QEX August 42 1985 $1

The ARRL Experimenters' Exchange

Assistant Technical Coordinators

As a reader of **QEX**, you should be interested in what's happening with the ATC program (and what's not). Also, as an amateur who is interested in the things technical you are probably an excellent candidate to become an ATC and share your expertise with others in your area.

As background, in January, 1985, the ARRL Board of Directors created the station-level post of Assistant Technical Coordinator. Several things prompted this. One was that a number of the then OO/RFI Coordinators recommended that someone else handle RFI. The section Technical Coordinator seemed a likely candidate. But the TC had no troops below section level. Why not RFI helpers level? The second was a need to move the Technical Information Service (TIS) from ARRL Hq to the field. TIS has been an unsolved problem from the founding of ARRL Hq: tons of mail, backlog, good answers but weeks too late, impersonality of letters and phone calls, impracticability of providing tutoring. Why not set up a system of ATCs who could not only handle RFI problems but any technical question? If a given ATC doesn't have the expertise, then he/she could refer the person to someone who does. If the guru is local, the next question and those that follow could be answered quickly. Good idea?

If you want some more background on the ATC program, see the article by Rick Palm, K1CE, on page 55 of the April 1985 issue of **QST**.

So why are we telling you all this? Your ARRL section needs volunteers for the ATC role. As an ATC you will become an important resource person in your community -- someone to help others when they're over their heads technically. Your advice can make the difference between a new Novice getting on the air or getting stuck with what seems to be an insurmountable problem worthy of hanging up one's earphones. You can talk people back on the air after an equipment malfunction or an attack of the RFI monster. These are the basic responsibilities of the ATC job. Of course, the sky's the limit for the truly effective ATC: speaking at club meetings and hamfests, lining up others to speak, organizing group technical projects, working with TVI committees, liaising with CATV companies on leakage matters, etc.

We'd like to congratulate Mort Cohan, KA1IU, Technical Coordinator for Eastern Massachusetts, for his fine start. He has already lined up sixteen ATCs to handle a wide variety of subjects. Of the first eight TIS letters referred to them from Hq, seven were duck soup.

What happens if an ATC can't answer a question? He/she can put the person in touch with another ATC or a TC who knows the answer. If that doesn't work, other field resources are available -- the ARRL Technical Advisor (TA). TAs aren't considered part of the formal field organization but were established to advise the Hq Technical Department on questions relating to technical articles. So, we can't say whether TAs would be able to respond to questions from other sources, but we plan to ask them. If the ATC, TC and perhaps a TA can't answer the question, it can be bucked to the ARRL Technical Department staff. Failing that, an inquiry can be made in **QEX**. We've had good luck with people getting answers that were right on the money.

When the ATC program was approved, the Hq staff began thinking about a computer listing of ATCs and other technical-inquiry resources. That grew into what we're now calling the **Field Resources Directory**. It is to have a computer listing of people in the field, industry listings and helpful information. What technical information would you like to see included? -- W4RI

Correspondence

Remote Transmitters and Receivers

Can you refer me to articles or information regarding the remoting of transmitters and receivers? There must be thousands of amateurs who live in antenna restricted areas such as small city lots, apartments, and condos. With radio controlled model boats, airplanes, and so on, I should think that the principle could be employed for several tuning functions.

I live on a small lot retirement village measuring 55 x 105 feet. I have several compromise antennas for 160 to 10 meters, but I can view (enviously) many acres of farm land from my roof! — George G. Scott, W2LFX, 34 Chatham Place, Vincentown, NJ 08088.

Another Look At ACSSB

The following is a brief summary of my activity with the ACSSB boards. So far, I haven't wrecked them...

Both the RF and audio boards checked out OK using an old two-meter VFO. After managing to split the transmit/receive circuit for full-duplex operation, I started to put together a TCXO with the intention of building two, one for transmit and one for receive. When May **QST** showed up, I quit building the TCXO. The circuit by John Reed, W6IOJ, on page 14, is exactly what I was looking for. I've already started constructing it!

John's up-converter amplifier may be the one I will use also, but I want to study it a little more before deciding. I don't expect to be using the circuit on OSCAR until winter. If all goes well, I should have it working to my satisfaction by then. There is nothing I would like better than to work on this circuit full time, but my priorities at times are elsewhere.

I've been keeping a log on the circuit activity with drawing changes and hopes of writing it up for others to use. I'll try to keep you posted. -- Roger Laroche, N6FOU, 1155 Hetrick Ave., Arroyo Grande, CA 93420.

I finally had some time to dig into the two ACSSB boards I received from the ARRL. The older board follows the schematic and I can understand it, but the newer one departs quite a bit from it. I find it difficult to tie it all together. I wrote to the factory over a month ago, but have since learned that the Sideband Technology Corp. has gone out of business.

In the newer version, the manufacturer switched to single gate FETs and have changed the AGC control on the first RF stage in the receiver among other things. If you have a schematic for the new board, I would be glad to pay for photocopies. — Bill Brosseau, KF6WM, 45300 Royal Dr., King City, CA 93930.

Feedback

On page 6 of the June 1985 issue of **QEX**, Robert Ball, WB8WGA, presents a short article, "TAPR TNC Modification for 12 V Use." Something was missing, however — his address. It is: 830 Riva Ridge Blvd., Gahanna, OH 43230.

Wire Transmission Lines Revisited

First let me thank you for publishing my request for impedance equations for the four and five wire transmission lines in the February 1985 issue of **QEX** (no. 36). The response was great (overwhelming really). I am enclosing a copy of some of the equations I received, as well as two letters, from Peter Sypher, KC4SI, who derived the equations for the four and five wire lines. I think some of you might like to take a look at his work. The math is over my head, but I can use the equations.

Perhaps a brief explanation of why I wanted the information would be in order. I live in a high rise apartment building in Honolulu and my space and ability to do any antenna work is very limited. The lines will be used as matching stubs on 2 meters and those with large center conductors might be used as support elements as well. There is now TV channel 26 here and I have thought of using it as a signal source, but have not gone beyond the idea stage yet. -- Jim Welch, KH6HEP, 419A Atkinson Dr., 1001, Honolulu, HI 96814.

5902 Queenston Street, Apt. E
Springfield, Virginia 22152

March 12, 1985

James W. Welch, KH6HEP
419A Atkinson Dr., #1001
Honolulu, HI 96814

Dear OM,

In response to your letter in the February QEX, I am enclosing some formulas which apply to figures 1 and 3 of your letter, the transmission lines in which all wires have the same diameter. Your transmission line problems aroused my curiosity, as I am an electrical engineer who sometimes deals with data transmission through landlines. I am not a professional worker in rf transmission lines, but I enjoy mathematical manipulations of the type called for in these problems, and that is why I addressed them.

PART 1

The formulas for capacitance per meter and inductance per meter of transmission line, which you need for computing the line impedance and velocity factor are approximate, as derived here, for the following reasons:

1) It is easier to assume the spacing betweenwires is much greater than the wire diameter when calculating capaditance/meter and inductance/meter than to not make this assumption.

2) The fact that charge on the surface of the conductors tends to concentrate on the portion of the wire nearest another conductor of opposite charge is ignored.

We also assume that there is a well-developed skin effect; that there is no current deep inside the conductors. As a consequence, there are no flux linkages deep inside the conductors. The flux linkages we will calculate will involve only the spaces between wires. At the radio frequencies hams use, the assumption of a well-developed skin effect is very accurate.

(continued on page 9)

Voltage-Controlled Capacitor Linearization

By George W. Allen, N1BEP
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Cape Coral, FL 33904

An empirical method has been found to improve the linearization of the relationship between frequency and shaft rotation when using potentiometers to tune voltage-controlled capacitors. An approximate fit to a straight line relationship can be achieved with a simple modification to the usual potentiometer circuit. The method works with most commercial voltage-controlled capacitors as well as Zener diodes used for the same purpose, provided the capacity-voltage function follows a smooth log relationship. Hyper-abrupt diodes, and types which have an inflection (S-shaped curve) in the capacity-voltage curve, are not suited for this type of compensation.[1] The same method has been used from the low megahertz range to 2.5 GHz.

