No originality is claimed for these amazing revelations, but it is surprising how infrequently they are published:

Impedance, Z, is related to L, C and Q in the following expressions:

$$Z = Qu.Xl = Qu.2.\pi.f.l \qquad ...1$$

where f is the tuned frequency in Hz and l is the inductance in Henrys

and $$Z = Qu.XC = \frac{Qu}{2\pi.f.C} \qquad ...2$$

C is expressed in Farads.

Q is largely dependent on the core and bobbin materials - together with the DC winding resistance:

$$Q = \frac{'R'}{\omega L} \qquad ...3$$

in fact, the analysis is a great deal more detailed, but the formulae given here will be quite sufficient for most of the practical situations that confront the circuit user.

Working an example from the TOKO IF range, take the LMC4202A 7mm 455IF:

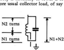
$Q = 105$ total turns 208
$C = 150.10^{-12}$

so, from (2) above

$$Z = \frac{105}{2.\pi.455.150.10^{-9}} = 244k\Omega$$

(all calculations used here will be rounded off for ease)

Fine, but most transistor and ICs need to work into much lower impedances than that, so a little transformer theory is necessary to lower Z to the more usual collector load, of say 37k.



Autotransformer tapping

The total tap point impedances are related by

$$\frac{Ztap}{Ztot} = \left[\frac{N1}{N1+N2}\right]^2$$

So, using our 37k value for Ztap:

$$\frac{37}{244} = \left[\frac{N1}{208}\right]^2$$

$$N1^2 = \frac{208^2 . 37}{244}$$

so $N1 = 80$

in fact, TOKO use 74 turns, but that isn't going to make a great deal of difference in practise.

The base coupling also requires a lowered impedance, and since it is not really desireable to employ another autotransformer type of tapping, a coupling secondary is used, where the primary and secondary are so tightly couled, that the basic analysis may be considered identical to that used in the autotransformer case.
In the case of the coil used here, the coupling is used for the detector - so a reasonably high value of 12k is used;

$$\frac{12}{244} = \left[\frac{N1}{208}\right]^2$$

$$N1^2 = \frac{208^2 . 12}{244}$$

thus N1 = 46 turns - and in this instance, the actual value used is 42 t.

The slight differences that occur are due to certain intentional mismatch conditions, designed to reduce the loaded Q. These considerations will be dealt with in greater detail in the next issue of Tecknowledgey.

In a perfectly matched system, the gain of an amplifier stage can be calculated from the following:

$$Av = \frac{Vout}{Vin} = K.gm.Rload.N.K$$

where N is the turns ratio $\sqrt{\frac{Zsec}{Zpri}}$

K represents the 6dB match loss at both input and output
gm is the transconductance in mho (reciprocal ohm)
Rload is the load impedance

In the example used so far, the gain of the final IF before the detector would be:

$$= ½.gm.37k.\sqrt{\frac{12}{244}} . ½$$

with a gm of 90mmho:

$$= 184 \text{ or } 45.3dB$$

The impedance may also be altered by a damping resistor; a popular means of lowering the Q and thereby stabilizing IFs that "take off" due to too much gain. Say to reduce a Q of 105 to Q of 40.

then $$Z = \frac{40}{\omega.\ 150.\ 10^{-12}}$$

$$= 93k\Omega$$

so to reduce our example, a parallel resistor is used, where

$$\frac{1}{93k} = \frac{1}{244k} + \frac{1}{R\ damp.}$$

$$R\ damp = 150k\Omega$$

Loading considerations

The loading of a tuned circuit of the type described here is first calculated by changing the tap impedance to a function known as the equivalent loading resistance, derived from the equivalent circuit:

The equivalent circuit:

Now $Ql = Rp \parallel Rl . Xc$

or $$= \frac{Rp \parallel Rl}{Xl}$$

as you might expect, Rl is taken from a similar transformation formula to the ones used so far:

$$Rl = \left[\frac{N1 + N2}{N1}\right]^2 . \text{ Input impedance}$$

with the example before

$$= \left[\frac{208}{74}\right]^2 . 37k$$

$$Rl = 294k$$

so $Ql = 244 \parallel 294 .Xc$

$244 \parallel 294 = 133.3k\Omega$

$$Ql = \frac{133,300}{2332} = 60$$

but dont forget the loading by the secondary winding !

$$Rl = \left[\frac{208}{42}\right]^2 . 12k = 294k -$$

now this is the same as Rl for the autotransformer tap, indicating that the optimum Rp would be 294k also.

