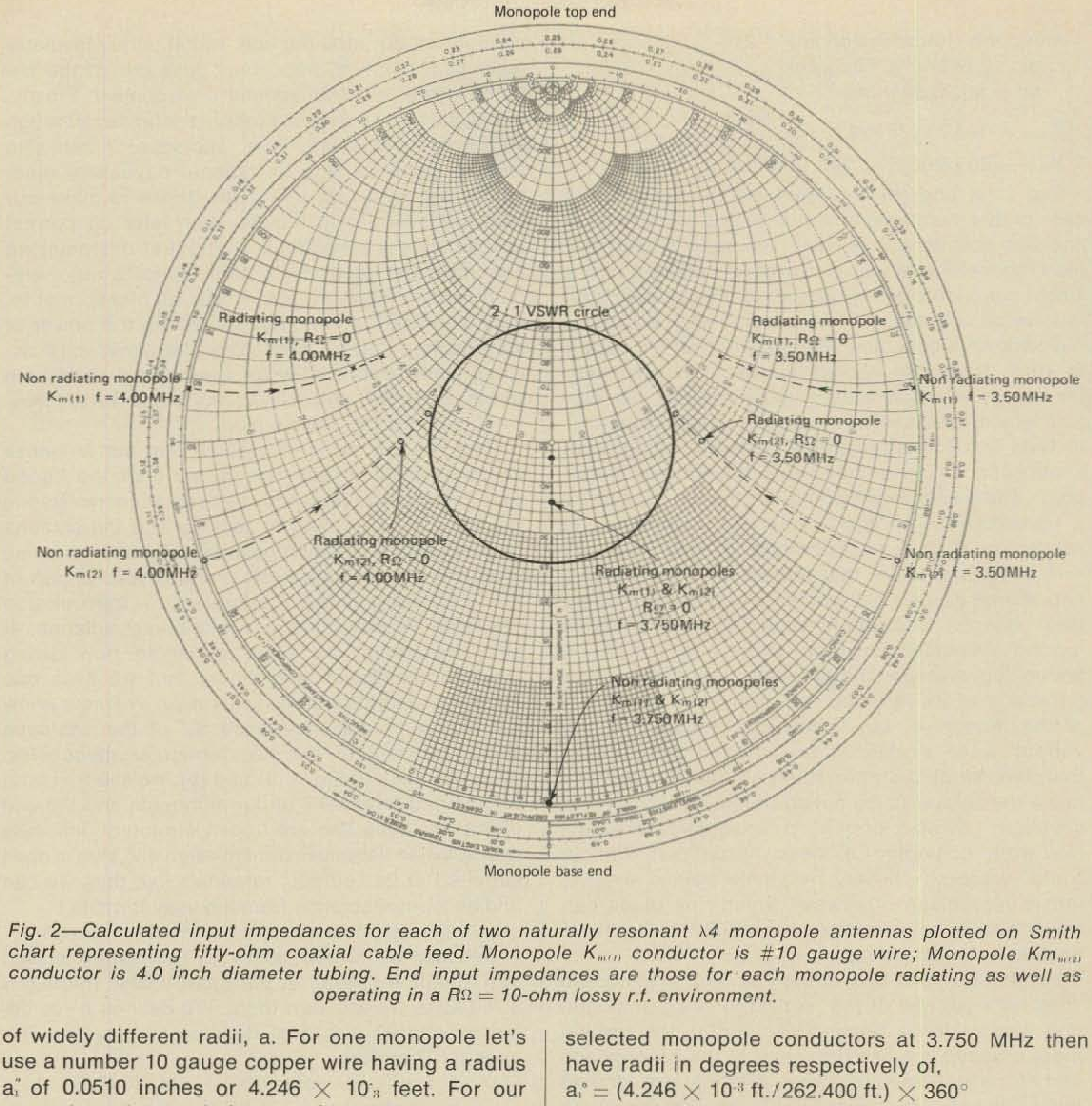

Fig. 2—Calculated input impedances for each of two naturally resonant λ4 monopole antennas plotted on Smith chart representing fifty-ohm coaxial cable feed. Monopole $K_{m(1)}$ conductor is #10 gauge wire; Monopole $Km_{m(2)}$ conductor is 4.0 inch diameter tubing. End input impedances are those for each monopole radiating as well as operating in a $R\Omega = 10$-ohm lossy r.f. environment.

of widely different radii, a. For one monopole let's use a number 10 gauge copper wire having a radius a_1° of 0.0510 inches or 4.246×10^{-3} feet. For our second conductor choice we will go all out and use aluminum tubing having a radius a_2° equal to 2.0 inches or 1.667×10^{-1} feet.

In stating our problem, we said the λ/4 monopole would be naturally resonant at f_0; by definition, this sets the monopole *electrical* length h° to be equal to 90 degrees, regardless of the conductor radius, a, at 3.750 MHz. By giving the length h° of the antenna in electrical degrees at f_0 instead of in feet, meters, or some other units we must then convert our selected conductor radii into degrees at the operating frequency to avoid having a case of apples and oranges. This is easy to do. At the frequency of 3.750 MHz, the wavelength λ in *free space* is just 984.00/3.750 MHz equals 262.400 feet; this wavelength dimension of 262.400 feet also represents 360 electrical degrees in *free* space at our f_0. The two selected monopole conductors at 3.750 MHz then have radii in degrees respectively of,

$a_1^\circ = (4.246 \times 10^{-3}\text{ ft.}/262.400\text{ ft.}) \times 360^\circ$
$= 5.825 \times 10^{-3}$ degrees.
$a_2^\circ = (1.667 \times 10^{-1}\text{ ft.}/262.400\text{ ft.}) \times 360^\circ$
$= 2.287 \times 10^{-1}$ degrees.

This data immediately "arms us for bear." We now know radii a° and electrical length h° for these two particular monopoles. To handle them as simple *lossless* r.f. transmission line stubs from here on, we convert both monopoles into their analogue lines by finding the value for $K_{m(1)}$ and $K_{m(2)}$ using Schelkunoff's formula (1.0-1.):

Monopole/transmission line #1:
$a_1^\circ = 5.825 \times 10^{-3}$ degrees;
$h^\circ = 90.00$ degrees.

$$K_{m(1)} = 60\left[2.3026 \log_{10} \frac{2\,(90)^\circ}{(5.825 \times 10^{-3})^\circ} - 1\right]$$
$$K_{m(1)} = 60\,[2.3026 \log_{10} 3.090 \times 10^{-4} - 1]$$
$$K_{m(1)} = 60\,[9.339] = 560.31 \text{ ohms}$$

Monopole/transmission line # 2:

$a_2^\circ = 2.287 \times 10^{-1}$ degrees

$h^\circ = 90.00$ degrees.

$$K_{m(2)} = 60\ [2.3026 \log_{10} \frac{2\,(90)^\circ}{(2.287 \times 10^{-1})^\circ} - 1]$$

$K_{m(2)} = 60\ [5.668] = 340.10$ ohms

Fig. 1 (a) and (b) represent the general picture case of this monopole antenna-into-r.f. transmission line stub conversion we have just made. Notice that the *monopole* antenna over ground converts into a single conductor laying parallel to the ground plane to form an *unbalanced* transmission line. We could have shown a coax line also, except that it's harder to draw. (If we had wanted to make our conversion for a doublet antenna high above ground, our picture would have shown a two wire, *balanced* transmission line floating in space.) One more thing: the shunt capacity to ground in an actual vertical monopole varies with height points on the monopole. As a transmission line's characteristic impedance is related to *distributed* conductor L and C as $Z_o = \sqrt{L/C}$, this means that the actual characteristic impedance K_m of a cylindrical conductor monopole also changes to different values as height above ground changes. Equation (1.0-1.) is based on taking the *average* characteristic impedance over the range of its variation from the base to the top of the monopole, but taking into account all the various wave modes on an antenna while doing that. We are glad that in doing all this, Schelkunoff made the answer come out so simple for us, but that simplicity covers a lot of electromagnetic "muscle". His work converted a mess of difficult "jungle trails" ventured on only by highly trained experts into a nice clean "freeway" which any of us can travel over to solve our problems quite easily. Now, back to fig. 1 (a) and (b).

On the left hand end of the line stub section, the "hot" or base end of the monopole or its analogue line has an input terminal labeled 1; the ground input terminal is labeled 2. Then at the far opposite end of the stub line conductor, a terminal is located there labeled 7; the "ground" for the top end of the monopole or analogue line end point is shown marked 8. Finally, the length h° of the analogue line is the same as the electrical height h° of the monopole antenna it represents, and the label K_m is shown representing the characteristic impedance of the line, but now viewed as if it were a *constant* value which does not change with analogue line length.

To be able to calculate the v.s.w.r. in a fifty-ohm feed line connected either to the monopole or to the analogue line, we have to find the total input impedance,

$$Z_{in(1\ 2)} = R_r + R\Omega + jX \quad , \quad \text{OHMS,} \qquad (1.0\text{-}2.)$$

where the R_r resistive part is the radiation resistance of the monopole. The $R\Omega$ resistive part is the ohmic, non-radiating loss of not only the antenna conductors but also the soil, radial wires, insulator leakage resistance, trap ohmic loss etc. of the antenna's *total electromagnetic environment*. Finally, the jX part is the antenna input reactance; although shown as a plus inductive reactance, it can also take on a minus sign to indicate capacitive reactance. We must now pull a few tricks to make our next steps easier to handle, then later on correct for this sneaky simplification. Right at this point we also must depart from Dr. Schelkunoff's more elegant mode theory to avoid use of higher mathematics in what follows, but still retain the power of his equation (1.0-1.). Therefore, the writer must assume sole responsibility for the method used here which, however, still follows the same wide "freeway" built by Dr. Schelkunoff.

First, we are going to assume that our antennas *do not radiate at all!* This gets rid of the "good part" of input impedance, the radiation resistance R_r. Then, we are going to assume that the antenna as well as its total environmental QTH has no ohmic loss; that discards the nasty and unloved $R\Omega$ part of $Z_{in(1\ 2)}$. All we have left is a *lossless* r.f. transmission line stub representing our monopole antenna. It certainly should be easy to handle now, using simple transmission line theory; but we have one tremendous thing going for us now: *We now know the characteristic impedance K_m of the analogue lines representing our two particular monopoles.* Glancing gack to Fig. 1 (a) and (b), we see that both the top end terminal 7 of our monopole and the end of the analogue line are "open circuited" into free space. When a lossless transmission line stub is open circuited at its "output" terminals like that, we can find its input reactance from the easy formula,[4]

$$jX_{in(1\ 2)} = -jK_m \cot h^\circ \qquad \text{OHMS} \qquad (1.0\text{-}3.)$$

Because we already know all the values to plug into equation (1.0-3.) at the band center frequency 3.750 MHz, we will start there. We defined h° as being exactly 90 electrical degrees long at that frequency. Therefore,

$$jX_{in(1\ 2)}\,f_o = -jK_m \cot 90^\circ = -jK_m\,(0.000)$$
$$= -j0 \text{ ohms}$$

Naturally we expected all along to get that kind of answer at the monopole's resonant frequency. With that 0.000 multiplier, there is no need for us to insert the actual values of $K_{m(1)}$ or $K_{m(2)}$ at f_o. But we suspect that things won't continue like this when we move over to f_{low} or f_{high} in the band. Take the low band limit of 3.500 MHz; Here, the frequency proportionality is 3.500 MHz/3.750 MHz equals 0.933. Our total electrical line length h° now shrinks to $90.000^\circ \times 0.933$, or 84.00 degrees. Equation (1.0-3.) now tells us,

(Continued on page 73)

[4] If a reader is a trifle rusty in trig or algebra, the author recommends "Basic Mathematics for Electronics," by M. Cooke and H. Adams, McGraw-Hill Book Company, Inc. New York.

Results

The array has been used for the past six months in monitoring SWBC transmissions from the Middle East. My receiver and tape recorder are operated by a timer, and I later remove the tapes and listen to them while commuting. This array has been far superior to the ones I had previously used—including a sloper and a bobtail bidirectional broadside curtain—in providing a strong, steady signal for tape recording.

On 40 meters, good results have been obtained in working European stations. I seem to be one of the earliest stations west of the Eastern seaboard to hear and work Europe. North American QRM, other than from VE1-VE2 and W1-W2 areas, appears to be attenuated to a worthwhile degree.

The design parameters chosen for this array were conservative. By increasing the and/or spacing, considerably higher gain can be achieved. Although the array is not small, it has not turned out to be troublesome to raise and maintain, except for getting the main support lines up and clear. Similar arrays for 80 or even 160 meters would not be unreasonable. For MARS use, or if additional hambands become available, the LPQA's broadband characteristics offer many advantages.

I would appreciate hearing from others who construct LPQA's. ■

The Powerlarm *(from page 47)*

larm is most useful. An intruder may pull the main switch to your house so that he can disconnect power being supplied to lights or various security guards which do not have a no-break power supply. A glance out the bedroom window to see if any other lights are visible will help establish whether the power outage is local or covers a neighborhood area. At any rate, you are forewarned.

The Powerlarm also serves a useful function when trying to locate a circuit breaker or fuse which controls a specific outlet you wish to disconnect in order to work on it. Plug the Powerlarm into the outlet and switch off the circuit breakers one by one until you hear the buzzer sound off. And if you plug the Powerlarm into the same receptacle as your deep freeze, you'll be able to tell when you've lost power to that unit. You may not notice for days that the external freezer panel light has gone out and that could mean the possible loss of all your frozen food and meats as the temperature rises above the safe deep freeze point.

The Powerlarm can be left plugged in permanently and dissipates about 3 watts power, the same as an electric clock. The Powerlarm can be easily assembled in an evening for a total cost of under $6.00. It only has to work once to pay for itself many times over. ■

On A Clear Day *(from page 46)*

that 'sneaky' — in your actions, I should say that you have been where things of great interest and dramatic impact have been occurring. Would you care to tell me what they were?"

"I guess you might say that I have been at the site of our newest 'repeater.' Now speaking of repeaters . . . we have a pretty active one in the basement. You might tell Thumbs what you told me about having to be up high to really produce, though. She doesn't know that."

"You mean she's . . . ?"