Fig. 1A illustrates a conventional capacitor diode tuning circuit, and Fig. 1B shows the resulting resonate frequency as a function of the potentiometer shaft rotation. The nonlinearity is typical of this method of tuning. Fig. 2A shows the modification to the circuit and 2B displays the resulting curve. The fit is not exactly linear, but substantial improvement can be seen. Parallelled Zener diodes were used to get sufficient capacity to tune the frequency range. The method worked equally well with one or more diodes in parallel.

The approximate match of the two describing functions fails at the extremes of the frequency-rotation curve. This match can be improved at the expense of the frequency coverage by truncating the ends of the curve. This may be done by inserting fixed resistances on either end of the potentiometer, as shown in Fig. 3. Ten turns of the potentiometer now cover a smaller frequency with improved linearity, as shown in curve 2B, region A - A.

Matching the two curves to compensate for the non-linearity is done graphically. The initial frequency range of the circuit is set without compensation to cover the band desired, with generous extension at each end to allow for the truncation. The shunt resistor, R4, is added and plots made for one or two values to determine the resulting correction. Values of R4 should lie between ten and twenty five percent of R1, the potentiometer resistance. A straight line is compared to the best curve, and linearity checked. It should be possible to get within two to three percent for the full capacity change. The curves shown came within about 2.5% for the full frequency range. To further improve the linearity, the range is limited to the center portion of the curve, between sections A - A. This gives a much better fit, for a smaller range. In the example shown, reducing the frequency span from 760 kHz to 460 kHz improved the linearity from 2.7% to 0.5%. This method of linearization does not prevent the use of series diode temperature compensation, where the temperature compensating diode load resistor is R4.[2] R4 is now 10 to 25 percent of the series combination of R1, R2 and R3, which should not be larger than 150 kilohms to limit the effect of back current on temperature. At the same time, minimum values less than about 25 kilohms will show discrete frequency jumps from turn to turn of the resistance winding as the potentiometer is rotated.

The power supply must be well regulated and filtered. The integrated circuit regulators should be bypassed as recommended to prevent any ripple or oscillation from modulating the circuit through the tuning diodes.

Summary

A method has been shown to provide a more linear relationship between applied voltage (shaft rotation) and frequency, for voltage-variable capacitors used as tuning diodes at frequencies from low MHz to low GHz. The method has been applied to commercial variable-capacitance diodes as well as Zener diodes used for the same purpose. In some cases, an improvement in linearity of about ten times may be expected.

References

[1] Motorola RF Device Data, third edition, Motorola Inc., Phoenix, AZ 85036.
[2] Application Note AN-551, Motorola Inc., Phoenix, AZ.
[3] Reference Data for Radio Engineers, sixth edition, Howard W. Sams & Co., Inc., New York.

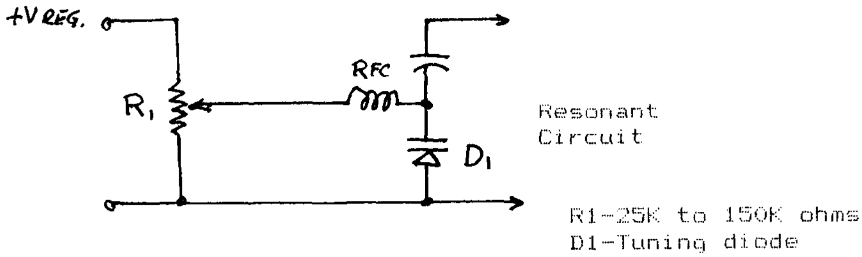

Fig. 1A -- Voltage-Controlled Capacitor tuning circuit.

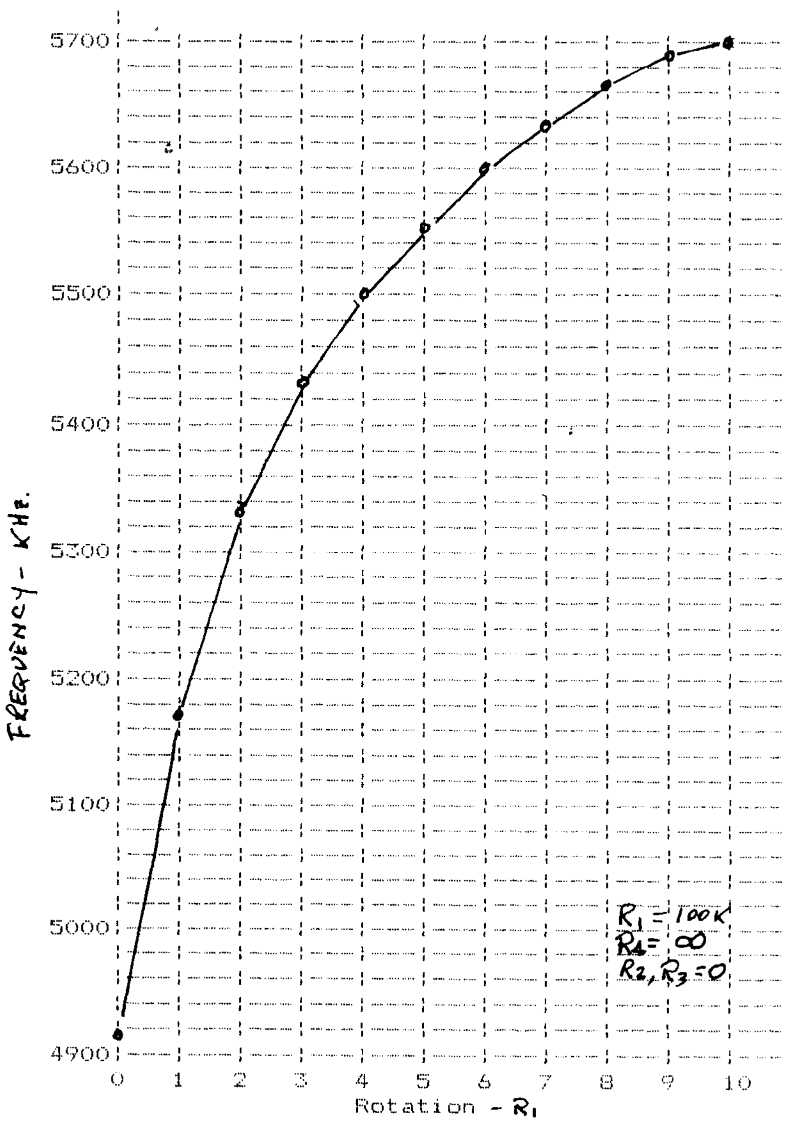

Fig. 1B -- Resonant frequency curve with voltage-variable capacitor tuning.

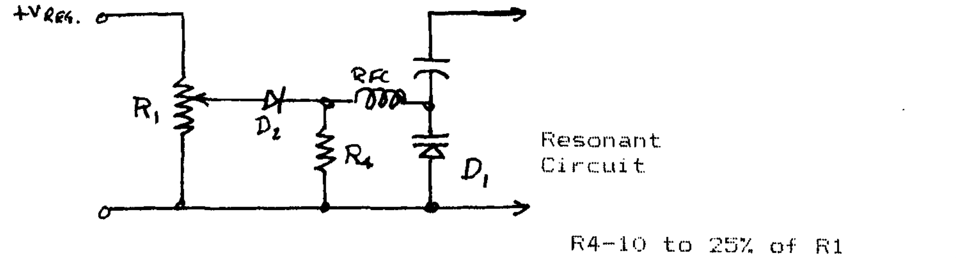

Fig. 2A -- Circuit modified for improved linearity.