Now $Ql = 244 \parallel 294 \parallel 294.Xc$

$$Ql = \frac{91,734}{2332} = 39.3 \text{ - which}$$

isn't far from the stated figure of 40.

So, to summarize the results:

$$\mathbf{Z = Rp = Qu.Xl = Qu.Xc}$$

$$\frac{Ztap}{Ztot} = \left[\frac{Tap\ turns}{Total\ turns}\right]^2$$

$$\textbf{Gain} = ½.gm.Rload.\ \overline{\frac{Tap\ turns}{Primary\ turns}} .½$$

The two "½" multipliers refer to the 6dB loss associated with input/out termination. It will be seen from the loaded Q formulae, that under best matching conditions, when all equiv. load impedances are the same, the Q is reduced to a Ql of (½.½) Qu. Since Q is directly related to V in a tuned circuit, this also implies the voltage is reduced by the (½.½) factor.

$$\mathbf{Ql = Rp \parallel Rl . Xc = Rp \parallel Rl . Xl}$$

and $$\mathbf{Rl} = \left[\frac{Total\ turns}{Tap\ turns}\right]^2 . \textbf{ Input imp}$$

Z = R in this feature - not to be confused with basic DC resistances.

Bandspread calculations for RF tuned circuits may look fearsome - but as long as a reasonable scientific calculator is used, the answers involve little difficulty.

The basic reason for bandspread in HF and communications recievers is simple: consider a general coverage application on SW3 (14MHz to 30MHz), and now think about the degree of electromechanical stability that is demanded of such a system when trying to resolve an SSB signal, where the carrier needs to be reinserted to within 50Hz. 50Hz on the basis of a coverage of 16MHz represents about one part in three million. This is not an easy task to acheive in terms of mechanical stability and tuning resolution on a dial where they may be only five to ten turns coverage. So the answer is to use a fine tuning capacitor connected in parallel across the main tuning capacitor, but having a greatly reduced capacitance-say one twentieth of the value of the main tuning gang. This approach is fine in many applications, but does not really solve the problem where long term electro-mechanical stability is essential. It merely facilitates the vernier tuning by the operator.

So the next technique is the expansion of the band to absorb the whole range of the main tuning gang - say 360pF - over a relatively small RF space, as in the type of reciever that covers 'amateur' bands, or broadcast bands only. In this way 21 to 21.5MHz is made to take up all 366pF of the tuning gang swing instead of just a pF or two. This means that small changes in the tuning capacitor due to mechanical shock, heat etc., are greatly buffered in terms of the final frequency shift.

Examples:

The tuned frequency of L/C parallel circuit is given by

$$f = \sqrt{\frac{25330.3}{L.C}}$$

Where f is in MHz
L is in microhenrys
and C is in pF

(Derived from $f = \frac{1}{2\pi\sqrt{LC}}$)

So, in the gnereal coverage application, to reach 30MHz with a minimum tuning capacity of 30pF - to allow for strays, trimmers etc - the inductance required is only 0.9uH (approximations will be used to avoid unecessary decimal complications.)

so at 21.5MHz, a capacitor of 61pF is required, and at 21MHz, a value of 64pF in other words, a change of only 3pF covers 500kHz at 21MHz. It isn't difficult to see that the mechanical susceptibility of such a system is very poor.

So in the process of spreading the band, the endeavour is to make all 366pF do the work of 3pF, and thus make all minor changes in C insignificantly small.

The basic considerations in bandspread calculations

In a tuned circuit arrangement that employs a variable capacitor for tuning (as nearly all outside car radios do), the frequency range covered is determined by the ratio of the maximum and minimum (including strays) capacity that appears across the inductance of the tuned circuit.