"Yes, she's! I have just come from the OB Ward. Several little feline offsprings have begun to appear there. There are certain traditional results from over-socializing on fox hunts, you know. You might say it was 'CFAR' all the way, too . . . Cute Felines Arriving Rapidly, Get it? When I left, things were what you in that 'Q Code' would call about 'ten four.' But things were happening pretty fast. She had an awful lot of help on that Fox Hunt. From what I saw, I think you'll be able to get on that other repeater and tell them that it looks like W9LC will soon be up to Ten *Twenty!*" ■

S.S.B. Theory *(from page 43)*

to the transmitter's abilities to reduce the carrier and unwanted sideband. Also, the terms "upper sideband transmission" and "lower sideband transmission" refer to which sideband the transmitter does not reject.

It is, however, not my purpose in writing this article to go into great detail about the actual makeup of single sideband systems. I do hope that, in presenting this alternate explanation, I may have shown that single sideband theory isn't completely senseless. ■

Multi-Band Traps *(from page 30)*

$jX_{in\ (1\ 2)}\ f_{low} = -jK_m \text{ cotan } 84^\circ = -jK_m\ (+0.105)$ ohms

Plugging in our respective K_m's for the two monopoles, we get,

$jX_{in\ (1\ 2)}\ f_{low} = -j560.32(+0.105) = -j58.834$ ohms
$jX_{in\ (1\ 2)}\ f_{low} = -j340.10(+0.105) = -j35.710$ ohms

Recalling that a linear antenna seems to act, at least in terms of its impedance behavior with frequency within a *single ham band,* like a series LC circuit, we see that we indeed obtain a capacitive reactance on the low frequency side of resonance like that predicted for such "circuit." At the high frequency band limit of 4.000 MHz, the frequency proportionality is now 4.000 MHz/3.750 MHz, so our analogue line length h° becomes $1.067 \times 90^\circ$, or 96.000 electrical degrees. Now equation (1.0-3.) tells us,

$jX_{in\ (1\ 2)}\ f_{high} = -jK_m \text{ cotan } 96.000^\circ$
$= -jK_m\ (-0.105)$ ohms

On the high frequency band limit the sign of cotangent 96.000 degrees flipped sign on us in the trig tables, so that minus times minus operation gets us,

$jK_{m\,(1\,2)}\,f_{high} = -j560.32(-0.105) = +j58.834$ ohms

$jK_{m\,(1\,2)}\,f_{high} = -j340.10(-0.105) = +j35.710$ ohms

Again, these inductive input reactances at the high frequency band edge look broadly ok as predicted from the series LC circuit idea. But notice a funny thing: the fat conductor monopole of $K_{m\,(2)}$ equal to 340.10 ohms gives us less input reactance at either band edge than that obtained for the skinny wire conductor monopole. Let's see what this means in terms of v.s.w.r. in the fifty ohm coax feeding our two monopoles on Eighty meters.

Oh oh! One reader just shouted, "Wait a darn minute there, OM! Those pure reactive impedance answers will give a v.s.w.r. ratio of infinity-to-one in that feed coax!" That reader is so right, but it was said we were being sneaky here. When we started out we just threw away the radiation resistance R_r of the monopole, and the ohmic QTH loss $R\Omega$. How could we get away with such a high-handed trick? Well, over the total frequency width f_{high}, f_{low} of any assigned h.f. amateur band, the radiation resistance R_r and ohmic loss $R\Omega$ change so little in value that we can regard them from a *practical* viewpoint as constants. It is only the jX_{in} reactive part of the antenna's complex input impedance which flies all over the place, madly changing value if we change r.f. frequency, change antenna conductor diameter, change guy wires, etc. You name it! It is the reactive part which we have to be mighty careful about in our antenna design in order to make our sky wires put out optimum strength signals on the air.

(To Be Continued)

CQ Reviews *(from page 25)*

but no doubt their life would be extended by keeping the input to no higher than 160 watts on c.w.

Keying characteristics are very much to my liking —just hard enough to make for good copying without any key clicks being evident. A.l.c. action on s.s.b. is effective, requiring considerable carelessness to splatter one's neighbors.

Harmonic radiation is rated by Kenwood at better than 40 db down from the output signal. Measured harmonic levels varied from 47 db down to 52 db down, depending on band. Carrier suppression was 48 db down and sideband suppression was 56 db down. Pretty good figures in anyone's book.

The r.f. speech processor operates at a frequency of 455 kHz. It is definitely effective, and can make the difference between solid and partial copy under crowded band conditions. Kenwood claims a low distortion level for this system, which is true. They also claim that it will not deteriorate the tonal quality of the voice. This claim is also true up to about 10 db of compression level.

Beyond that level, tonal quality changes drastically. Friends with whom you communicate regularly may find the change so drastic that they will ask you to turn off the processor. I have not found any r.f. processor unit that I did not find objectional as regards tonal quality at the higher compression levels. I say this so that you won't think I'm picking on Kenwood. If you are using some r.f. processing now, it may explain why your friends think someone is bootlegging your call.

Is the TS-820 worth the money? Yes, indeed, comparing features versus price, it's a good value in today's market. Would it be worth my trading in a TS-520 to get it? Yes and no, depending on your style of operating. I personally have a love affair with my TS-520. While some of its shortcomings have been eliminated in the design of the TS-820 and the latter has some luxury features that are not available with my TS-520, the TS-520 is more than adequate for my operating habits. A little s.s.b. and a little c.w. in moments of relaxation is my style. If I were an ardent DXer or contest operator the extra edge afforded by the TS-820 design would be worth every dime. If you choose the TS-820, rest assured you will not come up a loser. ■

Trindade Island *(from page 19)*

noise; even our frequent schedules with PY2FIQ in Sãn Paulo were unsuccessful. Our maximum hourly contact rate on 20 meters reached 200 QSOs per hour on voice and 150 contacts per hour on code.

Band openings to certain regions, notably the Western United States and Oceania, were severely limited. Nevertheless, a moderate number of contacts were made with stations in these areas. Unfortunately, because of our unexpectedly early arrival and departure, many DXers "missed the boat." Our abrupt disappearance caused considerable concern and speculation about our fate amongst DXers. The rumors, we are told, ran the gamut from rig failure to a shipwreck and similar catastrophies.

In addition to the aforementioned accomplishments of the DXpedition, I should point out a few of our more dubious distinctions. We spent more than six days on a ship for less than one day of operation. We both acquired mild sunburns *and* mild colds. I learned three words of Portuguese: *obrigado* (thank you) and *cinco-nove* (five-nine). And finally, we have gained enough memories to last a lifetime.

In concluding, we would like to say *obrigado* to the following people for their valuable assistance: PY1CGM, PY2QY, PY1AFA, PY1DMQ, Lucia Alves, the Brazilian Navy, the Commander of Trindade, the authorities of Dentel, and Rolf's XYL, Kitty. We are both eager to return to Trindade for an encore. ■

Even if you buy your trap antennas, understanding how they work can help you make better decisions.

The Multi-Band Trap Antenna—Part II

BY JOSEPH M. BOYER*, W6UYH

Part I, in which the author set the stage for an understanding of multi-band trap antennas by an analogy to transmission lines, appeared in February CQ. In part II he discusses the reaction of the antenna upon itself, and upon lumped reactance.

Now we will just reverse our earlier action and restore those discarded resistive parts R_r and $R\Omega$ in our input impedance, putting them *in series* with our newly found jX reactance values. This is easy with regard to the R_r part. Every *naturally resonant* $\lambda/4$ monopole ever constructed since the first growling spark c.w. signal flashed into the ether possessed a radiation resistance R_r very close to 36 ohms in value. The $R\Omega$ ohmic part is a bit different in nature, so we will defer consideration of it for a bit longer.

Fig. 2 shows a Smith impedance chart which represents a coaxial transmission line whose characteristic impedance Z_o equals fifty ohms feeding our two monopoles. (See now why we used K_m instead of Z_o to represent our analogue line?). The vertical line from top to bottom is calibrated in values of pure resistive ohms. At the chart center on the R line we have an impedance value of $R + jX = 50 + j0$ ohms, or a perfect impedance match point yielding a v.s.w.r. of 1.0:1 in fifty ohm coax.

Now the chart is printed on the page so that the chart notation is upside down. That's perfectly ok, because the point $R + jX = 0 + j0$ ohms is then located at the very bottom on the inside rim edge of the chart. That impedance point will temporarily represent the input terminals 1, 2 of our monopole at f_o when it is *not radiating*. Adjacent to this $0 + j0$ point (representing a short circuit) we see a zero marked on the very outermost electrical distance scale which is labeled Wavelengths Toward Generator (W.T.G.). It climbs circularly clockwise around the chart and again, adjacent to each of its calibrated $\lambda = h^\circ/360^\circ$ points of distance (along the line or antenna) we see points of pure inductive reactance $+$ jX located right on the inside rim scale of the chart when it is upside down. The pure reactive inside rim scale ends at the top of the chart at the point $+$ jX equals infinity. Adjacent to this infinite inductive reactance point we see that the W. T. G. scale is marked 0.250 wavelengths. That 0.250 λ point represents both terminal 7 at the very top end of the monopole and the "output terminal" of the analogue line.

From that top point on the Smith chart, if you inch a hair more distance beyond 0.250 λ on that W.T.G. scale, the sign of the pure reactance marked on the inside rim edge flips to a minus and you have entered the region of capacitive reactance, as the upside down printing on the chart tells us over on the right hand side. Again, you observe that $-$ jX changes from $-$ j infinity at the top down to $-$ j0 ohms at the bottom point, where the W.T.G. distance scale is marked 0.500 λ.

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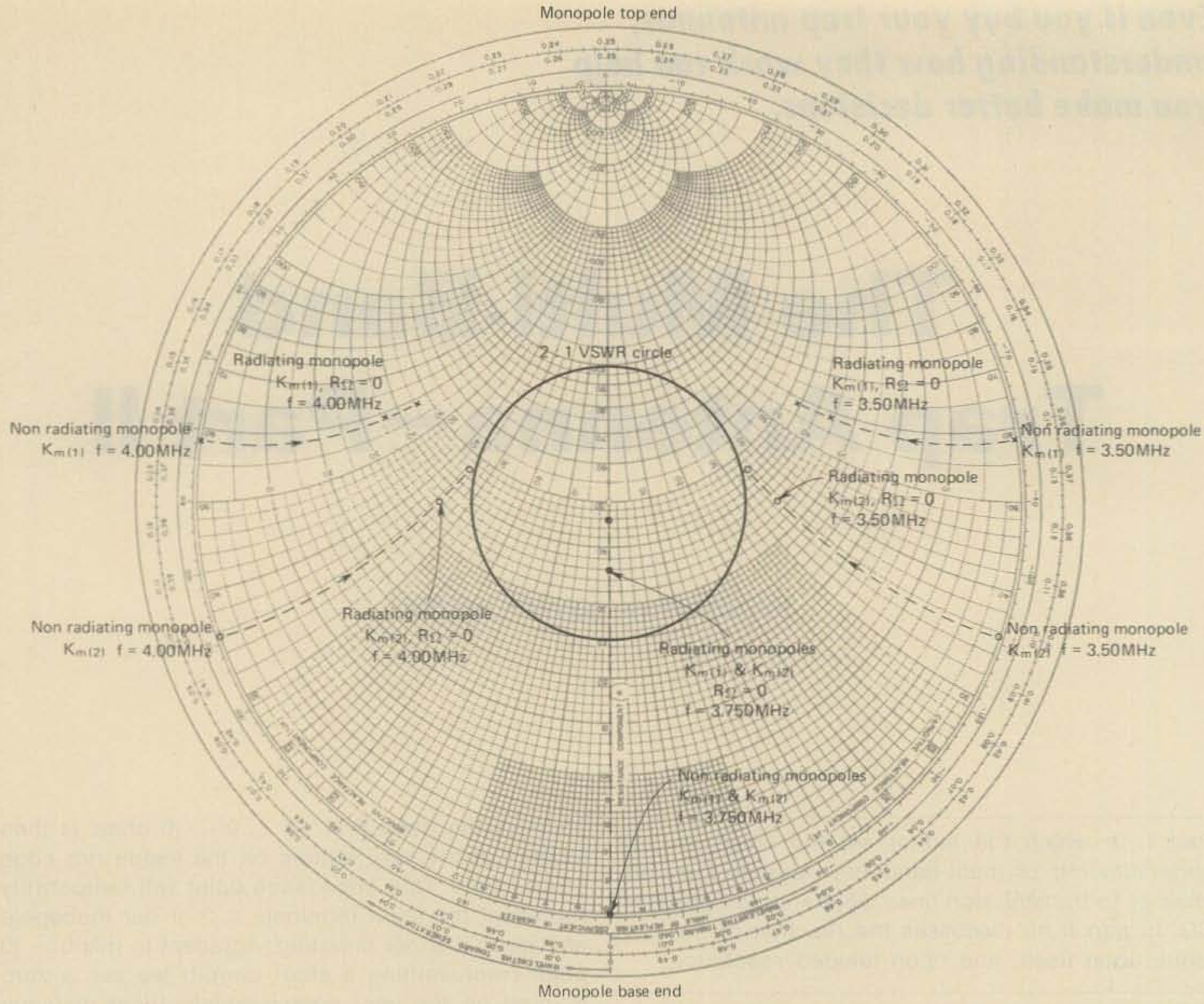

Fig. 2—Calculated input impedances for each of two naturally resonant λ/4 monopole antennas plotted on Smith chart representing fifty-ohm coaxial cable feed. Monopole $K_{m(1)}$ conductor is #10 gauge wire; Monopole $Km_{m(2)}$ conductor is 4.0 inch diameter tubing. End input impedances are those for each monopole radiating as well as operating in a $R\Omega = 10$-ohm lossy r.f. environment.

Now, for our *non-radiating* monopole case, those reactances we just calculated are shown marked as points on the inside rim edge scale. With the chart upside down, the points obtained for the high frequency band limit are located on the left as inductive reactance; those for the low frequency limit on the right as capacitive reactance. The reactances for the skinny wire monopole are indicated with small case x's; those for the fat monopole as small circles or dots. Band frequency and K_m indicators are marked next to these reactance points. If we now place our known value of R_r *in series* with each of these reactive points obtained, they travel upward along those dashed constant reactive line paths, and all stop movement when they encounter the circular line representing 36 ohms of resistance. Only the points of impedance for the band center frequency f_o lay over each other, moving up with the addition of R_r to end at a $Z_{(a, 1, 2)} = 36 + j0$ ohms on the pure resistive center line. The impedance points for 3.500 and 4.000 MHz end up in a no-man's land of complex impedance (both R and ± jX) outside that "magic" 2:1 v.s.w.r. circle marked on the chart. That 2:1 v.s.w.r. circle is "magic" to the modern day amateur for the following reason: unlike old time ham rigs which could use almost any random hunk of chicken wire as an antenna, modern commercially-built ham rigs will not load full output power into a feed line having much more than a 2:1 v.s.w.r. Of course we love those beautiful, shiny, computerized, digital-dial read-out, miniaturized wonders with their bells and whistles. But that is why we must revere that 2:1 circle on the chart when it comes to our antennas these days. Now, back to the chart.