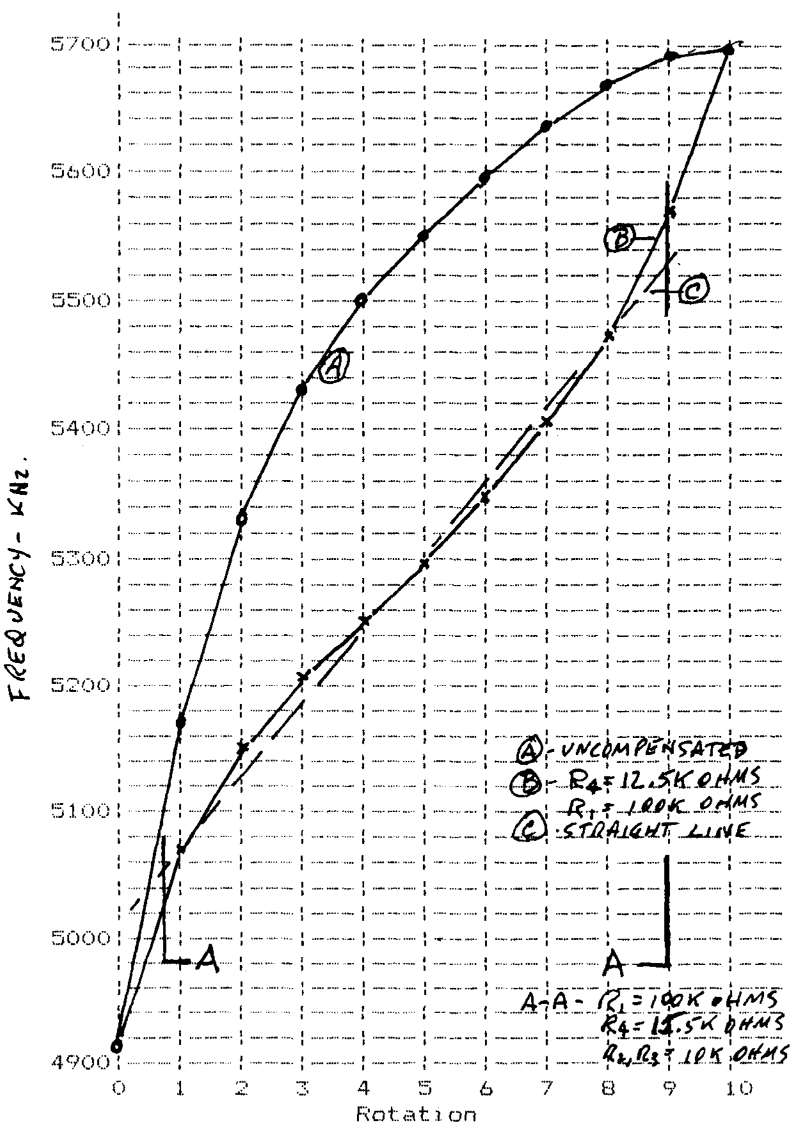

Fig. 2B — Resonant frequency curve with voltage-variable capacitor tuning. Linearity is improved by addition of R4 to reshape applied voltage curve. Section A - A shows most linear center region with ends of the curve truncated.

(Fig. 3 is shown on page 8)

A Variable-Frequency Oscillator for Communication Equipment

By Clint Bowman, W9GLW
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Prospect Heights, IL 60070

An earlier project I challenged using the Motorola MC1648P controlled oscillator led me to develop a VFO which has proven to be satisfactory for use in an amateur transceiver. Unlike the earlier use of the MC1648P, considerable attention was exerted to assure thermal stability of the VFO circuit. In addition, the output was transformed and filtered from square to a reasonably accurate sine wave.

My immediate requirement for a tunable oscillator with high stability and low noise-band products was for the 500 kHz from 6.2 to 6.7 MHz; therefore, the component values included in the schematic are for that range of frequencies. It is probable that little would have to be changed to cover the more common transceiver requirement of 5.0 to 5.5 MHz.

Construction was not particularly difficult. Compact, rigid assembly is suggested for best results. No air-dielectric capacitors were used thus eliminating microphonics from that source. Tuning is accomplished in 10 turns of the panel knob for the 500 kHz. Ideally, frequency calibration could be by electronic counter, however, for my purpose a dial string/pointer combination driven from the 0.250 diameter brass rod epoxied to the Miller no. 4400-6 tuning core has been quite satisfactory for all amateur bands including the SSB/CW portion of 2 meters.

With this combination, a slide-rule scale of approximately 8 inches in length can be fabricated. A piece of spring material cantilevered so as to ride against the brass rod will remove any tendency toward backlash should this become a problem. With this system of tuning, provision must be made for the panel knob to run in and out approximately 5/16 inch as the lead screw of the Miller coil form is turned. I found this to be of no consequence and is a small price to pay in order to eliminate complicated dial mechanisms using gears or planetary drives.

It is extremely important to shield the oscillator circuitry from stray air currents and from any heat producing components such as the voltage regulators. I used a formed box of rigid plastic sheet stock around the oscillator for that purpose. It can be removed easily for alignment and accomplishes its task satisfactorily. If metal is used for a thermal shield keep it at least an inch or more from the circuit.

All capacitors, except those in the power supply, will have an effect on the thermal stability of the oscillator. The 0.0001, 0.01 and 0.1 µF values are miniature Mylar, while the 82-pF capacitor is a Vitramon Porcelain. If available, polystyrene capacitors would be preferred. Radio Shack no. 272-1065 and 272-1069 may be satisfactory for the 0.01 and 0.1 units, although I haven't tried them.

The MC1648P has built-in buffering, and for my application no additional buffering was required when switching from receiver to transmit in a home designed transceiver (**QST**, July 1983, page 24). Additional buffering may be required for some applications, however.

The trimmer capacitors are Radio Shack 5-56 pF devices, the core for the output tuned circuit is Fair-Rite Products balun no. 2865002 402 and the trimming inductor in the main tuning circuit is a green core ferrite device of unknown origin. The inductance range of this coil is best determined experimentally since the ratio between this coil and the Miller main tuning coil will determine the range of frequencies covered. With the combination indicated in the schematic, along with the 5-56 pF trimmer, a coverage of exactly 500 kHz in 10 turns of the panel knob can be had from 6.2 to 6.7 MHz.