The required capacitance ratio, R,

$$(d^2) = \left[\frac{\text{Max frequency}}{\text{Min frequency}}\right]^2 \qquad (A)$$

let V = capacitance ratio of the tuning capacitor
Cv= maximum value of tuning capacitance
Cp= total parallel capacitance across tuning cap.
Cs= capacitance used in series with tuning cap.
BW= tuning range

$$\text{and BW} = [d-1]\sqrt{\frac{fmax.fmin}{d}} \qquad (B)$$

Parallel capacitors

There are two basic approaches to the techniques of electrical bandspread - for a variety of reasons, the usual result is a combination of the two, since the impedance of the tuned circuit is very low with a high value of parallel capacity - and thus not suited to many oscillator applications - or very high with a large value of inductor, where the stray capacities inherent in PCBs and wiring limit the overall tuning range through a tight restriction on the factor 'V' (Capacitor ratio)

$$Cp = \frac{Cv\ (V-R)}{V(R-1)} \qquad (C)$$

$$D = \frac{V(Cv+Cp)}{VCp + Cv} \qquad (D)$$

Now at the lowest frequency, the total tuning capacity is Ct = Cv+Cp

As an example, take an interpolation oscillator for tuneable IF of 10.6MHz to 10.8MHz

$$R = (1.02)^2 = 1.04$$

Using a BB104 varicap over a range of 2 to 10v bias C min is 12.5pF, and Cmax is 22.5pF so

$$V = \frac{22.5}{12.5} = 1.8$$

$$Cv = 22.5pF$$

substituting in (C)

$$Cp = \frac{22.5(\ 1.8-1.04)}{1.8(1.04-1)} = 237.5pF$$

So, in order to leave room for strays, use 220pF fixed 5% with a 2-22pF trimmer.

The value of the inductor is then derived from the basic formula for the resonant frequency, where f= 10.8 and C= 237.5+12.5 = 250pF

so L = 0.868uH,

but this leaves an impedance of Q.Xc
assume a Q of 100 and then Z = 100 x 5.89
which is only 589 ohms, and not generally much use in this context.

Before moving on, the tuning bandwidth may be confirmed from equation (B)

BW = 0.211MHz (since approx. are used)

So, the series capacitor method comes next:

This approach relies on the principal that a small capacitor placed in series with the Cv factor will reduce the effective parallel capacity across the tuned circuit to value that is

$$\frac{1}{\frac{1}{Cs} + \frac{1}{Cv}}$$

Using the various factors already discussed

$$Cs = \frac{Cv(R-1)}{V - R} \qquad (E)$$

(series C bandspread....)

the capacitance ratio, R

$$= \frac{VCs + Cv}{Cs + Cv} \qquad (F)$$

The total effective C accross the tuned circuit is also derived from

$$Cc = \frac{Cv(D-1)}{V-1} \qquad (G)$$

The example used will be based on the same problem so Cs

$$= \frac{22.5(1.04-1)}{1.8-1.04}$$

$$= 1.18pF$$

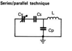
leading to L

$$= 201uH \text{ , and with a q of 100}$$

$$Z = 1.34 \text{ Mohm}$$

which is just about as unlikely as the result for basic parallel capacitor. The stability demands on the series capacitor are quite impossible to achieve - and no account of stray capacitance has been been made. So, to strike a useful medium, it is not surprizing to find that a combination of the two methods is used.

The Series/parallel technique



The circuit may be analyzed from a combination of the preceding formulae ((A) to (G)), using a mid-band value for the circuit impedance that is going to result in practical values and tolerances, or the following additional equations, which reduces the task to one of programming your calculator, and thinking of a few numbers:

Cta = Total maximum capacitance
Cps = Parallel capacitance
Ccs = Series capacitor
A = Intermediate capacitance ratio of the series arm of the network

and A is between the values of V and R

$$A = \frac{VCs + Cv}{Cs + Cv} \qquad (H)$$

$$Cpa = \frac{Cv(A-1)(A-R)}{A(R-1)(V-1)} \qquad (J)$$

$$Cta = \frac{RCv(A-1)^2}{A(R-1)(V-1)} \qquad (K)$$

$$\text{or } Cta = Cpa \cdot \frac{RCv(A-1)^2}{Cv(A-1)(A-R)} \qquad (L)$$

The value of A is found by introducing a few more variables:

$$Cj = 2RCv \quad \text{then}$$

$$Ck = Cj + Cta(R-1)(V-1)$$

$$\text{and } A = \frac{Ck + \sqrt{Ck^2 - Cj^2}}{Cj} \qquad (M)$$

which leads to

$$Cs = \frac{Cv(A-1)}{V-A} \qquad (N)$$

Now this technique is by far the most widely used in design and tracking of resonant circuits - and in the example used so far, where

$$R = (1.02)^2 = 1.04$$

$$V = 1.8$$

$$Cv = 22.5pF$$

Cta = chose a value that makes some practical sense here, say for Z of 10k ohms at 10.6MHz
= 150pF, which is quite manageable sort of choice, as the L is from

$$L = \frac{25530.3}{150(10.6)^2} = 1.5uH$$

the approx. value for A from (H)

$$Cj = 2 \,.\, 1.04 \,.\, 22.5 = 46.8$$

$$Ck = 46.8 + (150[1.04-1][1.8-1]) = 51.6$$

$$A = 1.5669$$

$$\text{so } Cs = \frac{22.5(1.5669-1)}{1.8-1.5669} = 54.72pF$$

The series capacitor is selected to be 56 pF in this instance bearing in mind it is going to be a great deal more satisfactory to place any trimming C in parallel, where one side will be RF earth to permit adjustments without stray errors.

Plugging this back into (H)

$$A = \frac{(1.8 \,.\, 56) + 22.5}{56 + 22.5} = 1.57$$

$$Cpa = \frac{22.5(1.57-1)(1.57-1.04)}{1.57(1.04-1)(1.8-1)} = 135.3pF$$

$$(A-1)^2 = 0.325$$

$$\text{so } Cta = \frac{1.04(22.5)(0.325)}{1.57(1.04-1)(1.8-1)}$$

$$= 151.37pF$$

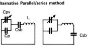
which confirms the original conditions

To allow for a trimmer, the final choice should be:

Cs = 56pF
Cp = 100pF
Ctrim= 0-60pF (in parallel with Cp)

remember that the distributed capacitance of the inductor will account for a few pF in Cp section of the equations- but this is not appreciable until layer wound LF coils are employed.

Alternative Parallel/series method



Csb = series capacitance
Ctb = total maximum capacitance
Cpv = max C in parallel arm
B = intermediate cap. ratio in parallel arm only (again, between V and R)

$$B = \frac{V(Cv + Cp)}{Cv + VCp} \qquad (O)$$

$$Csb = \frac{BCv(V-1)(R-1)}{V(B-1)(B-R)} \qquad (P)$$

$$Ctb = \frac{BCv(V-1)(R-1)}{V(B-1)^2} \qquad (Q)$$

As Csb will be fixed, Cp is solved using intermediate variables:

$$Cq = 4V \cdot Cv \cdot \left[\frac{Csb(V-R)}{R-1} - Cv\right] \quad \text{(R)}$$

$$Cr = Cv(V+1) + VCsb \quad \text{(S)}$$

$$Cp = \frac{\sqrt{Cq + Cr^2} - Cr}{2V} \quad \text{(T)}$$

(It should be noted that all these various formulae are basically algebraic manipulations of the basic LC resonance equation— and so derivations are not given here for reasons of space)

For a change, the example used here will relate to something different - coverage of the MW with the KV1210 varactor triplet. Reference to the data sheet of the KV1210 shows that from 2 to 9v bias, the capacity swing is from 400pF to 30pF typically (per diode)

$$V = 400/30 = 13.33$$
$$Cv = 400pF$$
$$BW = 1605/525 \text{ kHz}$$
$$R = BW^2 = 9.35$$

So take Csb = 10000pF and insert in the formulae:

$$Cq = 4 \cdot 13 \cdot 400 \left[\frac{10000\ (3.65)}{9.35-1} - 400\right] = 8.26 \times 10^7$$

$$Cr = 400(13+1) + 13(10000) = 135600$$

$$\text{so } Cq + Cr^2 = 1.847 \times 10^{10}$$

$$\text{and } Cp = \frac{1.36 \times 10^5 - 135600}{2 \times 13} = 17pF$$

as far as the RF sections of the MW are concerned, then Cp is simply a 7/35pF trimmer, set halfway and trimmed to take up strays.