We wanted v.s.w.r in our fifty-ohm coax line. Ok, the v.s.w.r. for the 3.750 MHz band center impedance is easy to obtain because there is *no reactance* at this frequency. For both our monopoles, VSWR is just,

$$VSWR = \frac{50\ \Omega}{36\ \Omega} = 1.39{:}\ 1\ (f_o)$$

Getting v.s.w.r. in the feed line at 3.500 MHz and 4.000 MHz gives us a bit of a problem due to the complex impedance at those frequencies. We could get fancy here and go to the bother of calculating what is called the complex reflection coefficient, and then get v.s.w.r from it. Most working antenna engineers don't bother; they use a lazy man's way to get v.s.w.r. when the impedance on the chart is complex: put the pin end of a drawing compass into the chart center point of 50 + j0 ohms, then pull out the pencil end and put its tip on say the 3.500 MHz impedance point for the radiating (but ohmic loss free) monopole of $K_{m\ (1)}$. At this radius a circle is lightly drawn on the chart. The point where such pencil circle cuts the pure resistive axis *below* the center point is "eye balled" in value. Doing this gives a pure resistive intercept of about 13.5 ohms for the $K_{m\ (1)}$ skinny monopole at 3.500 MHz, so

$$VSWR_{K_{m\ (1)}} = \frac{50\ \Omega}{13.5\ \Omega} = 3.7{:}1 \quad (f_{low})$$

Using the same lazy man's technique for the radiating but ohmic loss free fat monopole at 3.500 MHz gives us a resistive axis intercept of about 20.5 ohms, or a v.s.w.r. of about 2.44:1. If we had used the resistive axis intercept above the chart center (R larger than 50) and carried out the indicated division we would have v.s.w.r. going from 1.0:1 down to values *less than* 1.0. That is the way our British cousins and some Europeans figure v.s.w.r. It means the same thing, but can confuse us Yankees sometimes at first glance.

Again we see that the fatter $K_{m\ (2)}$ radiating monopole gives us a lower v.s.w.r value at the low frequency band limit than its skinnier counterpart. This illustrates a very important fact from antenna theory which says that larger diameter antenna conductors change reactance at a *slower rate* ($\Delta X/\Delta f$) with frequency and therefore give a *broader* impedance frequency *bandwidth* for a given limiting value of v.s.w.r. than skinny antenna conductors. What mysterious factor in our simple calculations here causes that effect to be seen? The neat value of antenna K_m Dr. Schelkunoff worked out for us to use.

Now we must consider the other ohmic loss term $R\Omega$. Everytime an antenna man eyes that term he winces inwardly; half of his time is spent fighting to keep it small; still it always exists in the real world of antennas. The ohmic environmental loss varies from QTH to QTH even for identical antennas. Discussing it could fill a text book. Here we can only note that in a "typical" US soil environment, using about six λ/4 radials and high conductivity antenna conductors, a "ball park" value for $R\Omega$ at the *lower HF ham frequency bands* would be about 10 ohms. If we now put this additional 10 ohms in series with R_r in our calculated antenna input impedance to get $Z_{in(1\ 2)} = 36 + 10 + jX$ ohms, this added "real" part of 10 ohms pushes all our calculated impedance points a bit farther along those little dashed constant reactance curves to stop on the R equals 46 ohm circular line, and a bit closer to the chart center. We see that adding non-radiating ohmic loss also lowers feed line v.s.w.r. in this case (but that is a poor way to lower antenna v.s.w.r.!). Now that our fatter, naturally resonant monopole is radiating and operating in an environment possessing a realistic ohmic loss, its $Z_{in(1\ 2)}$ impedance has almost reached our magic 2:1 circle at the 3.500 MHz band edge. Unfortunately, this is not true for the wire monopole.

Someone says, "Hey, my pencil drawn circle through the impedance at the low band limit also cuts right through the impedance point for the high band limit. Do real world antennas have pretty, symmetrical impedance curves like that, spaced equally on either side of f_o ?" Uh . . . no! Not quite! In our calculations we didn't include a *lumped* capacity which *always exists* between the base end of a monopole and, via the dielectric constant of the base insulator, to the ground plane. This lumped capacity is connected in *parallel* across the *series* input impedance of the monopole. What it does is to "skew' the shape of the Z_{in} curve with frequency about f_o so that the magnitude of Z_{in} at f_{low} is not equal to that at f_{high}, with just a change of reactance sign.[5] The use of a hollow base insulator as well as tapers in the conductor at the feed point can reduce this effect so that in well designed antennas it is minimal *at h.f.* The same thing happens in a center fed doublet from the *lumped* capacity across the feed insulator.

The Antenna Reaction Upon Itself And On Lumped Reactance

If we give a small child some adult gadget like a wind-up alarm clock to play with, the child invariably starts trying to take it apart. This inborn human instinct is intelligent and commendable—it eventually results in a better understanding of how things work.

When we try to take a λ/4 monopole apart we just end up with a number of shorter lengths of antenna; in terms of the analogue of the antenna, we then have a collection of electrically shorter transmission line sections $h_1 + h_2 + h_3 + \ldots h_n$ which *all add up* to a total h° of 90 degrees. The natural question which arises in our minds is this: how do each of these individual short sections of the total antenna end up interacting reactively on one another to produce an input reactance $jX_{in(1\ 2)} = j0$ ohms as resonance?

It turns out that the answer to this seemingly academic question is an important step along the road to easy design not only of the Morgan trap antenna, but many other kinds as well. Let's take a look at this problem. To do so won't require a lot of analogue

[5] Also, the radiation resistance R_r is *slightly* larger on the f_{high} side of f_o than on the f_{low} side, and this adds to non-symmetry in Z_{in}.

line sections; just two will do. Such line dissection is shown in fig. 1 (c). All we have done is to cut the single analogue transmission line section of Fig. 1(b) into two shorter length sections h_1 and h_2. A new terminal, 3, now appears on the conductor as the "output" terminal of the left hand line section h_1, and directly below it a corresponding ground terminal, 4. A new terminal, 5, on the conductor now represents an "input" terminal for the right hand line section h_2, and directly below it a companion ground terminal, 6. All that the dual listing of K_m under both of the line sections means is that the line characteristic impedance is of the same value in both. The total electrical length $h^\circ = h_1 + h_2$ of the analogue line is still 90.000 degrees, although h_1 and h_2 may be proportioned *any way we like* as long as they sum up to this specified total length. Oh yes, there is a little dashed line conductively connecting terminals 3 and 5 together. It represents a wire "pig tail" lead of "zero" electrical length. Right now let's just reach out and break that pig tail lead so that terminals 3 and 5 are temporarily insulated from one another.

Now we know from our previous exercise with the $\lambda/4$ monopole, that when we look into the input terminals of an RF transmission line stub section when its output terminals are open circuited, equation (1.0-3.) will give us the reactanve $jX_{in\,( )}$ present at the *input terminals* of such stub line. The "output" terminals 7 and 8 of line section h_2 are obviously open circuited, so if we look into the input terminals of this right hand line section we should see,

$$jX_{in\,(5\text{-}6)} = -jK_m \cot h_2 \quad \text{Ohms}$$

Because we set a limit on the total line length h° *at frequency* f_o, the right hand side line section h_2 just has to be less than 90 degrees at f_o at least, unless the line section h_1 ceased to exist by making its length equal to zero degrees. Therefore, if the length of section h_1 is more than zero degrees long, the input reactance $jX_{in\,(5\text{-}6)}$ of the right hand line section will *always* come out as a capacitive reactance $-jX_{in\,(5\text{-}6)}$; in other words, line section h_2 (representing the upper section of the monopole) will *always* look like a *condenser* connected across the "output" terminals of the left hand line section h_1 (when we put our pig tail wire back in place). That sure seems strange, because we *also know* that if we move over to the left and look into the line terminals 1 and ground 2, $jX_{in\,(1\text{-}2)}$ *always* comes out to be j0 ohms *there* when the total h° equals 90 electrical degrees. Somehow that first line section h_1 does *something* to cancel out the capacitive reactance $-jK_m \cot h_2$ at its "output" terminals *no matter how the lengths of the two line sections are proportioned.* To *amplify* our mental attack on this idea, let's employ a "thinking equation" at this point which could look like this:

$$j(K_m \text{ SOMETHING } h_1) + (-jK_m \cot h_2)$$
$$= j0.00 \text{ ohms}$$

Peering at the above relation, we see that if $j(K_m$ SOMETHING $h_1)$ had the same absolute magnitude as $jK_m \cot h_2$, but an opposite reactive sign, then the two terms *would* add up algebraically to j0 ohms. We know that a coil and condenser connected in series will do that when their $+jX_L$ and $-jX_c$ have the same electrical size in ohms at f_o. Therefore, in such approach the $j(K_m$ SOMETHING $h_1)$ term could be made to look like a series inductive reactance to its "condenser" load. This idea is sketched in fig. 1 (d). Now one way to make a transmission line section *look like* a series inductance is to change the first term in our "thinking equation" so it looks like this:

$$jK_m \tan h_1 + (-jK_m \cot h_2) = jO \text{ ohms,} \quad (1.0\text{-}4.)$$

or,

$$jK_m \tan h_1 = jK_m \cot h_2,$$

and even,

$$\frac{jK_m \tan h_1}{jK_m} = \cot h_2 \quad (1.0\text{-}5.)$$

Notice that in equation (1.0-4.), when h_1 and h_2 add up to either *less or more* than 90 electrical degrees, the answer becomes finite in magnitude and flips in reactive sign to go capacitive or inductive in reactance. But that is just what our equation (1.0-3.) did with the $\lambda/4$ monopole on either side of f_o; therefore, equation (1.0-4.) converts into equation (1.0-3.) when h_1 goes to zero degrees in length. Now we can see how a naturally resonant $\lambda/4$ monopole reacts on itself, length by length, to "resonate itself" at f_o; or become plus or minus reactive on *either side* of f_o. However, our equation (1.0-5.) seems kind of dumb. It just says, "Tell me how long h_1 is, and I'll tell you how long h_2 is." But we already know the answer to that question by means of first grade arithmetic; say h_1 equals 47.000 degrees. Obviously then, $h_2 = 90^\circ - 47.000^\circ = 43.000$ degrees. What do we need this more complicated equation for? It merely echo's: $\tan 47.000^\circ = \cot 43.000^\circ$

Well, now let's become sneaky again. What if we opened up that gap between terminals 3 and 5 and inserted there a *coil* whose inductive reactance at 3.750 MHz was equal to + j150 ohms? Because we've all played with *electrically short,* coil loaded mobile whip antennas, we see that this is what we would create here: a coil loaded monopole, with the "loading coil" located at an electrical height h_1 of 47.000 electrical degrees from the base input terminals. We know something else, too: that *top* line section h_2 can not remain at 43.000 electrical degrees if we wish to obtain a resonant j0 ohms input reactance at 3.750 MHz. Our new *coil loaded* monopole and its equivalent analogue line is shown in fig. 1 (e) and (f). Now we know we must *shorten* the electrical length h_2 of our monopole top conductor section, *but by just how much?* Let's try using our new found equation (1.0-4.) to get an answer, by sticking that *known* coil reactance in series with the other terms.

$jK_m \tan h_1 + j150.00 + (-j \cot h_2)$
$= j0.00$ ohms (f_o)

Also, let's reach back and use our $K_{m(1)}$ of 560.32 ohms representing our skinny number 10 gage wire monopole. Plugging in that K_m value, plus our known length h_1 of 47.000°, we get:

$j560.32 \tan 47.000° + j150.000 + (-j560.32 \cot h_2) = j0.00$ ohms (f_o)

$j560.32\ (1.0724) + j150 + (-j560.32 \cot h_2) = j0.00$ ohms (f_o)

or,

$$\frac{j560.32\ (1.0724) + j150.000}{j560.32} = \cot h_2 =$$

$$\frac{740.887}{560.32} = 1.340\ (f_o)$$

$\cot^{-1} 1.340 = 36.731°$ (at f_o)

Is that correct? Let's stick our found length for top section h_2 back in and see:

$j560.32\ (1.0724) + j150.00 + (-j560.32 \cot 36.731°) = j0.00\ (f_o)$

$j750.887 + (-j560.32 \times 1.3401) = j0.00\ (f_o)$

$j750.887 + (-j750.887) = j0.00\ (f_o)$ ohms.

Now we can go into the coil loaded mobile whip business if we wish. We just specify how high in electrical degrees h_1 we will place our loading coil of so many $+jX$ ohms, from the base input terminals of the whip; calculate the whip conductor K_m from (1.0-1.) on the band of interest, and then our modified equation (1.0-4.) and its conversion (1.0-5.) tells us how long electrically our top conductor section *must be* to get resonance. A reader says, "Sure, and you can turn it around and use it to solve for the needed loading coil reactance $+ jX_L$ when you know *coil height* h_1, and top section length h_2." He pauses a minute, and then adds, "And it even works for coil *base loading* when you let h_1 go to zero degrees." That reader is so right! Another OM out there pops up with, "Aren't we working a bit too hard? I mean , if K_m is uniform in value in both line section h_1 and h_2, why don't we just divide through by K_m and get rid of it to make our equation even more simple." Ok, let's do that:

$$\frac{jK_m \tan h_1 + jX_L + (-jK_m \cot h_2) = j0.00 \text{ ohms } (f_o)}{jK_m \qquad jK_m \qquad jK_m}$$

$$\tan h_1 + \frac{X_L}{K_m} + (-\cot h_2) = j0.00 \text{ ohms } (f_o) \quad (1.0\text{-}6.)$$

The above equation is still perfectly valid, except now we would have to *multiply the final answer* (if other than j0.00 ohms at resonant frequency f_o) by K_m ohms to make it come out in the *actual value* of ohms. As it stands, the above equation is in *normalized* form: normalized with respect to the antenna or analogue line's K_m. Normalization is just a fancy word for the process engineers use in making it less work to calculate impedances in circuits, antennas, and RF transmission lines. It is another "lazy man's" trick, except that this one in no way reduces accuracy as our other compass and eyeballing resistive values to get v.s.w.r. did.