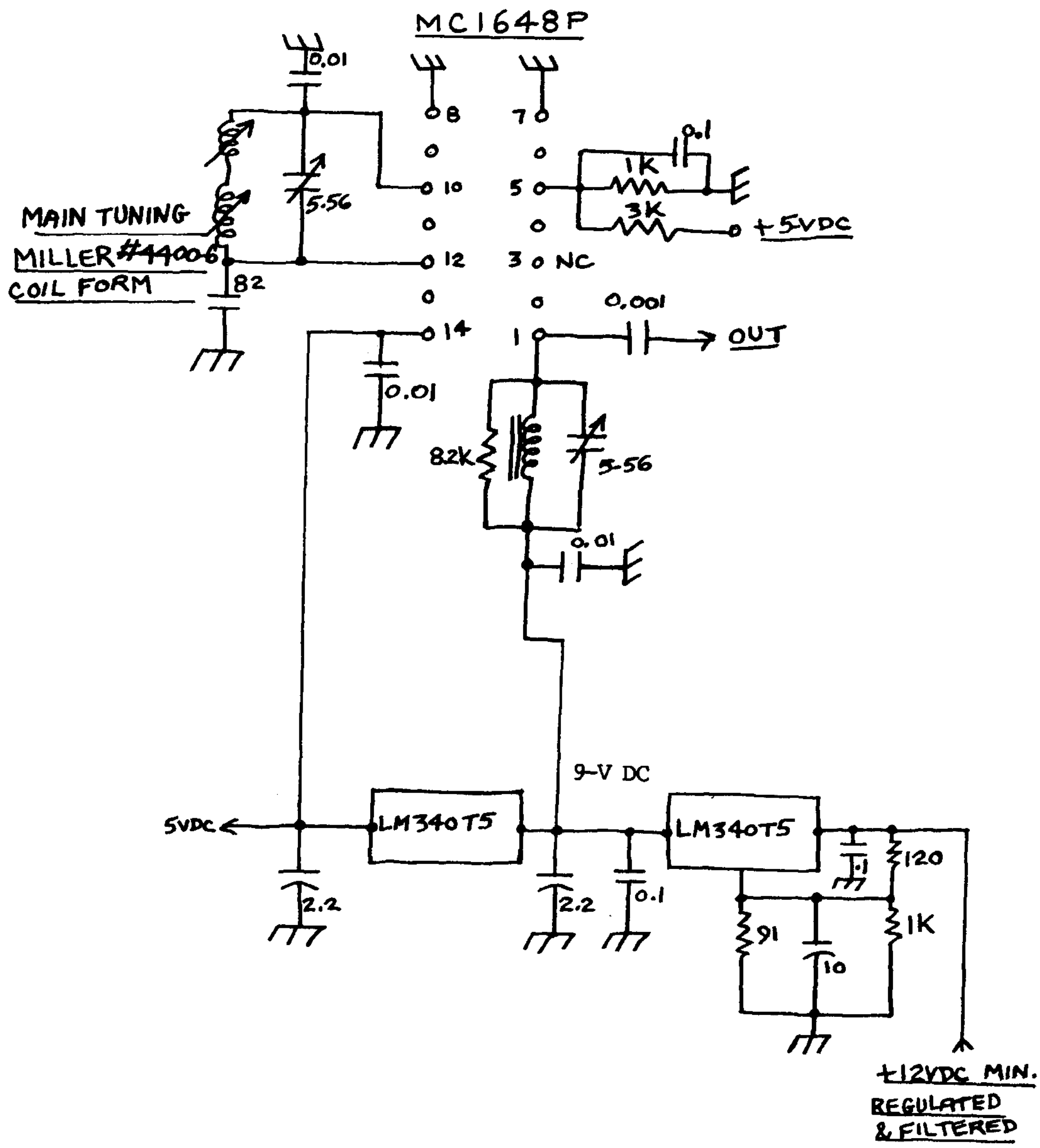
The trimmer capacitor in the output circuit should be adjusted for best sine wave over the tuning range. Amplitude will be quite uniform over 500 kHz, and should be adequate for driving doubly balanced active mixers such as the LM or MC1496.

Voltage regulation is extremely important to assure optimum performance from this circuit. Series regulation with high-grade regulators is strongly recommended.

The Miller no. 4400-6 coil form was prepared with 12 turns of no. 24 enameled copper wire with a winding length of 7/16 inch, which is held firmly in place with a coating of urethane varnish. The leading end of the core during 10 turns of adjustment must be contained entirely within this coil for best uniformity of tuning characteristic. The 0.250-inch diameter brass shaft is threaded internally to match the lead screw of the coil assembly and epoxied firmly in place. A formed metal rectangle supports the coil bushing on one face and the opposite face has a clearance 0.250-inch diameter bushing to accommodate the brass rod. The clearance bushing as well as the brass shaft was salvaged from an old potentiometer. In addition to the slide-rule string drive arrangement, I have added a 360-degree Waldom no. 303R knob which, in spite of the parallax inaccuracy caused by the in and out movement of the knob, provides some additional usable resolution beyond that of the slide-rule scale. For band-end operation of course, a frequency standard or counter should be consulted!

Over the years I have built a number of variable frequency oscillators with moderate success, but I believe the performance of this one comes about as close to that of a crystal oscillator as any I have constructed.

VFO 6.2-6.7 MHz



Convert BA Screws to American Sizes

By Ladimer S. Nagurney, WA3EEC
73 Blackberry Lane
Amherst, MA 01002

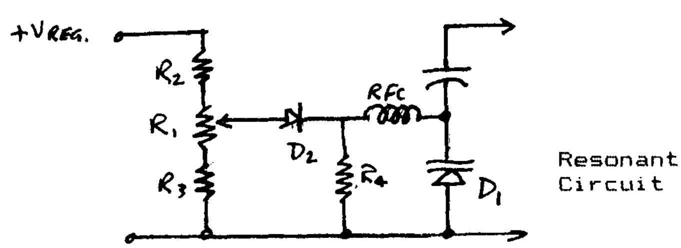
While designing several modifications to circuits in **The Microwave Newsletter Technical Collection** by the RSGB (available from the ARRL), I noted that many of the machine screw sizes were listed as British Association (BA) and followed by a number. I quickly noted that these numbers did not correspond to American sizes. While American National and, more recently, metric hardware is available, I did not have a source of BA hardware. Even various machinist handbooks did not list these or include them in the tap drill and clearance drill tables.

The RSGB Radio Data Reference Book (out of print) lists some sizes and tap and clearance drill sizes, while an old flier from Small Parts, Inc., provided the rest. Since most U. S. and Canadian amateurs would want to use available hardware, the table below would allow you to convert to available sizes. While I do not understand the history of these threads, it is interesting to note that some of the BA sizes are almost identical in diameter to American screws, although the pitch (and probably the shape) of the thread is different. In general, the thread is finer than standard American thread, but coarser than our NF thread. The tables below indicate the BA thread, its closest American equivalent, and the drill sizes. American sizes followed by asterisks indicate the American thread has the same diameter. I hope this is of use to hams converting British plans.

BA Thread Table

BA Size	Diameter (in.)	Threads per inch	Drill Tap	Drill Clearance	Closest American
BA-16			74	66	
BA-15			71	63	000-120 *
BA-14			70	58	000-120
BA-13			64	55	00-90
BA-12			62	54	
BA-11			56	1/16	0-80
BA-10	.067	73	55	50	0-80
BA-9	.075	65	53	46	1-64/72 *
BA-8	.087	59	50	42	2-56/64 *
BA-7	.098	53	46	36	3-48/56
BA-6	.110	48	42	32	4-40 *
BA-5	.126	43	38	30	5-40 *
BA-4	.142	39	32	25	6-40
BA-3	.161	35	29	18	8-32/36 *
BA-2	.185	31	24	9	10-24
BA-1	.209	28	17	2	12-24/28
BA-0	.236	25	9	C	

(continued from page 5)

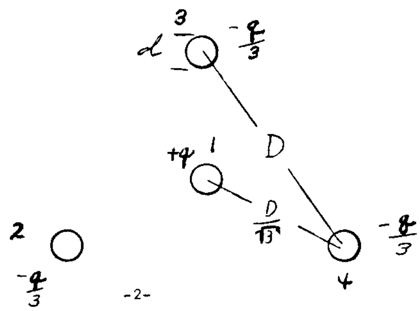


R4=10 to 25% of R1+R2+R3

Fig. 3 — Truncating range extremes to get further improvement in linearity.

(continued from page 2)

We proceed with the calculation of capacitance per meter of the four-wire line.