The inductance at a Ct of 417pF and a frequency of 525kHz is then

$$= \frac{25330.3}{417 \cdot (0.525)^2} = 220uH$$

which should also occur at 1605kHz and 47pF

$$= \frac{25330.3}{47 \cdot (1.605)^2} = 217uH$$

(The slight error is due to use of 400pF without taking into account the effect of the 10000pF in series)

The final part of this series of bandspread and tracking details will appear in the next issue - it concerns the tracking of the local oscillator at (signal frequency + IF) and covers both parallel gangs (where both the antenna(e) and oscillator sections of the tuning capacitors are the same) and non- parallel gangs eg 160+80 pF, as often found in imported MW only radio applications

In the next issue of Tecknowledgey:

At the time of writing this, all consideration of the next issue seems the height of folly - after all, this issue has been delayed for a string or reasons, but primarily because we have been busier for the time of the year than we anticipated, and we want the whole production to be carried by Ambit staff who are aquianted with the products - and generally in the "thick of things". We have not managed to get in all we wanted to - so additional supplements are going to be produced on things such as

* The Radiometer Ferret
* The PCB/dalo pens
* Databooks
* Hardware including tuning pots etc
* Spaghetti farming in Tuscany

But in the next issue proper.........

All that's new and wonderful in radio ICs, we hope to include details of the first Europeanized frequency synthesiser for FM and AM, some improved DFM chips (and ways to remove their RFI problems).
A full report on the new family of noise blankers, introduced for the first time in this issue with the KB4423.
Words and details of UHF tunerhead techniques, with a TV preamp (tuneable) with printed inductors.
But since we aim to keep as up-to-the minute as possible, it is difficult to predict much more than that.

We would appreciate any contributions from our readers/customers. These will be assessed quickly, and payment will be made on acceptance. The features must naturally relate to modern techniques, and modern components with a main theme of radio - or based on products appearing in our catalogue/price list. We particularly invite articles on the following:
Switching radios with the MA1012/MA1023
Use of the TDA1083 in communication type circuits.
The KB4423 as a record scratch blanker/ IF noise blanker at IF frequencies.
A perpetual motion device that works.

PS:
We know about the wonder decoder announced in April WW - the TCA4500A - and have given it a check At present, we think its fine for car radio, but overall, the HA1196 is superior in all normal tuner applications. In fact, the HA1196 can be fitted with variable blend with a simple FET voltage dependant resistor in the separation circuit.
A new IC from Hitachi with .02% stereo THD and 86dB S/N is likely to be our next decoder feature, when available. But all this is bit acedemic when you consider broadcast standards at the transmitter don't promise better than some 65dB S/N and we have heard of one BBC station broadcasting 3% THD, which was later reduced when they found out !
Trends seem to be towards Dolby broadcasting, and it can already be usefully employed in conjunction with certain IBA stations who we think are already testing. The BBC variable pre-emphasis technique also benefits from a Dolby approach to de-emphasis.

This issue is something of new concept of combining sales promotion with feature articles (original ones, that is) and so we would like to know what you think of it. The editorial on page 3 sets out our basic approach, namely this is a magazine for engineers, by engineers who can still remember which end of the soldering iron gets hot. Do you want more basic radio reference theory, cut out from transmission line principles that most humble wireless followers don't ever really use? Is it too boring? Has it been of any use? Would you buy another?
By using the term "engineers" we try to encompass both the professional, and of course the enthusiast, and since these are frequently the same creature - we don't really think a distinction in terms is necessary. There is certainly a distinction in terms of basic knowledge and experience, but since we know many hobby enthusiasts who know a whole lot more than many professionals, the choice of terms is difficult if one or other of our two main readerships is not to feel insulted. You're all really very wonderful.