Well, we have warmed up our trig and algebra, and old Mister Sprinkle back in H.S. 59 would be proud of us for that, but what the heck has the subject of shortened, coil loaded antennas got to do with the Morgan multi-band trap antenna?

Sorry OM's, but I've been kinda sneaky again here! We have just *completed* the electrical design of a *two band* Morgan trap antenna made of number 10 gage copper wire! It is operating on the eighty meter ham band, resonant on 3.750 MHz. On eighty meters, its forty meter "quarter wave" bottom conductor section h_1 ended up being only 47.00 electrical degrees in length. That 150 ohm series "loading coil" turns out to be what the forty meter *parallel* LC trap looks like at the frequency 3.750 MHz.

What we were *actually doing* was solving to obtain the *needed* electrical length of the *next* conductor section h_2 located *above* the non-resonant forty meter band trap in order to make the *entire eighty meter monopole section* of the two band Morgan end up resonant at 3.750 MHz. Notice that $h_1 + h_2$ representing the total electrical lengths of the Morgan *conductors* no longer add up to 90 electrical degrees as they did before in the *naturally* resonant $\lambda/4$ monopole at f_o. Instead, in our skinny wire monopole, that off-resonant forty meter band trap *added* 6.27 electrical degrees to our Morgan on eighty meters; We had to remove that amount of electrical degrees from our *conductor* length *above* that trap, therefore to make our Morgan resonant at 3.750 MHz. In our fatter monopole of $K_{m(2)}$, the *same* 150 ohm magnitude of series trap inductive reactance, located at the precisely *same* electrical height h_1 from the base input terminals of the monopole, would require h_2 to be only 33.455 electrical degrees in *conductor* length; the *same* non resonant trap would have added 9.45 electrical degrees to the fat monopole on eighty meters. You will find out later that this "loading" effect by the traps, when non-resonant, places a limitation on the performance of the Morgan antenna when any band section of it is compared to that of a naturally resonant monopole for that band using identical conductor diameter $d = 2a$. But we will learn how to make this limitation minimal with our new-found sneaky ways.

Now we are armed; our design muscle has grown; we are now getting a feel for this antenna/ transmission line analogue tool. We are now ready to take on a Morgan trap antenna covering all the ham bands from ten to eighty meters and make it play correctly. It is just a matter of repeating the process we used here, band-by-band, stringing all those band traps and conductor sections in series as "the foot bone's connected to the ankle bone, and the ankle bone" on and on. When we reach the "head

(Continued on page 72)

A.M. Is Not Dead *(from page 57)*

stand in line to get to use the repeater and then get told how to use it.

Now, a.m. *is* fun; it is *not* a nostalgia trip for a bunch of old fogies—there are plenty of young fellows on who are smart enough to realize that you don't need a fat wallet to enjoy amateur radio. They have discovered the *tremendous,* almost lost, satisfaction to be had from building their own rigs and actually using them on the air to talk with other amateurs with the same interests. Much air time is spent not just on the technicality of producing "broadcast quality" a.m. but in tracking down sources for the high power transmitter components we can no longer buy over the counter. (Even if we could, the prices would be out of this world.) This is why "flea markets" and salvage yards are so popular with these fellows who build. This is why so much old fashioned horse-trading takes place.

A.M. And The S.W.L.

You know, if you stop to think, *a.m.* is what the s.w.l. listens to, not s.s.b. It is astounding to learn how many present day a.m. stations on 75 and 160 have gotten letters and 'phone calls from short wave listeners. No small number of these s.w.l.'s become amateurs because listening to *a.m.* is their introduction to amateur radio. Could the proliferation of s.s.b. be a significant factor in the failure to attract newcomers to amateur radio? Don't knock a.m., use it properly, in the right part of each band. Remember the dire predictions that RTTY would spread all over the bands when f.s.k. was first authorized on the h.f. bands? Well, it didn't happen. RTTYers stuck close to agreed frequencies and they still do. The same thing is happening on 75. Just about all a.m. stays between 3850 and 3900, and many s.s.b. operators respect this and move away from the few a.m. QSO's in progress. (There *is* enough space for all.)

To say it again, Harry, a.m. is *not* dead. Don't let the FCC and the ARRL sweep it under the rug under the guise of "deregulation." Let's keep this basic kind of radio, the radio of a.m. broadcasting, alive by local radio club activity, by active horse trading in old components, by letting magazine editors know that you want to see articles on a.m. Amateur radio *can* still be fun!

Multi-Band Trap Ant. *(from page 55)*

bone" on our last band of coverage, we're through.

Well yes, there *are* still a few minor matters yet to be covered: how to design band traps in terms of their Q and L/C ratio; how to convert conductor sections first found in electrical length over to conductor length in feet when the conductor radius is taken into account. Also, in the process, we will simplify the steps a bit more so things will be even easier to work out. We will have to do all that, however, in parts III and IV, as space just ran out. Oh, and I promise you this: no more sneaky tricks from here on out. ■

(To Be Continued)

DXosis Okinawa* Style *(from page 18)*

Ocean stations, you might try the Pacific Inter-Island Net that meets on 14305 at 0800 GMT.

Now that you DXosis patients have worked all that DX, how do you go about getting confirmations for your contacts? What is the best method to assure a high percentage of returns on QSLs? I wish I knew the answer to this one! I do share with you a method that has netted me over 90% returns from KA6DE. My procedure is simply this. The same day I work a new country, I prepare a QSL card and place it in a business size air mail envelope with sufficient International Reply Coupons for the station operator to purchase sufficient postage for air mail return of my card. IRCs cost 26¢ each at your friendly Post Office or are sometimes available for 15¢-20¢ each from QSL managers. If a station is located in a country that does not honor IRCs you might send him one large green stamp (a US dollar) and tell him you hope this deflated piece of currency will pay for his postage. I also include a self-addressed air mail envelope on which I have typed my name, address and callsign. I also type on the DX station's return address. You will never know how much DX stations appreciate this little favor. DX stations spend much time in addressing QSLs and every little thing you can do to save them time will increase your percentage of returns. You have read a lot about the great success rate of some US operators who adorn their envelopes with those big, beautiful commemorative stamps. I do not do this because it only draws attention to the envelope and a few postal workers in some countries are known to be stamp collectors. I know! They have many of my commemoratives from earlier years. I also suspect that some of them have paid parts of their food bills with my IRCs. So, keep the envelope business-like and do not draw attention to amateur radio on it.

I hope this article has been of interest to those of you who have just recently contacted DXosis. As I said in the beginning, I certainly do not have all the answers on how to build your DXCC total. However, I did have a great time operating as KA6DE these last 15 months, and it has been fun sharing these ideas with you. My DXosis problems have diminished as of August 1975 when I return to The World for a new assignment at Lowry AFB, Colorado. As my premedical student roommate knew, all skin problems subside only to return again, so will my DXosis. However, I am not really that concerned about my disease. As a matter of fact I really enjoy having DXosis. I hope you do too. ■

A close look at the electrical nature of lumped LC trap circuits.

The Multi-Band Trap Antenna—Part III

BY JOSEPH M. BOYER*, W6UYH

This paper discusses the function and design of the multi-band trap antenna invented by Howard K. Morgan† in 1940. Up to this point, a picture of this periodically resonant antenna as a collection of single band naturally resonant radiators, isolated from one another by parallel resonant LC circuits, magically snapping "open" and "shut" like switches activated by frequency sensitive gremlins, has been avoided.

Instead, in Part I, a method was introduced to permit conversion of any antenna into its equivalent r.f. transmission line. This simple concept was then used in Part II as a tool to inspect the calculated input impedance behavior of two monopole antennas which differed from one another only in conductor diameter. In one case the two monopoles functioned as naturally resonant quarter-wave vertical antennas; in the second case, a "loading coil" was inserted in series with each monopole at an electrical height of 47 degrees above the input terminals, and the influence of such coil's reactance on the monopole was inspected quite closely at a center frequency of 3.750 MHz. It turned out that when we were through, we had "inadvertantly" designed a two band Morgan antenna. The "loading coil" reactance had originated from the Morgan forty-meter-band trap operating out of resonance on eighty meters. In the process we had worked out the basic steps necessary to design, analyze, and understand the function of such multi-band antennas.

In this part of the paper, the electrical nature of the lumped LC trap circuits will be looked at, then placed into the Morgan multi-band antenna and its design carried out in progressive band-by-band steps for coverage in the ten to eighty meter ham bands. At the conclusion of the series, some of the limitations inherent in the lumped LC trap, multi-band antenna will be discussed.

The Trap

In describing the impedance behavior of the monopole antenna over the relatively narrow frequency width of a single ham band, the word *series circuit* was used. Although a useful term, the writer always mentally bites his tongue when employing the word *circuit* to describe antenna function. No antenna is really a circuit: if it were it would not radiate at all. In a gross sense, the essential difference between a *true circuit* composed of *lumped* inductance L, lumped capacity C and ohmic loss resistance RΩ and that of any antenna is electrical *size*. True circuits employed in electronic "black boxes" are exceedingly minute in size when their physical dimensions are compared to the wavelength λ of the r.f. energy flowing through them.

When, a passive device (no tubes or transistors) composed of L, C, and R is no longer sufficiently small in terms of the operating wavelength, it begins to function as an RF *transmission line*. All r.f. transmission lines are antennas, radiate waves, and thereby possess an R_r term in their total impedance. This is the basis for all the newer type, electrically small antennas such as the DDRR[1,2],

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†Reference number 1, part I.

[1] J. M. Boyer, "Hula Hoop", pp. 44-46, *Electronics*, January 11, 1963.

[2] U.S. Patents: J. M. Boyer: 3,151,328; 3,247,515; RE 26, 196.

the King[3] BAR antenna, and others. The only way the radiation resistive term R_r can be cancelled in an r.f. transmission line is to bring *another antenna* —or a set of antennas—close enough to it to affect its electromagnetic field. Although strange, this is the correct way to think about a non-radiating coaxial transmission line: the inner conductor is viewed as one *antenna* surrounded concentrically by an infinite number of other parallel, out-of-phase antennas which cancel out the *time average* radiation resistance of the inner conductor over one complete r.f. cycle. In the true sense, there is *no such thing* as a "shield" for radio waves.

A true lumped LC circuit can only function once to simulate a "closed" or "open" switch. Here the term *once* means "at a single RF frequency". A "closed" circuit switch function can be accomplished by L and C connected in series at its single resonant frequency f_o. An "open" circuit switch function is represented by L and C connected in parallel at the single resonant frequency f_o. At all other frequencies over a total RF bandwidth in which the circuit is sufficiently small in electrical size compared to λ, no circuit switch action will occur. Off resonance, a true circuit displays only a rising or falling reactance magnitude, plus of course its ohmic loss $R\Omega$.

In marked contrast, an antenna or r.f. transmission line may be said to act like a periodically "opening" and "closing switch: that is, at an infinite number of different r.f. frequencies, an antenna or transmission line keeps flipping back and forth between a low and then a high impedance condition. The frequencies at which such "switching" functions occur need not even be in harmonic relationship: if you connect two r.f. transmission lines in series, each having different characteristic impedances Z_o (or K_m), you obtain a *compound* transmission line[4]. Such a line flips back and forth between high and low impedance at portions of the frequency spectrum in a non-harmonic fashion.

You might say to yourself, "He certainly is making a big deal out of this business of lumped LC circuits and transmission lines!" Well, we must do that here because we will soon connect what we will initially *assume* to be an ideal lumped LC circuit trap in series with a linear antenna. This is a mating of two electrical systems of quite different behavior: like marrying a lion to a pussy cat.

Ok; as has been said, an ideal parallel LC circuit "opens" only once at a single resonant frequency f_o to yield a high magnitude of impedance Z_p through its terminals. Such parallel resonant impedance can be expressed by the relation,

$$Z_p = \frac{(R_C - jX_C)\,(R_L + jX_L)}{(R_C + R_L) + j(X_L - X_C)} \text{ Ohms} \quad (2.0\text{-}1.)$$

In the above equation, R_C is the ohmic loss resistance of the circuit capacitor; R_L is the ohmic loss resistance of the circuit coil, and X_C and X_L denote the capacitive and inductive reactances of the capacitor and coil respectively at the resonant frequency f_o. In what follows, we will only consider what may be called "good quality" traps: those in which an air or vacuum insulated capacitor is employed. The ohmic loss R_C of a well designed air or vacuum capacitor is so microscopically small (when protected against moisture) at high frequencies that it may be neglected in equation (2.0–1.). Therefore, in high quality band traps, all ohmic circuit loss resistance may be considered to reside only in the coil used. The coil ohmic loss resistance is expressed as,

$$R_L = \frac{X_L}{Q} \text{ Ohms} \quad (2.0\text{-}2.)$$

Because we can now regard all ohmic loss to be in the circuit coil, we can then avoid dealing with the complex impedance $R+jX$ seen in equation (2.0–1.), by using a more simple relation for this the *resonant* parallel circuit impedance:

$$Z_p = \frac{(X_L)^2}{R_L} \text{ Ohms} \quad (2.0\text{-}3.)$$

Once we know the parallel impedance of our trap at the resonant frequency f_o, we may then use Z_p to get the non-resonant trap impedance when it is operated at a much lower frequency f. When f is lower than f_o, the trap will look like a series inductance. When f is displaced from f_o by a minimum factor of 3/Q, we may also conveniently forget about coil loss resistance R_L and obtain that very important trap non resonant series reactance X_s by the relation,

$$X_s(\,) = \frac{1}{Q\,M\left[\frac{1}{(M)^2} - 1\right]} \times Z_p \text{ Ohms,} \quad (2.0\text{-}4.)$$

where M is equal to the given band operating frequency f to trap resonant frequency f_o ratio f/f_o. We will also be using that factor M in terms of the changing electrical length of our Morgan conductors, so it is worth a second glance. Of course circuit Q is equal to X_L/R_L, where we get R_L from equation (2.0–2) for our particular coil. As it's a "ring around" situation, we first choose Q then get R_L afterward.