-2-

The outer conductors are separated a distance D, the wire diameter is d, and the distance between an outer conductor and the center conductor is $D/\sqrt{3}$. There is a charge of +q on the center conductor and a charge of −q/3 on each of the outer conductors. These values were chosen so that the net charge is zero, as observed at a great distance from the line. The voltage at a point x distance from the conductor is

$$V=\int_{d/2}^{x}\vec{E}\cdot d\vec{r}=\int_{d/2}^{x}\frac{q}{2\pi\epsilon}\cdot\frac{dr}{r}$$

What we have to do is go from one of the outer conductors, say 2, to conductor 1, taking into account the electric fields of all four conductors.

$$V_{2\to1}=\overset{\text{cond 2}}{\int_{d/2}^{D/\sqrt{3}}}\frac{q}{6\pi\epsilon}\frac{dr}{r}-\overset{\text{cond 1}}{\int_{D/\sqrt{3}}^{d/2}}\frac{q}{2\pi\epsilon}\frac{dr}{r}+\overset{\text{cond 3}}{\int_{D}^{D/\sqrt{3}}}\frac{q}{6\pi\epsilon}\frac{dr}{r}+\overset{\text{cond 4}}{\int_{D}^{D/\sqrt{3}}}\frac{q}{6\pi\epsilon}\frac{dr}{r}$$

$$=\frac{q}{6\pi\epsilon}\left(\ln\frac{2D}{\sqrt{3}d}+3\ln\frac{2D}{\sqrt{3}d}-\ln\sqrt{3}-\ln\sqrt{3}\right)$$

$$=\frac{q}{6\pi\epsilon}\left[\ln\left(\frac{2D}{\sqrt{3}d}\right)^4-\ln 3\right]=\frac{q}{6\pi\epsilon}\ln\frac{\left(\frac{2D}{\sqrt{3}d}\right)^4}{(3^{1/4})^4}$$

$$=\frac{q}{6\pi\epsilon}\ln\left(\frac{2D}{3^{1/4}\sqrt{3}d}\right)^4=\frac{2q}{3\pi\epsilon}\ln\frac{2D}{3^{1/4}d}.$$

$$V=\frac{2CV}{3\pi\epsilon}\ln\frac{2D}{3^{1/4}d},$$

$$C=\frac{3\pi\epsilon}{2\ln\frac{2D}{3^{1/4}d}}\quad\text{farad/meter}$$

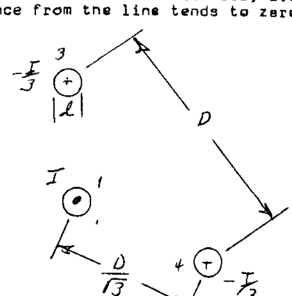
ϵ is the absolute dielectric constant of the medium surrounding the wires. For free space, and very nearly for air,

-3-

$$\epsilon=\frac{1}{36\pi}\times10^{-9}\quad\text{farad/meter},$$

$$C=\frac{4.17\times10^{-11}}{\ln\frac{2D}{3^{1/4}d}}=\frac{1.81\times10^{-11}}{\log_{10}\frac{2D}{3^{1/4}d}}\quad F/m.$$

The inductance per meter of this transmission line is the total perpendicular component of flux passing between the center conductor and an outer conductor. The current in each of the conductors is considered in the calculation. In the figure below, the line cross-section is shown again, this time with conductor currents having magnitudes such that the line is balanced, i.e., the magnetic field at a great distance from the line tends to zero.



The flux passing through, say, the plane containing wires 1 and 2 is:

$$\psi=\overset{\text{cond 1}}{\int_{d/2}^{D/\sqrt{3}}}\frac{\mu I}{2\pi}\frac{dr}{r}+\overset{\text{cond 2}}{\int_{d/2}^{D/\sqrt{3}}}\frac{\mu I}{6\pi}\frac{dr}{r}-\overset{\text{cond 3}}{\int_{D/\sqrt{3}}^{D}}\frac{\mu I}{6\pi}\frac{dr}{r}-\overset{\text{cond 4}}{\int_{D/\sqrt{3}}^{D}}\frac{\mu I}{6\pi}\frac{dr}{r}$$

$$=\frac{\mu I}{6\pi}\left(3\ln\frac{2D}{\sqrt{3}d}+\ln\frac{2D}{\sqrt{3}d}-\ln\sqrt{3}-\ln\sqrt{3}\right)$$

$$=\frac{\mu I}{6\pi}\left[\ln\left(\frac{2D}{\sqrt{3}d}\right)^4-\ln 3\right]$$

-4-

$$\psi=\frac{\mu I}{6\pi}\ln\frac{\left(\frac{2D}{\sqrt{3}d}\right)^4}{(3^{1/4})^4}=\frac{2\mu I}{3\pi}\ln\frac{2D}{3^{1/4}d}$$

$$L=\frac{2\mu}{3\pi}\ln\frac{2D}{3^{1/4}d}\quad\text{henry/meter}.$$

μ is the absolute magnetic permeability of free space, and the value $\mu=4\pi\times10^{-7}$ holds very nearly for air and most insulators.

$$L=2.67\times10^{-7}\ln\frac{2D}{3^{1/4}d}=6.14\times10^{-7}\log_{10}\frac{2D}{3^{1/4}d}\quad H/m$$

The characteristic impedance for an <u>air-insulated</u> four-wire line is:

$$Z=\sqrt{\frac{L}{C}}=184\log_{10}\frac{2D}{3^{1/4}d}$$

$$=184\log_{10}\frac{D}{1.14d}\quad☆$$

The phase propagation speed for this air-insulated line is

$$v=\frac{1}{\sqrt{\mu\epsilon}}=3\times10^{8}\quad m/s$$

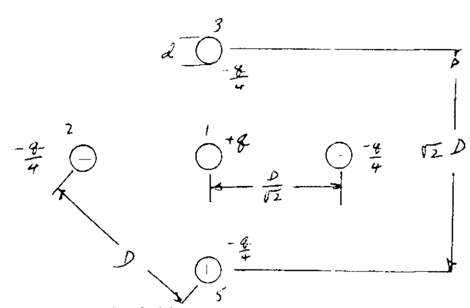
This latter result emphasizes the fact that, for free-space values of μ and ϵ for the medium surrounding the wires, the velocity factor is 100%. This works out even for coaxial cables. The fact that the velocity factor for many coaxial cables is 66% is attributable to the dielectric constant of the polyethylene between the conductors, and not the geometry of the conductors.

The velocity factor for the four-wire line having an insulating medium of relative dielectric constant ϵ_r is $V.F.=\frac{100\%}{\sqrt{\epsilon_r}}$

The approximate parameters of the five-wire line are derived below at a faster pace than before.

☆ In general, $Z=\frac{184}{\sqrt{\epsilon_r}}\log_{10}\frac{D}{1.14d}$.

-5-



Capacitance/meter calculation:

$$V_{2\to1}=\overset{\text{cond 1}}{\int_{D/\sqrt{2}}^{d/2}}\frac{-q}{2\pi\epsilon}\frac{dr}{r}+\overset{\text{cond 2}}{\int_{d/2}^{D/\sqrt{2}}}\frac{q}{8\pi\epsilon}\frac{dr}{r}+\overset{\text{cond 3}}{\int_{D}^{D/\sqrt{2}}}\frac{q}{8\pi\epsilon}\frac{dr}{r}+\overset{\text{cond 4}}{\int_{\sqrt{2}D}^{D/\sqrt{2}}}\frac{q}{8\pi\epsilon}\frac{dr}{r}+\overset{\text{cond 5}}{\int_{D}^{D/\sqrt{2}}}\frac{q}{8\pi\epsilon}\frac{dr}{r}$$

$$= \frac{q}{8\pi\epsilon}\ln\frac{2D}{\sqrt{2}d} + \frac{q}{8\pi\epsilon}\ln\frac{2D}{\sqrt{2}d} - \frac{q}{8\pi\epsilon}\ln\frac{\sqrt{2}D}{D}$$

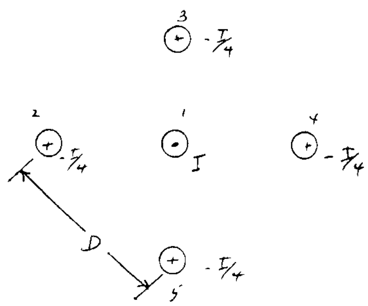
$$-\frac{q}{8\pi\epsilon}\ln\frac{2D}{D} - \frac{q}{8\pi\epsilon}\ln\frac{\sqrt{2}D}{D}$$

$$= \frac{q}{8\pi\epsilon}\left(5\ln\frac{\sqrt{2}D}{d} - \ln\sqrt{2} - \ln 2 - \ln\sqrt{2}\right)$$

$$= \frac{q}{8\pi\epsilon}\left[\ln\left(\frac{\sqrt{2}D}{d}\right)^5 - \ln 4\right]$$

$$= \frac{q}{8\pi\epsilon}\ln\left(\frac{\sqrt{2}D}{4^{1/5}d}\right)^5 = \frac{5q}{8\pi\epsilon}\ln\frac{D}{2^{-0.1}d}$$

Inductance per meter calculation:



$$\psi = \int_{d/2}^{\frac{D}{\sqrt{2}}\ \text{cond }1}\frac{\mu I\,dr}{2\pi r} + \int_{d/2}^{\frac{D}{\sqrt{2}}\ \text{cond }2}\frac{\mu I\,dr}{8\pi r} - \int_{\frac{D}{\sqrt{2}}}^{D\ \text{cond }3}\frac{\mu I\,dr}{8\pi r} - \int_{\frac{D}{\sqrt{2}}}^{\sqrt{2}D\ \text{cond }4}\frac{\mu I}{8\pi}\frac{dr}{r} - \int_{\frac{D}{\sqrt{2}}}^{D\ \text{cond }5}\frac{\mu I}{8\pi}\frac{dr}{r}$$

$$= \frac{\mu I}{8\pi}\left(4\ln\frac{\sqrt{2}D}{d} + \ln\frac{\sqrt{2}D}{d} - \ln\sqrt{2} - \ln 2 - \ln\sqrt{2}\right)$$

$$= \frac{\mu I}{8\pi}\left[5\ln\left(\frac{\sqrt{2}D}{d}\right) - \ln 4\right]$$

$$= \frac{\mu I}{8\pi}\ln\left(\frac{\sqrt{2}D}{4^{1/5}d}\right)^5 = \frac{\mu I}{8\pi}\ln\frac{D}{2^{-0.1}d}$$

$$= \frac{5\mu I}{8\pi}\ln\frac{D}{0.933d}$$

$$L = \frac{5\mu}{8\pi}\ln\frac{D}{0.933d} \quad H/m$$

In air,

$$L = 2.50\times10^{-7}\ln\frac{D}{0.933d} = 5.76\times10^{-7}\log_{10}\frac{D}{0.933d} \quad H/m$$

$$Z = \sqrt{\frac{L}{C}} = 173\log_{10}\frac{D}{0.933d} \quad \text{ohm}$$

$$v = \frac{1}{\sqrt{LC}} = 3\times10^{8}\ m/s$$

for $\mu = \mu_0$ and a relative dielectric constant ϵ_r

$$Z = \frac{173}{\sqrt{\epsilon_r}}\log_{10}\frac{D}{0.933d}$$

I'm happy to report that this answer for the five-wire line is confirmed on page 22-22 of the 5th edition of the ITT Reference Data for Radio Engineers, 1968, and on page 24-22 of the 6th edition (1975) of this handbook.

The expression for the phase velocity is:

$$v = \frac{1}{\sqrt{(5.76\times10^{-7})(1.93\times10^{-11})\epsilon_r}} = \frac{1}{3.33\times10^{-9}\sqrt{\epsilon_r}}$$

$$= \frac{3\times10^{7}}{\sqrt{\epsilon_r}} \quad m/sec$$

$$\text{velocity factor} = \frac{100\%}{\sqrt{\epsilon_r}}$$

Sincerely,

Pete Sypher

Peter Sypher, KC4SI

5902 Queenston Street, Apt. E
Springfield, Virginia 22152

April 18, 1985

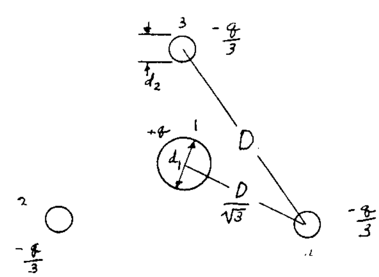
Jim Welch, KH6HEP
419A Atkinson Dr., #1001
Honolulu, Hawaii 96814

Dear Jim,

Thank you for your recent letter. It has spurred me to continue the analysis, to arrive at formulas for the 4- and 5-wire lines with a large-diameter center conductor. The exact expressions I alluded to at the end of my first letter are very difficult to derive, and the final results would be tedious to work with, pretty much like the formula for the 9-wire line you were kind enough to put in your letter to me. The formulas below are approximate, like those of the first letter. In fact, they are derived in the same way as the previous formulas were; only the limits of the integrals are changed to account for the different conductor diameters. The same assumptions listed at the beginning of Part 1 apply here, also.

PART 2

We proceed with the calculation of capacitance per meter of the four-wire line.



The same dimensions are used here as in Part 1, except that d_1 is the diameter of the center conductor, and d_2 is the diameter of each of the outer conductors.

$$V_{2\to1} = \int_{d_2/2}^{\frac{D}{\sqrt{3}}\ \text{cond. }2}\frac{q\,dr}{6\pi\epsilon r} - \int_{\frac{D}{\sqrt{3}}}^{\frac{d_1}{2}\ \text{cond }1}\frac{q\,dr}{2\pi\epsilon r} + \int_{D}^{\frac{D}{\sqrt{3}}\ \text{cond }3}\frac{q\,dr}{6\pi\epsilon r} + \int_{D}^{\frac{D}{\sqrt{3}}\ \text{cond. }4}\frac{q\,dr}{6\pi\epsilon r} \quad \text{volts}$$

$$= \frac{q}{6\pi\epsilon}\left(\ln\frac{2D}{\sqrt{3}d_2} + 3\ln\frac{2D}{\sqrt{3}d_1} - \ln\sqrt{3} - \ln\sqrt{3}\right)$$

$$= \frac{2q}{3\pi\epsilon}\ln\frac{2D}{3^{3/4}d_1^{3/4}d_2^{1/4}} \quad \text{volts}$$

$$= \frac{2CV}{3\pi\epsilon}\ln\frac{2D}{3^{3/4}d_1^{3/4}d_2^{1/4}}$$

$$C = \frac{3\pi\epsilon_o\epsilon_r}{\ln \frac{2D}{3^{3/4} d_1^{3/4} d_2^{1/4}}} \quad \text{farad/meter,}$$

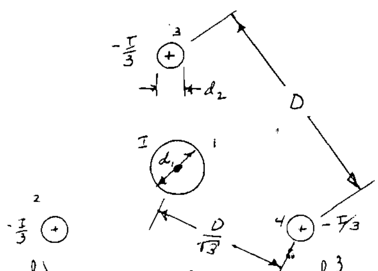
where $\epsilon_o = \frac{1}{36\pi} \times 10^{-9}$ farad/meter,

ϵ_r = relative dielectric constant, which is ≈ 1 for air.

$$C = \frac{4.16\times 10^{-11}\,\epsilon_r}{\ln\left(\frac{2D}{3^{3/4} d_1^{3/4} d_2^{1/4}}\right)} = \frac{1.81\times 10^{-11}\,\epsilon_r}{\log_{10}\left(\frac{2D}{3^{3/4} d_1^{3/4} d_2^{1/4}}\right)} \quad \text{F/m}$$

-3-

Inductance calculation for the 4-wire line:



$$\psi = \int_{d_1/2}^{D/\sqrt{3}} \frac{\mu I\,dr}{2\pi r} + \int_{d_2/2}^{D/\sqrt{3}} \frac{\mu I\,dr}{6\pi r} - \int_{D/\sqrt{3}}^{D} \frac{\mu I\,dr}{6\pi r} - \int_{D/\sqrt{3}}^{D} \frac{\mu I\,dr}{6\pi r}$$