Now, a parallel connected LC circuit operates like the "mirror image" of a series connected LC circuit: its input reactance goes inductive below the resonant frequency f_o, and goes capacitive at frequencies above f_o. Therefore, in a Morgan antenna, all the traps closer to the antenna input

[3] R. W. P. King et al, "Transmission Line Missile Antennas, *IRE Transactions on Antennas and Propagation*, Vol. AP-8, pp. 88-90, January 1960.

[4] *Very High Frequency Techniques*, Volume II, pp. 922-925, McGraw-Hill Book Company, Inc., N.Y., First Edition.

terminals 1, 2 than the *one resonant trap* at a given frequency band f_o $(_n)$, will look like a string of series loading coils spaced at certain points along the length of that portion of the Morgan operating on this particular band. In the "active" band in which that one trap is resonant, it also displays this low frequency side inductive, high frequency side capacitive reactance behavior; the difference is that in the active band of the trap (a) the magnitude of reactance is much larger than its off band X_s and (b) you can not neglect R_L in calculating such trap impedance. That means you have to use *equation (2.0–1.) for such case.

From what has just been said, we can see that the correct design and analysis of the Morgan trap antenna operating across many ham bands can become very involved and difficult unless a sharp little tool like the antenna analogue concept is put to work to *snip* the total antenna into small parts which can then be easily handled on a band-to-band step basis.

Band Trap Design

The L to C ratio and Q parameters of the parallel resonant circuits selected for the traps in a Morgan multi-band antenna have a first order effect on the final electrical lengths of its conductors and also on its on-the-air performance. We secured an introduction to the conductor shortening effect of the trap "loading coil" influence in part I. Therefore, we will at this point plunge directly into the design of a Morgan five bander using one set of trap parameters. Later, the effect of alternate trap parameters will be discussed. Armed with such information, the reader may then modify his own Morgan design accordingly. Our initial trap design here is based on the following objectives:

(1) Use of standard capacity values for the trap condensers.

(2) Attainment of approximately the same parallel resonant trap impedance Z_p in each amateur band.

The resonant frequency f_o for each band trap will be selected in the center of each ham band from ten to eighty meters. Our f_o list will therefore be:

$f_o(_{10}) = 28.850$ MHz;
$f_o(_{15}) = 21.225$ MHz;
$f_o(_{20}) = 14.175$ MHz;
$f_o(_{40}) = 7.150$ MHz;
$f_o(_{80}) = 3.750$ MHz.

At the lowest frequency band of coverage, eighty meters, no band trap is required. Consequently, we need only four trap capacity values (in farads):

$C_{10} = 25 \times 10^{-12}$ F; $C_{20} = 50 \times 10^{-12}$ F.
$C_{15} = 35 \times 10^{-12}$ F; $C_{40} = 100 \times 10^{-12}$ F.

*The interested reader will find an excellent, lucid treatment of LC circuits by a down-to-earth master engineer in F. E. Terman, *"Radio Engineer's Handbook,"* pp. 135-171, McGraw-Hill Book Company, Inc., N.Y., N.Y.

At the above listed f_o's the capacitive reactance $-jX_C$ of the selected condensers is found by the relation,

$$-X_C = \frac{1}{2\pi f_o\,(\text{Hz})\,C\,(\text{F})}\ \text{Ohms}$$

We therefore obtain the following capacitive reactance values, using three decimal place accuracy:
$X_C(_{10}) = -220.665$ ohms; $X_C(_{15}) = -214.242$ ohms;
$X_C(_{20}) = -224.557$ ohms; $X_C(_{40}) = -222.594$ ohms

Because we know that at resonance the magnitude of X_C must equal that of the coil inductive reactance X_L, we can just remove the minus sign in front of each of the capacitive reactances obtained and substitute these magnitudes into the relation,

$$L = \frac{+X_L}{2\pi f_o(\text{Hz})}\ \text{Henries},$$

to obtain the necessary calculated values of coil inductance needed in each band trap:
$L_{10} = 1.217\times10^{-6}$H; $L_{15} = 1.606\times10^{-6}$H;
$L_{20} = 2.521\times10^{-6}$H; $L_{40} = 4.945\times10^{-6}$H.

Because of the persistance of the idea that the trap serves as an "antenna insulator" and must therefore have super high Z_P, the designer feels the urge to select a very high value of trap Q which must reside in his coil. For reasons which will be seen later, we will not do that here. Instead we will select a reasonable value of 100 for all trap coils. Using this value of Q with each of the resonant frequency reactances we obtained above, we may employ (2.0-2) to get the values of ohmic resistance R_L predicted for each coil in its band of resonance. This easy step yields,
$R_L(_{10}) = 2.206$ ohms; $R_L(_{15}) = 2.142$ ohms;
$R_L(_{20}) = 2.246$ ohms; $R_L(_{40}) = 2.226$ ohms.

Now that we know both R_L and the resonant coil reactance X_L, there is no problem in using equation (2.0-3) to find the resonant parallel impedance Z_p offered by the traps at their f_o's:
$Z_p(_{10}) = 2.207\times10^4$ ohms; $Z_p(_{15}) = 2.143\times10^4$ ohms;
$Z_p(_{20}) = 2.246\times10^4$ ohms; $Z_p(_{40}) = 2.226\times10^4$ ohms.

To utilize that "coil loading" effect of the traps on the length of the conductor sections when they are operating at band frequencies below f_o, we must also calculate the *series* X_s inductive reactance value produced by *each* trap on all the f_o's lying in the bands below its resonance. We use equation (2.0-4.) for this little chore. Here, remember that M is the ratio of operating frequency f to trap f_o:

Ten Meter Trap (f_o=*28.850* MHz; $Z_p = 2.207\times10^4$ ohms)

On $f_o(_{15})$:
$M = 21.225/28.850 = 0.736$
$X_s(_{10})_{15} = 353.801$ ohms

On $f_o(_{20})$:
$M = 14.175/28.850 = 0.491$
$X_s(_{10})_{20} = 142.946$ ohms

On $f_o(_{40})$:
$M = 7.150/28.850 = 0.248$
$X_s(_{10})_{40} = 58.276$ ohms

On $f_o\ (_{80})$:

$M = 3.750/28.850 = 0.130$

$X_s(_{10})_{80} = 29.180$ ohms

Fifteen Meter Trap (f_o=*21.225* MHz; Z_p=2.143x10^4 ohms)

On $f_o(_{20})$:

$M = 14.175/21.225 = 0.668$

$X_s(_{15})_{20} = 258.345$ ohms

On $f_o(_{40})$:

$M = 7.150/21.225 = 0.337$

$X_s(_{15})_{40} = 81.431$ ohms

On $f_o\ (_{80})$:

$M = 3.750/21.225 = 0.177$

$X_s(_{15})_{80} = 39.082$ ohms

Twenty Meter Trap (f_o=*14.175* MHz; Z_p=2.246x10^4 ohms)

On $f_o(_{40})$:

$M = 7.150/14.175 = 0.504$

$X_s(_{20})_{40} = 151.883$ ohms

On $f_o\ (_{80})$:

$M = 3.750/14.175 = 0.265$

$X_s(_{20})_{80} = 63.861$ ohms

Forty Meter Trap (f_o=*7.150* MHz; Z_p=2.226x10^4 ohms)

On $f_o\ (_{80})$:

$M = 3.750/7.150 = 0.524$

$X_s(_{40})_{80} = 161.049$ ohms

To explore the $jX_{in}(_{1,\ 2})$ on the band edges, repeat the above calculating process, but use $M = f_{low}/f_o$ or f_{high}/f_o, to get the *change in X_s at each band frequency limit.*

Monopole/ Transmission Line Characteristic Impedance K_m

When we explored the two monopoles in part I, you will recall that we obtained two different values of K_m; for our number 10 gauge wire monopole conductor, $K_m(_1)$ was 560.32 ohms; for our tubing conductor of 2.0 inch radius, $K_m(2)$ was 340.10 ohms in value. These values of K_m apply *only* to our two monopoles when they are operated in the eighty meter band. When we shift our design to any of the other ham bands, we must calculate new K_m values for the characteristic impedance of our antenna analogue transmission line, even though we are still employing the same two conductors of fixed physical conductor radii. To understand this, recall that in Schelkunoff's equation (1.0-1.) everything is a constant except the ratio 2(h)/a. Now, a "monopole" in a Morgan antenna is *only* that portion of the total antenna extending from the base input terminals up to the trap which happens to be resonant at a given ham band. Because the r.f. wavelength λ_o at each new band center is different, that 2(h)/a ratio in equation (1.0-1.) keeps changing band-by-band. To account for this, we just repeat the steps we went through in part I to get K_m on eighty meters:

A. Calculate $\lambda_o'=984/f_o$(MHz) in the given band.

B. Get the conductor radius in degrees as $a^\circ=a'/\lambda_o'$.

C. Plug the new value of a_1 or a_2 into (1.0-1), letting 2(h) always equal $2(90)^\circ$.

When we do this, band-by-band, still using our old conductor radii of a_1=4.245x10^{-3} ft. and a_2=1.667 x10^{-1} ft., we get the resulting table of K_m values for each band and each monopole:

a_1=4.245x10^{-3} ft.	a_2=1.667x10^{-1} ft.
$K_m(_1)_{80}$=560.32 ohms	$K_m(_2)_{80}$=340.10 ohms
$K_m(_1)_{40}$=521.60 ohms	$K_m(_2)_{40}$=301.37 ohms
$K_m(_1)_{20}$=480.54 ohms	$K_m(_2)_{20}$=260.31 ohms
$K_m(_1)_{15}$=456.32 ohms	$K_m(_2)_{15}$=236.10 ohms
$K_m(_1)_{10}$=437.90 ohms	$K_m(_2)_{10}$=217.67 ohms

Notice that the *average* characteristic impedance of the monopole analogue line is largest in value at the very lowest band frequency $f_0(_{80})$, and smallest at our highest band frequency $f_0(_{10})$. If we had included $f_0(_6)$ it would have dropped in value even lower. Alright! That does it! We now have a complete list of preliminary design parameters for our multi-band Morgan. Sure we've worked a bit to get them, but now its all down hill, and will be fun the rest of the way.

We will now use these tabulated values to *climb* up our monopole, band-by-band, to obtain only one unknown value: the electrical length of the *last* top conductor section which ends in the resonant trap on a given band. We did this "inadvertantly" in part I to get a two-band Morgan. The only difference between our old two-band Morgan and this new five bander we're about to tackle is in the number of conductors and traps we must juggle. We don't want to do this in some careless way which can get us mixed up.

A Check Out Diagram

In science, when we deal with a problem involving a number of different values of constants and a range of variables, it's kind of nice to have a map or diagram of each step needed along the way to avoid making errors; a sort of simple computer program; after all, the pro's use them to keep from getting mixed up, so why shouldn't we?

Fig. 1 shows a general diagram of the Morgan antenna in the form needed for us to climb up the monopole length sections and also inch around each off-resonant trap reactance X_s, until we reach and obtain the electrical length of the last conductor section in that band's active monopole antenna. It isn't as bad as it looks at first glance (nothing ever is!). Down at the bottom, the very first ten meter band conductor section is labeled $h^\circ(_{10})$. Then, at the top end of conductor $h^\circ(_{10})$, the ten meter band

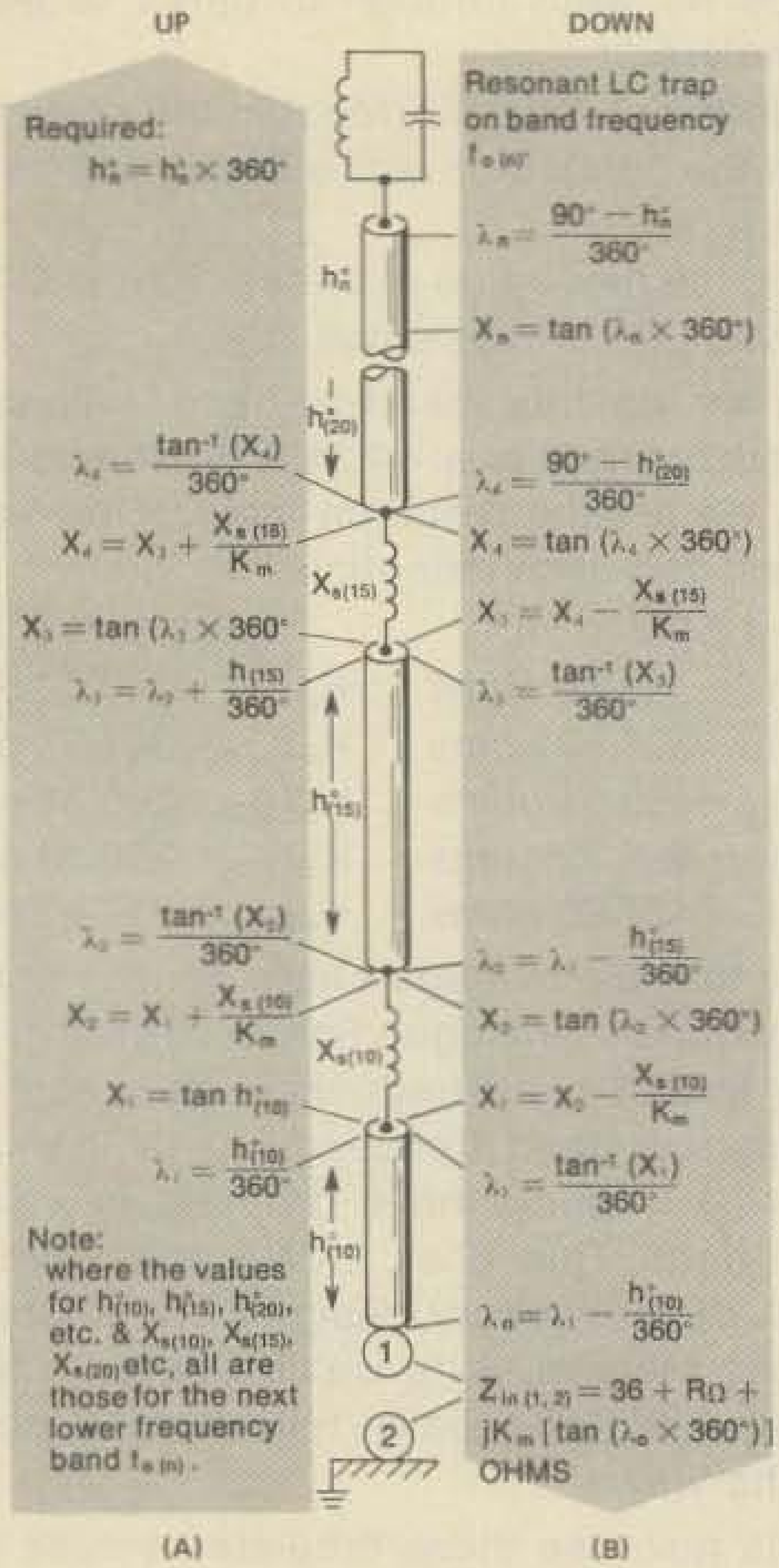

Fig. 1—(a) Successive steps for "climbing" Morgan monopole in a given ham band to determine last top conductor electrical length $h^\circ_{(n)}$ needed to establish monopole resonance at a given band frequency $f_o(_n)$. (b) Successive steps for "descending" monopole, after length $h^\circ_{(n)}$ has been found, too determine input impedance $Z_{in}(_{1,2})$.

trap is shown as a series inductive reactance $X_{(10)}$, representing it as out of resonance at our operating frequency $f_o(_n)$. Above the ten meter band trap extends another conductor for fifteen meter band operation labeled $h^\circ_{(15)}$; at its top is the fifteen meter band trap, again shown as non resonant at $f_o(_n)$ thereby inserting more inductive reactance $X_s(_{15})$ in series with our monopole. Above the non-resonant fifteen meter trap, a short part of the next conductor section $h^\circ_{(20)}$ is seen. We then use our imagination to realize that $h^\circ_{(20)}$ will also end in the twenty meter band trap, and above that there will be another conductor. It's just the old string of "bones" thing in electrical form.