(cond. 1, cond. 2, cond. 3, cond. 4) weber-turns/meter

$$= \frac{\mu I}{6\pi}\left(3 \ln \frac{2D}{\sqrt{3}\,d_1} + \ln \frac{2D}{\sqrt{3}\,d_2} - \ln\sqrt{3} - \ln\sqrt{3}\right)$$

$$= \frac{2\mu I}{3\pi} \ln \frac{2D}{3^{3/4} d_1^{3/4} d_2^{1/4}} \quad \text{weber-turns/meter}$$

$$L = \frac{2\mu}{3\pi} \ln \frac{2D}{3^{3/4} d_1^{3/4} d_2^{1/4}} \quad \text{henry/meter,}$$

where $\mu = 4\pi\times 10^{-7}$ henry/meter

$$L = 2.67\times 10^{-7} \ln \frac{2D}{3^{3/4} d_1^{3/4} d_2^{1/4}} = 6.14\times 10^{-7} \log_{10} \frac{2D}{3^{3/4} d_1^{3/4} d_2^{1/4}} \quad \text{H/m}$$

-4-

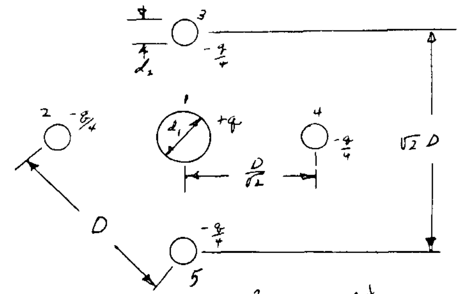
For the 4-wire line with a large-diameter center conductor:

$$Z = \sqrt{\frac{L}{C}} = \frac{184}{\sqrt{\epsilon_r}} \log_{10} \frac{2D}{3^{3/4} d_1^{3/4} d_2^{1/4}} \quad \text{ohm}$$

$$= \frac{184}{\sqrt{\epsilon_r}} \log_{10}\left(\frac{D}{1.14\, d_1^{0.75} d_2^{0.25}}\right) \quad \text{ohm}$$

$$\text{velocity factor} = V.F. = \frac{100\%}{\sqrt{\epsilon_r}}$$

We go on to the capacitance/meter of the 5-wire line:



$$V_{2\to1} = \int_{D/\sqrt{2}}^{d_1/2} \frac{-q\,dr}{2\pi\epsilon r} + \int_{d_2/2}^{D/\sqrt{2}} \frac{q\,dr}{8\pi\epsilon r} + \int_{D}^{D/\sqrt{2}} \frac{q\,dr}{8\pi\epsilon r} + \int_{\sqrt{2}D}^{D/\sqrt{2}} \frac{q\,dr}{8\pi\epsilon r} + \int_{D}^{D/\sqrt{2}} \frac{q\,dr}{8\pi\epsilon r}$$

(cond. 1, cond. 2, cond. 3, cond. 4, cond. 5)

-5-

$$V = \frac{q}{8\pi\epsilon}\left(4 \ln \frac{\sqrt{2}\,D}{d_1} + \ln \frac{\sqrt{2}\,D}{d_2} - \ln\sqrt{2} - \ln 2 - \ln\sqrt{2}\right)$$

$$= \frac{5q}{8\pi\epsilon} \ln \frac{D}{2^{-0.1} d_1^{4/5} d_2^{1/5}} \quad \text{volts}$$

$$= \frac{5CV}{8\pi\epsilon} \ln \frac{D}{2^{-0.1} d_1^{4/5} d_2^{1/5}} \quad \text{volts}$$

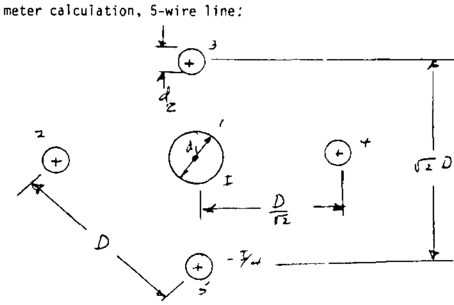
$$C = \frac{8\pi\epsilon}{5 \ln \frac{D}{2^{-0.1} d_1^{4/5} d_2^{1/5}}} \quad \text{farad/meter}$$

$$= \frac{4.44\times 10^{-11}\,\epsilon_r}{\ln \frac{D}{2^{-0.1} d_1^{4/5} d_2^{1/5}}} \quad \text{F/m}$$

$$= \frac{1.93\times 10^{-11}\,\epsilon_r}{\log_{10}\left(\frac{D}{0.933\, d_1^{0.8} d_2^{0.2}}\right)} \quad \text{F/m}$$

-6-

Inductance per meter calculation, 5-wire line:



$$\psi = \int_{d_1/2}^{D/\sqrt{2}} \frac{\mu I\,dr}{2\pi r} + \int_{d_2/2}^{D/\sqrt{2}} \frac{\mu I\,dr}{8\pi r} - \int_{D/\sqrt{2}}^{D} \frac{\mu I\,dr}{8\pi r} - \int_{D/\sqrt{2}}^{\sqrt{2}D} \frac{\mu I\,dr}{8\pi r} - \int_{D/\sqrt{2}}^{D} \frac{\mu I\,dr}{8\pi r}$$

(cond. 1, cond. 2, cond. 3, cond. 4, cond. 5)

$$= \frac{\mu I}{8\pi}\left(4 \ln \frac{\sqrt{2}D}{d_1} + \ln \frac{\sqrt{2}D}{d_2} - \ln\sqrt{2} - \ln 2 - \ln\sqrt{2}\right)$$

$$= \frac{5\mu I}{8\pi} \ln \frac{D}{2^{-0.1} d_1^{4/5} d_2^{1/5}} \qquad \text{weber-turn/meter}$$

$$L = \frac{5\mu}{8\pi} \ln \frac{D}{2^{-0.1} d_1^{4/5} d_2^{1/5}} \qquad \text{henry/meter}$$

$$= 5.76 \times 10^{-7} \log_{10} \frac{D}{0.933\, d_1^{0.8} d_2^{0.2}} \qquad \text{henry/meter}$$

$$\boxed{\begin{aligned} Z &= \sqrt{\frac{L}{C}} = \frac{173}{\sqrt{\varepsilon_r}} \log_{10} \frac{D}{0.933\, d_1^{0.8} d_2^{0.2}} \quad \text{ohms} \\ V.F. &= \frac{100\%}{\sqrt{\varepsilon_r}} \end{aligned}}$$

-7-

Well, Jim, there they are. The four formulas for line impedance. I included the mathematical song-and-dance here because you said you were intending to send in the results to QEX, and the folks at QEX might want to know how these results were obtained, in view of the fact that only a formula for the 5-wire-same-diameter case has been found in a respected reference (so far as I know).

I might note that these formulas in this letter apply to a transmission line having a center conductor smaller, as well as larger, than the outer conductors. In fact, the formulas of Part 1 are just special cases of the Part 2 formulas. If in the Part 2 formulas, you set $d_1 = d_2 = d$, you will find that these reduce to the Part 1 formulas.

Remember, these formulas are approximate. I have come across the exact formula for the 2-wire transmission line (in the ITT Handbook) having unequal diameters, and from playing with this formula, I have concluded when $\frac{D}{d_1}$ or $\frac{D}{d_2} \doteq 2$, the impedance error is about 20%. For wider spacings, that is, when $\frac{D}{d_1}$ and $\frac{D}{d_2} > 2$, this error is less than 20%, and when the spacing is closer, the error increases rapidly.