However, on the one particular band frequency $f_o(_n)$ we are concerned with at the moment, there will be one last conductor of unknown length $h^\circ_{(n)}$ (where n takes on the given band length in meters) which we need to determine in electrical degrees. If this last conductor section turned out to be $h^\circ_{(40)}$, we'd be just one band away from finishing our design task; if $h^\circ_{(n)}$ represented $h^\circ_{(80)}$ we'd get its length in one fell swoop and finish the job. Finally, in fig. 1, there is one band trap which is shown as a parallel LC circuit. That circuit represents the one resonant trap in the band at which we are working.

Along the left hand side of fig. 1 is a label λ_1 at the very top of conductor section $h^\circ_{(10)}$. This label λ_1 represents the distance $\lambda_1 = h^\circ_{(10)}/360^\circ$ from that point on the monopole down to the base terminals 1, 2 in wavelength at the operating frequency. Right near to λ_1 there is a reactance symbol X_1. It represents the reactance of the antenna at the same point in normalized ohms. The little formula next to X_1 says its normalized reactive magnitude is equal to the tangent of $h^\circ_{(10)}$ degrees. Above the ten meter trap another height label λ_2 is shown, and at the same electrical height above the monopole base there is another value of X_2 ohms. Above these points are more λ and X labels, each of them being equal to a little formula which shows how to reach these points in terms of those below. Everything ends, in terms of λ and X, right at the base end of the last conductor section $h^\circ_{(n)}$ remaining below the resonant band trap. All these λ's and X's are our climbing steps needed to obtain the length $h^\circ_{(n)}$ band-by-band.

Finally, over on the right hand side of the figure the same λ and X labels appear, except that now the little formulas are subtractions instead of additions. You guessed it: those notations on the right hand side are used to *climb down* the monopole *after* we have obtained a calculated length for $h^\circ_{(n)}$ in order to (a) check on the accuracy of our answer for $h^\circ_{(n)}$ in a given band in terms of whether $jX_{in}(_{1,2})$ comes out very close to j0 ohms and (b), to shift frequency in the band over to the edges in order to find out what $jX_{in}(_{1,2})$ calculates to be there so we may find v.s.w.r. in our fifty ohm feed line. As we remember from part I, we first obtained $jX_{in}(_{1,2})$ then put R_r (and even RΩ) in series to get total antenna $Z_{in}(_{1,2})$ and thus find v.s.w.r. Remember; if we don't like the answers we get for band edge v.s.w.r., we can always do something like changing K_m (by selection of different conductor radii) to make things more to our liking. This is the great thing about "paper antennas"; they can be "erased" and reworked without waste of expensive tubing or trap components. Now let's begin climbing our Morgan to design it.

The Ten Meter Band

In using the antenna analogue tool, we always start at the bottom and climb up the antenna to design it; then climb down again to check on the accuracy of our answer. On ten meters we start by finding the electrical length of $h^\circ_{(1)}$ in degrees. This step is so easy we have to restrain ourselves from laughing. In a Morgan trap antenna, as we said, its

(Continued on page 72)

Multi Band *(continued from page 50)*

first monopole section always "theoretically" functions as a naturally resonant $\lambda/4$ monopole antenna (later we will modify that word "naturally" just a little bit). By defining $h^\circ(_{10})$ as being naturally resonant at one quarter wavelength, we know instantly that $h^\circ(_{10})$ equals 90.000 degrees electrically at $f(_{10})$. That does it for ten meters. We're finished (See, we said it would be easy).

However, from here on there will be no more naturally resonant $\lambda/4$ sections in our Morgan on any lower frequency band; just "coil loaded" monopoles whose *total* electrical length of *both* the conductor electrical lengths *and* that added by the traps below $h^\circ(_n)$ will equal 90 degrees at $f_o(_n)$. A "loaded" monopole is *not* naturally resonant when its *total* electrical length is 90 degrees: it is a monopole "forced" into resonance at $f_o(_n)$ by lumped reactance!

The Fifteen Meter Band

When we snap the band switch of the rig to the fifteen meter band and spin the v.f.o. dial over to 21.225 MHz, the ten meter band trap goes out of resonance to present a reactance $X_s(_{10})_{15}$ ohms in series between conductor sections $h^\circ(_{10})$ and $h^\circ(_{15})$. But we know what $X_s\,(_{10})_{15}$ is because of our sweating a bit earlier. All we have to do is find out how long electrically to make $h^\circ(_{15})$ so that with the ten meter trap reactance in series, we still come out resonant at $f(_{15})$. Reaching back to our earlier lists, we obtain the reactive value for $X_s(_{10})_{15}$ as equal to 353.801 ohms. That's neat, but what about the length of conductor $h^\circ(_{10})$ at $f_o(_{15})$? It's simple: we just put our factor M to work. We get $M = 21.225/28.850$ equals 0.736. Therefore $h^\circ(_{10})_{15} = 90^\circ \times 0.736$ equals 66.213 degrees. On *all bands* the converted length of $h^\circ(_{10})$ will be the *same* for both the $K_m(_1)$ and $K_m(_2)$ monopoles. Now, we first list our known data for each monopole then hitch up our belts and begin climbing.

$a'(_1) = 4.245 \times 10^{-3}$ ft.	$a'(_2) = 1.667 \times 10^{-1}$ ft.
$K_m(_1)_{15} = 456.32\Omega$	$K_m(_2)_{15} = 236.10\Omega$
$h^\circ(_{10})_{15} = 66.213^\circ$	$h^\circ(_{10})_{15} = 66.213^\circ$
$X_s(_{10})_{15} = 353.801\Omega$	$X_s(_{10})_{15} = 353.801\Omega$

(We now use the steps shown on the left hand side of fig. 1 to climb the monopole to the height $\lambda 1$.)

$\lambda_1 = h^\circ\,(_{10})_{15}/360^\circ = 66.213^\circ/360^\circ = 0.1839\lambda$
$X_1 = \tan h^\circ\,(_{10})\,_{15} = \tan 66.213^\circ = 2.2687\Omega$
$\lambda_1 = 66.213^\circ/360^\circ = 0.1839\lambda$
$X_1 = \tan 66.213^\circ = 2.2687\Omega$

(We must now climb around the off-resonant ten meter trap reactance by adding its normalized reactance X_s/K_m to X_1.)

$$X_2 = X_1 + \frac{X_s(_{10})_{15}}{K_m(_1)} = 2.2687 + \frac{353.801}{456.32}$$

$$X_2 = 2.2687 + 0.7753 = 3.044\Omega$$

$$X_2 = X_1 + \frac{353.801}{236.10}$$

$$X_2 = 2.2687 + 1.4985 = 3.7672\Omega$$

In this last step we have reached the base end of the last conductor section $h^\circ\,(_{15})$. We notice that the reactance X_2 now differs in value on our two monopoles. To find the needed length $h^\circ(_{15})$ in each of the monopoles, we take the next steps indicated in fig. 1:

$X_2 = 3.0440\Omega$

$$\lambda_2 = \frac{(\tan^{-1} X_2)^\circ}{360^\circ} = \frac{(\tan^{-1} 3.0440)^\circ}{360^\circ}$$

$$\lambda_2 = \frac{(71.8139)^\circ}{360^\circ} = 0.1995\lambda$$

$X_2 = 3.7672\Omega$

$$\lambda_2 = \frac{(\tan^{-1} 3.7672)^\circ}{360^\circ}$$

$$\lambda_2 = \frac{(75.1337)^\circ}{360^\circ} = 0.2087\lambda$$

(then:)

$h\lambda(_{15}) = 0.250\lambda - \lambda_2$

$h\lambda(_{15}) = 0.250 - 0.1995 = 0.0505\lambda$

$h^\circ(_{15}) = 0.0505 \times 360^\circ = 18.186^\circ$

$h\lambda(_{15}) = 0.250\lambda - \lambda 2$

$h\lambda(_{15}) = 0.250 - 0.2087\lambda = 0.0413\lambda$

$h^\circ(_{15}) = 0.0413 \times 360^\circ = 14.866^\circ$

Actually, we could have used another "lazy man's" step just after we had obtained the reactance X_2 at the base of the unknown length section $h^\circ(_{15})$. Because X_2 is in normalized ohms in both monopoles we would find $h^\circ(_{15})$ degrees immediately as:

$h^\circ(_{15}) = (\text{cotan}^{-1} X_2)^\circ$
$h^\circ(_{15}) = (\text{cotan}^{-1} 3.0440)^\circ = 18.186^\circ$
$h^\circ(_{15}) = (\text{cotan}^{-1} X_2)^\circ$
$h^\circ(_{15}) = (\text{cotan}^{-1} 3.7672)^\circ = 14.866^\circ$

You can only pull this lazy man's step* when you finally reach $X(_n)$ of the last conductor section whose length $h^\circ(_n)$ *you need to find* in a given band. Also, you have to get the electrical height λ_n at the base of an intervening conductor of *known* electrical length to climb to its top end. As we haven't done that yet, this OM will stay in the QSO for one more band to make sure that the gang knows not only how to climb around non resonant traps but also shinny up a known conductor length. Bye the bye, notice how the conductor lengths between monopole $K_m(_1)$ and $K_m(_2)$ are changing.

(To Be Continued)

*Recall how we discovered this "lazy man's" trick back in our equation (1.0-5) back in part II when we were taking a $\lambda/4$ monopole apart to see what made it tick.

Conclusion—Getting it all together.

The Multi-Band Trap Antenna—Part IV

BY JOSEPH M. BOYER*, W6UYH

Part I appeared in February, Part II in March, and Part III in April. In this, the concluding part, the author completes his analysis of the Morgan trap antenna.

The Twenty Meter Band

When we shift to the next lower frequency band, twenty, we are savvy to what happens in the antenna now: on 14.175 MHz, *both* the ten and fifteen meter band traps are non-resonant, so two traps are now acting as series "loading coils" in the monopole. But we know these $X_s(_{10})_{20}$ and $X_s(_{15})_{20}$ reactive values. At the same time our first conductor section $h^\circ(_{10})_{20}$ is now 14.175/*28.850* $\times$ 90° equals 44.220 degrees long electrically in *both* our monopoles. However, those two different length $h^\circ(_{15})$ conductors just obtained have to each be multiplied by the common M factor 4.175/*21.225* = 0.668. Doing this gives us $h^\circ(_{15})$ equals 12.145 degrees for the skinny $K_m(_1)$ monopole and $h^\circ(_{15})$ equals 9.928 degrees in our fatter $K_m(_2)$ monopole. Starting our climb up the monopole again to find out what the needed $h^\circ(_{20})$ conductor length must be, we again split our steps into two columns, first listing our known data:

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$a'_1 = 4.245 \times 10^{-3}$ ft.
$K_m(_1)_{20} = 480.540\Omega$
$h^\circ(_{10})_{20} = 44.220^\circ$
$h^\circ(_{15})_{20} = 12.145^\circ$
$X_s(_{10})_{20} = 142.946\Omega$
$X_s(_{15})_{20} = 258.345\Omega$

$a'_2 = 1.667 \times 10^{-1}$ ft.
$K_m(_2)_{20} = 260.310\Omega$
$h^\circ(_{10})_{20} = 44.220^\circ$
$h^\circ(_{15})_{20} = 9.928^\circ$
$X_s(_{10})_{20} = 142.946\Omega$
$X_s(_{15})_{20} = 258.345\Omega$

$\lambda_1 = h^\circ(_{10})_{20}/360^\circ = 44.220^\circ/360^\circ = 0.1228\lambda$
$X_1 = \tan h^\circ(_{10})_{20} = \tan 44.220^\circ = 0.9731\Omega$

$$X_2 = X_1 + \frac{X_s(_{10})_{20}}{K_m(_1)_{20}} = 0.9731 + \frac{142.946}{480.540}$$

$$X_2 = 0.9731 + 0.297 = 1.271\Omega$$

$$\lambda_2 = \frac{(\tan^{-1} X_2)^\circ}{360^\circ} = \frac{(51.796)^\circ}{360^\circ} = 0.1439\lambda$$

$\lambda_1 = 44.220^\circ/360^\circ = 0.1228\lambda$
$X_1 = \tan 44.220^\circ = 0.9731\Omega$

$$X_2 = 0.9731 + \frac{142.946}{260.310}$$

$$X_2 = 0.9731 + 0.549 = 1.522\Omega$$

$$\lambda_2 = \frac{(56.697)^\circ}{360^\circ} = 0.1575\lambda$$

(Right here, everybody take a firm grip, as we are going to shinny up the known lengths of $h^\circ(_{15})_{20}$:)

$$\lambda_3 = \lambda_2 + \frac{h^\circ(_{15})_{20}}{360^\circ} = 0.1439 + \frac{12.145^\circ}{360^\circ}$$

$$\lambda_3 = 0.1575 + \frac{9.928^\circ}{360^\circ}$$

$$\lambda_3 = 0.1776\lambda$$

$$X_3 = \tan(\lambda_3 \times 360)^\circ = \tan 63.936^\circ$$

$$X_3 = 2.0445\Omega$$

$$\lambda_3 = 0.1851\lambda$$

$$X_3 = \tan(\lambda_3 \times 360)^\circ = \tan 66.628^\circ$$

$$X_3 = 2.3140\Omega$$

(Hold it there: stay in reactance because we must now climb around the non resonant fifteen meter trap:)

$$X_4 = X_3 + \frac{X_s(_{15})_{20}}{K_m(_1)_{20}} = 2.0445 + \frac{258.345}{480.54}$$

$$X_4 = 2.0445 + 0.538 = 2.5821\Omega$$

$$X_4 = 2.3140 + \frac{258.345}{260.310}$$

$$X_4 = 2.3140 + 0.992 = 3.3060\Omega$$

(Having reached the base of our unknown length section $h^\circ(_{20})$, we may now use the lazy man's move to obtain the length of $h^\circ(_{20})$ in one additional step:)

$$h^\circ(_{20}) = (\cot^{-1} X_4)^\circ = (\cot^{-1} 2.5821)^\circ$$

$$h^\circ(_{20}) = \underline{21.170^\circ}$$

$$h^\circ(_{20}) = (\cot^{-1} 3.3060)^\circ$$

$$h^\circ(_{20}) = \underline{16.829^\circ}$$

We now see how two non-resonant traps in series, but of fixed reactance value, affect our two monopoles differently due to the K_m effect of the conductor radii. Such effect also applies, of course, to ordinary coil-loaded short monopoles. Notice that in the "computer steps", when you have to climb up an intervening antenna *conductor* of know electrical length, you stay in height λ at the conductor base, then shinny up it by adding *its length* $h_n^\circ/360^\circ$ to its base point height λ. However, when you must climb above a trap you add the X value obtained at the *bottom end* of the trap to *its actual reactance* X_s divided by the monopole K_m (as the X_s/K_m step normalizes trap reactance, you climb around a trap by adding trap reactance to antenna reactance—both in normalized form).

We may use a "score board" in this monopole climbing process to do two things: first, to keep track of our moves in the design and, second, to be able to actually "watch" things happen along the Morgan antenna. Nothing does a better job of "etching" an understanding of antenna function in terms of impedance behavior on our minds (even a pro's) than studying such a "score board". Naturally, our score board is the one invented by P. H. Smith[5], and it is shown in fig. 2.0. It is identical to the one used in our part I work, except that the impedance R+jX at its center is now 1.0+j0 instead of 50+j0 ohms. Due to this, all impedance on the chart is normalized so it may be used for a transmission line of *any* characteristic impedance Z_o or K_m. Here we will use it to *simultaneously* represent our $K_m(_1)$ and $K_m(_2)$ characteristic impedances. Because we are now familiar with normalized impedance and reactance, only a word is needed to understand the data presented in fig. 2. Each of our climbing steps along our two twenty meter monopoles are shown in normalized form, along the left hand side sector of the chart. As we are not letting our monopoles radiate or have ohmic loss (R_r = $R\Omega$ = 0 ohms), all reactance is shown as points along the inside rim edge (+jX) as in part I. All heights λ along the monopole at increasing distance from the base are listed on the Wavelengths Toward Generator (W.T.G.) outermost distance scale. Again the chart is printed "upside down".

The short radial lines projecting outside the chart represent our steps for the skinny $K_m(_1)$ wire monopole; those projecting inward into the chart are for our fatter $K_m(_2)$ monopole. Notice that both the outward and inward pointing lines touch both the W.T.G. distance scale and also the inside rim reactance scale. Also, observe that both monopoles are "neck-to-neck" at the top of the $h^\circ(_{10})_{20}$ conductor section, but don't stay that way farther on up the antennas. Finally, over on the right hand side sector of the chart, the *climbing down* steps are shown as a check *after* we've found $h^\circ(_{20})$ to make sure we end up at $jX_{in}(_{1,2})$ = j0 ohms on $f_o(_{20})$. On the band edges, our part I work tells us now that at the f_{low} band edge that circumferential distance from λ_1 on the right down to the monopole base would come out *short* of j0, to end up a short distance on the *right hand* capacitive reactance region of input impedance $-jX_{in}$. Conversely, at f_{high}, that same arc distance would inch *beyond* j0, to end in the inductive $+jX_{in}$ region. Finally, we remember that to get those normalized reactances $jX_{in}(_{1,2})$ "out" of the chart, we just multiply them either by $K_m(_1)$ or $K_m(2)$ to convert them to *actual values* of reactance (or impedance). Then, to get v.s.w.r. in a fifty ohm feed coax, we'd just add R_r (and $R\Omega$ if we wish) to these *actual reactance* values and plot them on a 50+j0 ohm Smith chart to get v.s.w.r. in such cable versus operating frequency.

Now that the gang is getting experienced in climbing up the monopole on any band, this OM will leave the QSO while the gang works out the remaining conductor lengths for forty and eighty, and QRX for a traffic sked. However, we'll be back on 3.750 MHz at 1600 hours GMT to compare a final list of calculated conductor lengths for each monopole with those obtained by the fraternal brothers.

[5] P. H. Smith, "Transmission Line Calculator," *Electronics, 12*, pp. 29-31, January, 1939. Note: Printed pads of Smith chart blank forms are available from General Radio, West Concord, MA 01781. The 50 ohm and normalized types are most useful to amateurs.

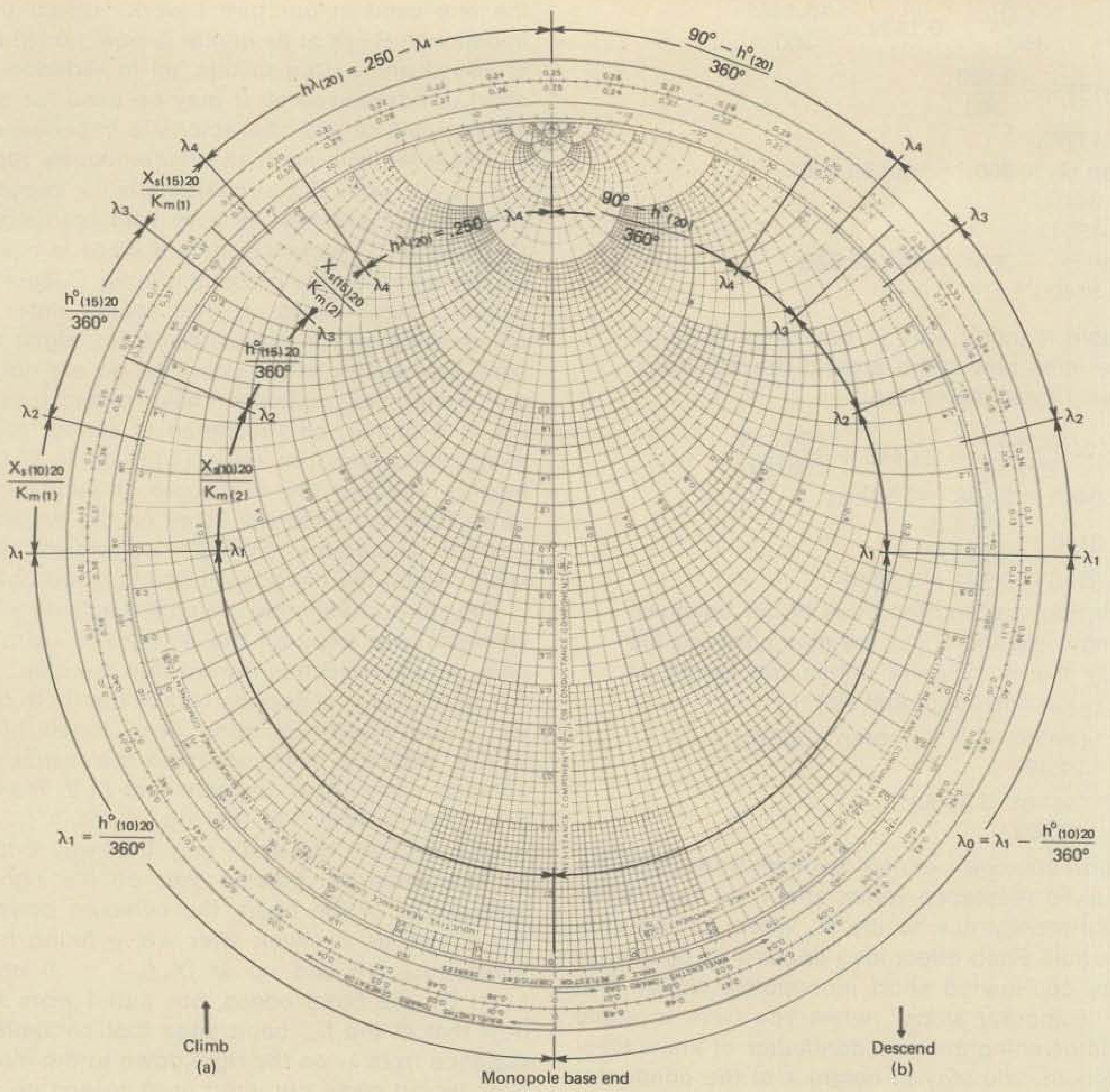

Fig. 2—Normalized Smith impedance chart showing analogue design steps for twenty meter band monopole portion of ten to eighty meter Morgan multi-band antennas. Analogue characteristic impedance $KM_{(1)20}$ is 480.54 ohms for #10 gauge wire; $Km_{(2)20}$ is 260.31 ohms for four-inch diameter conductor. See text.

Eighty Meters

Well we see everybody finally reached the base end of conductor $h°_{(80)}$ and used the lazy man's step to find its electrical length for each of the two monopoles. Our paper Morgan antenna design is complete, so let's all relax, compare notes, and chew the fat. The conductor lengths we obtained from a table. Each needed conductor length is listed in electrical degrees and electrical wavelengths at the band f_o *at which it was first found.* The list should look like the one given below, which is *based solely* on use of the particular band trap parameters selected at the start of design.

$a_1' = 4.245 \times 10^{-3}$ ft.

$h°_{(10)} = 90.000°$;$h\lambda_{(10)} = 0.2500\lambda$

$h°_{(15)} = 18.186°$;$h\lambda_{(15)} = 0.0505\lambda$

$h°_{(20)} = 21.170°$;$h\lambda_{(20)} = 0.0588\lambda$

$h°_{(40)} = 33.900°$;$h\lambda_{(40)} = 0.0942\lambda$

$h°_{(80)} = 34.346°$;$h\lambda_{(80)} = 0.0954\lambda$

$a_2' = 1.667 \times 10^{-1}$ ft.

$h°_{(10)} = 90.000°$;$h\lambda_{(10)} = 0.2500\lambda$

$h°_{(15)} = 14.866°$;$h\lambda_{(15)} = 0.0413\lambda$

$h°_{(20)} = 16.829°$;$h\lambda_{(20)} = 0.0467\lambda$

$h°_{(40)} = 27.881°$;$h\lambda_{(40)} = 0.0774\lambda$

$h°_{(80)} = 29.972°$;$h\lambda_{(80)} = 0.0833\lambda$

In comparing calculated conductor lengths to those above, the gang need not be concerned about small differences in lengths obtained in their work; it's ok if agreement is within say ± 1.0 degrees or so, for reasons we will see shortly. If we go back and add up *all conductor* electrical lengths in a given Morgan monopole band section, we see that they do not total up to 90.00 electrical degrees; however, in each case if we then *add* the electrical lengths contributed by *each trap in that band,* the total antenna electrical length then *does equal 90.00* degrees and produces "forced" resonance except on Ten meters. We now see that it is the trap "coil loading" effect *plus* the influence of the conductor related to K_m, which produces substantial *conductor shortening* in *electrical degrees*.

Although such antenna conductor "shrinkage" is inevitable in a lumped LC trap multi-band antenna, we shall shortly see that we can't let this conductor miniaturization *go too far* if we wish to obtain optimum on-the-air performance from our Morgan antenna.

Conversion From Electrical To Physical Conductor Length

Now it was said that the conductor lengths we obtained by using the antenna analogue steps came out in electrical degrees at the band f_o. This means that the $h\lambda(\ )$ lengths listed apply to dimensions of wavelength λ_o in free space. Any real antenna conductor of finite physical diameter $d = 2a$ will be actually shorter in physical length than the free space wavelength dimension given, in inverse proportion to the conductor diameter. Unfortunately, space does not permit a discussion of the very interesting theory behind such conductor shortening effect. Here, we can only give the steps needed to convert those calculated conductor lengths from electrical to physical length.

(a) Starting at ten meters, then moving progressively to the center of the next lower ham band, compute the free space wavelength $\lambda_o = 984.00/f_o$ (MHz) in feet.
(b) Multiply your particular conductor radius (ft) by two to get its physical diameter $d = 2a$ in feet.
(c) Divide your conductor diameter into the λ_o' for the ham band being considered.
(d) Use this λ_o'/d' number obtained from step (c) to enter the Wavelength in Diameter horizontal bottom scale of fig. 3.0.
(e) Move up the chart at the entered value until you intercept the graphed curve of the figure.
(f) Move horizontally at that found height on the curve to find the conductor length correction factor P calibrated along the left hand side scale of the figure.
(g) The corrected physical conductor length in feet for that particular band is then: $S = P \times h\lambda(\ ) \times \lambda_o$.