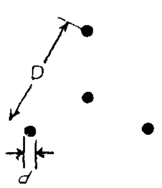
I hope these formulas are helpful to you. Don't hesitate to write if you want to discuss this matter further.

Very 73,

Pete Sypher

Pete Sypher, KC4SI

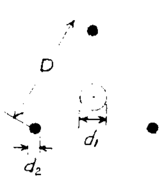
$D >> d$ ALL LOG = LOG_{10}



$$Z_o = \frac{184}{\sqrt{E}} \text{LOG} \frac{D}{1.14d} \qquad (1)$$

Fig 1

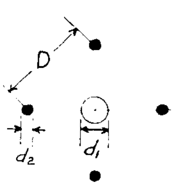
$D >> d$



$$Z_o = \frac{184}{\sqrt{E}} \text{LOG}\left(\frac{D}{1.14 d_1^{0.75}\ d_2^{0.25}}\right) \qquad (1)$$

Fig 2

$D >> d$



$$Z_o = \frac{173}{\sqrt{E}} \text{LOG}\left(\frac{D}{0.933 d_1^{0.8}\ d_2^{0.2}}\right) \qquad (1)$$

Fig 3

$h >> a$

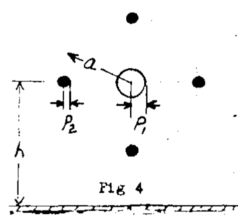


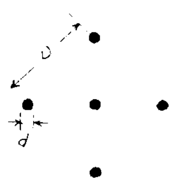
Fig 4

$$Z_o = 138\left(\text{LOG}\frac{2h}{P_1} + K\ \text{LOG}\frac{2h}{a}\right) \qquad (2)$$

$$K = -\left(\frac{4\ \text{LOG}\frac{2h}{a}}{\text{LOG}\frac{2h}{P_2} + \text{LOG}\frac{2h^2}{a^2} + \text{LOG}\frac{h}{a}}\right)$$

$$h = \infty \quad Z_o = 138\left(\text{LOG}\frac{a}{P_1} + \frac{1}{4}\text{LOG}\frac{a}{P_2} - \frac{1}{4}\text{LOG}\,4\right) \qquad (3)$$

$D >> d$

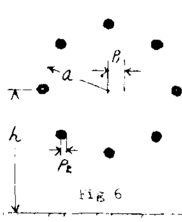


$$Z_o = \frac{173}{\sqrt{E}} \text{LOG}\frac{D}{0.933d} \qquad (4), (6)$$

$$Z_o = \frac{173}{\sqrt{E}} \text{LOG}\ 1.14\frac{D}{d} \qquad (5)$$

Fig 5

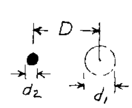
$h >> a$



$$Z_o = 138\left(\text{LOG}\frac{2h}{P_1} + K\ \text{LOG}\frac{2h}{a}\right) \qquad K = -\left(\frac{8\ \text{LOG}\frac{2h}{a}}{\text{LOG}\frac{2h}{P_2} + 2\text{LOG}\frac{2h}{0.77a} + 2\text{LOG}\frac{2h}{a\sqrt{2}} + 2\text{LOG}\frac{2h}{1.36a} + \text{LOG}\frac{h}{a}}\right)$$

WHEN $h >>> a$ $K \rightarrow 1.0$ & $Z_o \rightarrow 138\ \text{LOG}\frac{a}{P}$ (2)

Fig 6



$$Z_o = \frac{60}{\sqrt{E}} \text{COSH}^{-1} N \qquad (4)$$

$$N = \frac{1}{2}\left[\left(\frac{4D^2}{d_1 d_2}\right) - \left(\frac{d_1}{d_2}\right) - \left(\frac{d_2}{d_1}\right)\right]$$

$$\text{COSH}^{-1} X = \text{LN}\left(X + \sqrt{X^2 - 1}\right) \qquad (7)$$

Fig 7

$\text{LN} = \text{LOG}_e$

(1) Equation derived by Peter Sypher, KC4SI
(2) Radio Antenna Engineering - E. A. Laport
(3) Equation submitted by Albert E. Weller Jr. WD8KBW
(4) Reference Data for Radio Engineers - ITT
(5) Antenna Engineering Handbook - Henry Jasik
(6) Also derived by Peter Sypher, KC4SI
(7) Mathematical Tables from Handbook of Chemistry & Physics, 7th Edition

Bits

OKI Expands Line of CMOS Microcomputer Chips With Advanced Version of 80C51 8-Bit MCU

An advanced version of the standard 80C51 single-chip 8-bit microcomputer for multitasking control applications in automotive, instrumentation and telecommunications systems has been developed by Oki Semiconductor earlier this year. The device is fully compatible with a wide range of HMOS devices from Intel Corp., NEC and a host of other manufacturers, and with larger CMOS MCUs currently under development by Oki.

To ease the task of software development, the 80C51 offers the use of 111 instructions, including hardware multiply and divide. With its 131,456-bit memory space, it can execute the same instruction set used with the HMOS 8051.

The device also has the same on-chip 4k x 8-bit ROM and 128 x 8-bit RAM configurations as the slower 80C50. Contiguous addressing of ROM is unique to the 80C51, eliminating the page mode addressing that is necessary with other MCUs. It can also address 64 kbits of data and the same amount of programmable space. The 80C51 is architecturally configured for five-source two-level interrupts.

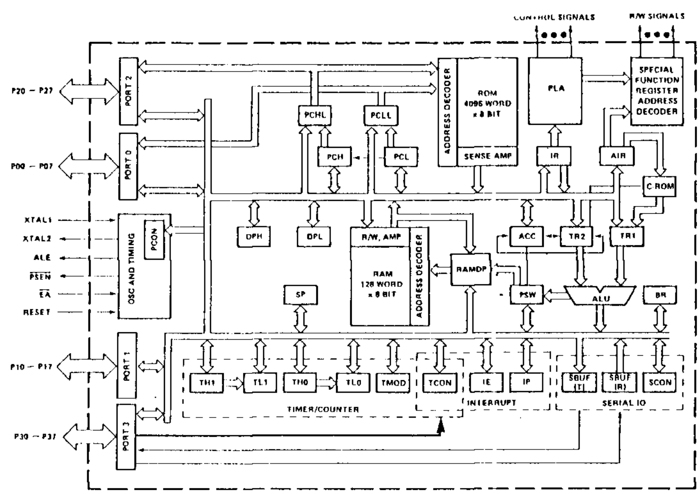
Operating frequency is 12 MHz over the full temperature range of -40 to +85 degrees Celsius. At this frequency, the length of an instruction cycle is 1 microsecond. Two full 16-bit timers are provided. Oscillator and clock circuitry are located on the chip.

There are 32 I/O lines in the 80C51, organized into four 8-bit ports. An I/O port expander, the 82C43, is available for use with the 80C51. The expander adds another 16 I/O lines, providing a total of 52.

The device contains a full-duplex serial channel that can be operated under four different modes. The chip also contains an on-board Boolean processor, capable of performing multiplication and division of bits and bytes. Operating modes can be selected by software and enables the user to lower power consumption to a few microwatts, while the contents of the RAM remain safe.

The Oki 80C51 is manufactured with the company's proprietary 2-micron, dual-layer-metal, P-well CMOS technology. The device operates over a frequency range of 1.2 MHz to 12 MHz and a Vcc range of 4.0V to 6.0V.

The 80C51 was first available in May 1985. The price per device is $14 in production quantities. More information can be obtained on the 80C51 by writing Irwin Schwartz or Anil Bedi, Oki Semiconductor, 650 North Mary Ave., Sunnyvale, CA 94086, tel. (408) 720-1900. Oki Semiconductor is a leader in memory, speech synthesis and CMOS microprocessor products. They manufacture and market very-large-scale integrated circuits for use in computers, communications systems, and consumer products.



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