As an example, take the $h\lambda(_{10}) = 0.250\ \lambda$ length given for each monopole at ten meters:

$\lambda_o'(_{10}) = 34.107$ ft.
$d' = 2x\ 4.245 \times 10^{-3}$ ft.
$d' = 8.49 \times 10^{-3}$ ft.
$\lambda_o'(_{10})/8.49 \times 10^{-3} = 4{,}017.$
$\lambda_o'(_{10}) = 34.107$ ft.
$d' = 2x\ 1.667 \times 10^{-1}$ ft.
$d' = 3.33 \times 10^{-1}$ ft.
$\lambda_o'(_{10})/3.33 \times 10^{-1} = 102.$

We see immediately that the λ/d scale in fig. 3.0 ends at a maximum value of 3,000. This means that even on ten meters, the λ/d value obtained for a number 10 gauge wire is so large that we do not need to make a correction between its electrical and physical length* in *practical* antenna design. We just let $P = 1.0$ to get the conductor length S in feet for the skinny $K_m(_1)$ monopole. For number 10 gauge wire, the λ_o'/d' ratio gets increasing larger at the lower frequency bands so no correction is needed at lower freqencies either. The fatter monopole, however, gives a λ_o'/d ratio of only 102. Entering figure 3.0 at this value, we "eyeball" a P factor of about 0.89 on ten meters. Therefore, our four-inch diameter (twice the value of *a*) tubing monopole of $K_m(_2)$ comes out as $S' = 0.89 \times 0.250 \times 34.107$ ft $= 7.6$ feet in length to form a resonant $\lambda/4$ monopole.

But we had to "eyeball" a value off the chart, which implies that we might have been a bit foolish in using even three decimal place accuracy in our design calculation steps. Actually, we were not being foolish: we should always try to obtain accuracy in design calculations based on theory. Later we'll see how to get our calculated values "right on" when we make our Morgan antenna play.

The Real World Of Antennas

Up to now we have been climbing up and down our Morgan antenna, designing it with the idea that our band traps function like ideal lumped LC circuits. But earlier it was said that ideal circuits have zero electrical size in λ at the operating frequency!

*Amateur antenna handbooks give a length correction factor of about 0.950 to apply to antenna conductors. This factor applies *only* to the conductor-shortening effect produced by antenna end insulators. It is valid, but it is an entirely different length correction than discussed here.

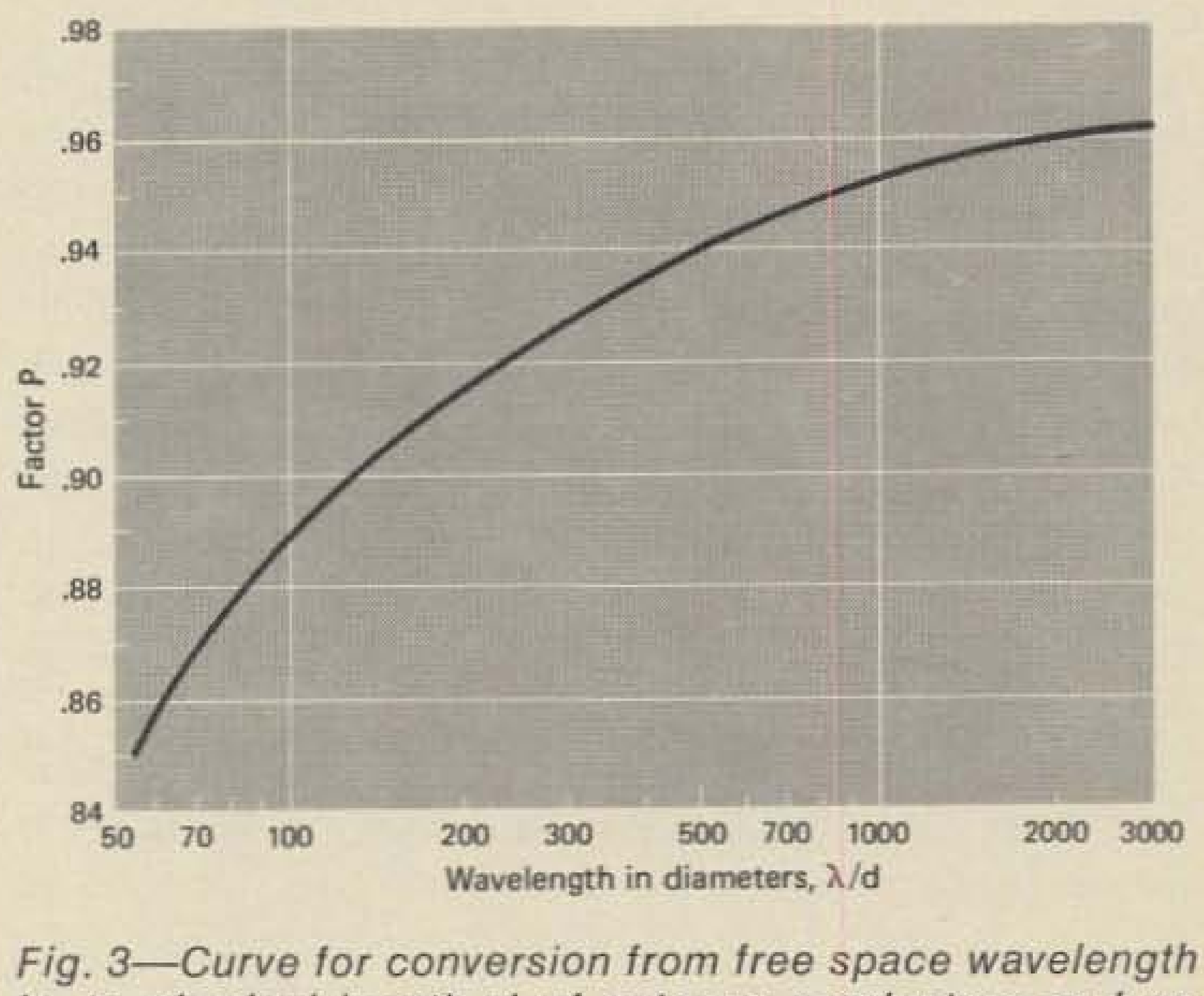

Fig. 3—Curve for conversion from free space wavelength hλ to physical length s' of antenna conductors as function of wavelength to diameter ratio.

If we constructed even the ten meter band trap with its 25 picofarad capacitor and 1.22 microhenry coil we'd find it was not exactly microscopic in size. The eighty meter band trap would constitute a good handful. Here is what happens when we use such traps in a real Morgan antenna:

First, that hunk of trap conductor geometry sitting on the top of a resonant monopole section of the Morgan has a substantial capacity over to the ground plane (or to the other side of a doublet) and will act like a top "capacity hat" on the monopole. This throws that monopole section out of resonance, by making it too long electrically from "top loading"; here is the reason we made a conditional remark about obtaining *natural* $\lambda/4$ monopole resonance on ten meters. There will also be an additional capacity existing from each of the coil turns and from the capacitor frame and plates over to the conductors connected above *and* below the resonant band trap. Such capacity can reduce the "isolation" effect of the trap in "cutting off" the rest of the Morgan above it, so that some *energy* can be coupled to the rest of the antenna. The effect is very complicated and subtle, and dependant upon what the actual impedance is existing at the *upper terminal* of the trap at that frequency. When substantial, this effect can excite a "long wire" mode in the upper parts of the Morgan to generate spurious high angle radiation pattern lobes. Such effect, if present, influences the gain performance of the Morgan primarily at the higher frequency bands of coverage.

Now, what if we played games with the trap parameters? Ok, let's do. The first thing we would find is that our instinctive desire for extremely high trap Q is somewhat misleading. Increasing coil Q does increase Z_p at in-band frequencies *very close* to the trap resonant frequency f_o. Yet even at the frequency limits of a given active ham band, Z_p falls back down to impedance values presented by lower Q traps *using the same L to C ratio.* For example, our ten meter band trap of 100 Q falls to a Z_p of 3.7×10^3 ohms at 28.00 MHz. A trap having the same L and C, but using a 300 Q coil will still fall to almost an identical Z_p at 28.00 MHz. Well, does that mean we were stupid in asking for super high Q air or vacuum condensers? No, because when we go to less ideal condensers we insert a condenser loss R_C and coil loss R_L of unequal value into the trap parallel circuit. This changes the Z_p/f curve in an undesirable manner, as well as adding to ohmic loss. Incidentaly, Z_p at f_o changes in proportion to the Q ratio in compared traps.

What about changing the L to C ratio? Well now that does produce a major effect in the Morgan. If we *increase* capacity C, the trap Z_p is reduced in magnitude at f_o (which is not too important) and also reduces the magnitude of reactance X_s of the trap in lower frequency bands. Now this trap reactance decrease is in proportion to how much we increase capacity ratio in the compared traps. However, if we *lower* the capacity C, we increase the magnitude of the non-resonant trap reactance X_s. This can cause problems. We have clearly seen the shortening effect produced by the trap X_s on the conductor length. A substantial reduction in trap C can really shorten conductor electrical length. At first thought, we might think this a great idea: boy, a miniaturized height, all band antenna!

Unfortunately, Mother Nature is always standing over us with a big club in her hand waiting to bash our technical heads in. Only conductors and their current distribution contribute to antenna radiation, and thus R_r magnitude; traps don't! When conductor electrical length shortening becomes substantial, the radiation resistance R_r of the active Morgan monopole section falls to values less than the 36 ohms of a naturally resonant $\lambda/4$ monopole. We see that with our antenna operating in a fixed ohmic loss environment, the $R\Omega$ remains constant but R_r now moves down closer to it in value. The antenna radiation efficiency is,

$$\text{Efficiency} = \frac{R_r}{R_r + R\Omega} \times 100 \quad \text{Percent.}$$

From this relation we see that the amount of input power to the antenna which is converted to radiated signals decreases in such miniaturized height antenna case. But say we went all out and used a $\lambda/2$ diameter wire ground plane at f_o, with a lot of wire radials and had a QTH in the middle of a salt marsh! Then $R\Omega$ goes down, but that lower value of R_r now will decrease the antenna frequency bandwidth in terms of impedance variation for a given conductor K_m. If one is a devoted c.w. or phone man, reduced bandwidth may not mean much. However, commercials can lay down super dense and large diameter wire ground planes: few hams can. Therefore, small R_r value antennas pose difficulties in ham radio unless certain newer type antennas of military origin are employed. We can't get into a QSO about those at this time. Therefore, in a Morgan of good on-the-air performance we should choose trap L to C ratio to keep the total antenna height from being sharply reduced, even though the idea may appear attractive.

Putting A Morgan On The Air

One of the gang breaks in at this point and says, "If all those weird things go on in the actual antenna from effects which we didn't or couldn't figure in, what good does all this fancy antenna design buy us?" Good question! What careful preliminary antenna design does is to place us squarely in the right ball park in terms of dimensions of conductors, trap parameters and so forth. We then face the same problem seen by pro's who must make real antennas play in a real world QTH. We "prune" to make the antenna perform correctly, guided by our common sense and what we know technically about the way the antenna functions from our study of antenna theory. This approach is one hundred decibels better than the blind groping of the old cut-and try-school! Now let's make our Morgan play.

Say we have chosen the fatter conductor Morgan because its calculated design results look better on paper. We build our traps* and carefully "tweak" them on the bench to the f_o of each band, using a *loosely* coupled grid-dip oscillator and the shack receiver as a frequency meter for the g.d.o. signal. We then cut our tubing to the corrected lengths found, put down a radial wire ground plane either on the soil or on a roof location, and install a low mass ceramic base insulator. Here is what we then do to compensate experimentally for all the "weird" effects:

(1) Stretch out a length of coax reaching from the rig in the shack to the antenna installation location. Connect one end of such coax to the receiver in the shack, and station a friend there to track the T4 note of the g.d.o. on the calibrated receiver dial. On the other end of the coax solder a temporary wire pig tail to the inner conductor to act as a pick up antenna for the g.d.o. signal. Ground the shield of the coax to a radial wire.

(2) Erect *only* the $h^\circ{}_{(10)}$ conductor section on the base insulator, holding it to a vertical position with a temporary "glas line" guying system. Solder a jumper wire from a lug connected to the base of section $h^\circ{}_{(10)}$ (terminal #1) to a ground radial wire directly below. Fire up the g.d.o. and shack receiver in the ten meter band and make sure the g.d.o. signal can be heard on the remote receiver. The b.f.o. helps.

(3) Install the ten meter trap at the top of the conductor section, but do not bother trying to tune it at this time. We now have a top "capacity hat" ten meter monopole.

(4) *Lightly* couple the g.d.o. to the monopole base grounding wire and find the dip in the ten meter band. Invariably, it will be on the low frequency side of f_o. Wobble frequency of the g.d.o. slightly to make sure the friend in the shack identifies it in the QRM from other signals. Then, in small steps, guided by how close the dip was from f_o, prune the length of the ten meter conductor section until the dip falls closely on 28.850 MHz. Let the man in the shack give you frequency: don't use the g.d.o. dial for such data.

(5) Install the $h^\circ{}_{(15)}$ conductor section in place above the trap, and connect it to the other trap terminal. Again use a glas line guy set to hold it in place if necessary. Now bring the g.d.o. dip back on f_o solely by tuning the ten meter band trap. Coil turn bending or turn pruning may be needed. An air padder variable is nice here, but be sure plate spacing is adequate for power input level. As a low loss dielectric weather housing is a must around good quality band traps, make sure you put such cover *back in place* before the man at the receiver lets you know you have reached the ten meter f_o.

(6) Shift the g.d.o. and shack receiver to the fifteen meter band. Install the fifteen meter trap as a "top capacity hat" at the top of the $h^\circ{}_{(15)}$ conductor section. Repeat steps (3) to (5) on the fifteen meter band, as well as on each lower frequency band of coverage until the Morgan is resonant on $f_{o(80)}$.

When you have finished, and have the Morgan working correctly on all its bands of coverage, do not ruin all your good work by installing a set of conventional wire guys even if such wire guys are broken into short sections by compression egg insulators. It's capacitive effect on the Morgan will cause a shift of the antenna resonance. Instead use glas line rope for the final guying system. Even with it, however, install an egg compression insulator at both the antenna attachment end and about midway down the length of the guy to prevent influence on the antenna in a down pour of rain. You might find it interesting to then carefully measure the v.s.w.r. reaches the magic 2:1 circle. Carefully log this data, and compare it to the results obtained when climbing down your paper Morgan on the Smith chart at the band limits. Good DX hunting. ■

*In a Morgan monopole you have to use a non-conducting member to carry mechanical stress around the gaps in the conductor needed for insertion of the band traps. Marine junk yards are a source of husky Navy-type cylindrical insulators of high-quality ceramic removed from old ship antennas. They come in diameters up to several inches, and may be several feet in length. Use them as a core around which to build the trap. The ends of an insulator are inserted in the upper and lower antenna conductor.