FERRITE IMPEDANCE TRANSFORMERS FOR ANTENNA SYSTEMS

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Generalities

I'm writing this short article to clarify some useful formulas for the calculation of ferrite impedance transformers used in antenna systems. This passage has a merely practical application (the core of this exposé is the example of its practical realization at the end of it) and it simply aims at presenting the problem without necessarily being a strict academic discussion. It goes without saying that a bibliography can easily be found at at the bottom of the passage for whoever would want to dive deeper into the topic.

Theory

Let's take into consideration the power supply system of an antenna such as the one shown in figure 301; where the antenna is just a resistance it can be represented by its equivalent Thevenin [1], with the impedance Z_L=R_L, which is supplied by a transmission line having the same characteristic impedance of the antenna Z_0 being Z_0=R_L. In this condition we are able to obtain the maximum power possible from the line to the antenna because we do not experience any wave reflection on the line itself (let's assume that the power supply is impedance matched for the line). Yet, often times obtaining the perfect match between the line and the load cannot be possible because commercial lines have the characteristic impedance Z_0 which is not equal to R_L. In this latter situation, in order to get a broadband match from line to load, an impedance transformer has to be placed at the height of the antenna connector. The transformer shows the load Z_b to the line, the real part R_b of which is equal to the characteristic impedance of the line (figure 302).

Occasionally, an autotransformer is used in place of a transformer (which has two windings galvanically separated) to reduce the parasitic elements, even though it only links the secondary winding to part of the primary one; therefore, the balancing effect between the input and the output resulting from the galvanic separation is annulled. That's the price to pay.

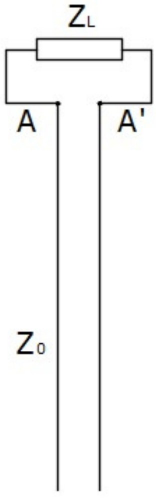

Fig.301

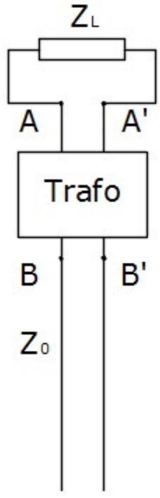

Fig.302

The transformer (as well as the autotransformer), not only does it transfer energy, but it also converts the load impedance Z_L to be as close in value to the characteristic impedance of the line z_0. The link between the two impedances (in this case only resistances because we assumed that the antenna is a pure resistance) is standard for a transformer.

$$Z_b = n^2 Z_L \qquad \text{(n is the turns ratio between } n_1 \text{ and } n_2\text{)} \qquad 3.0$$

Referring to figure 303, the load impedance Z_L, consisting only in its real part R_L it is seen at the connecting terminals B-B', as equivalent to the real part R_b with the magnetizing reactance of the primary winding L_m in parallel, followed by an ideal transformer (figure 304).

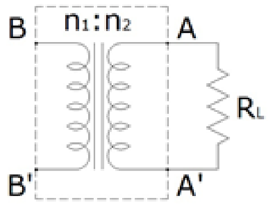

Fig.303

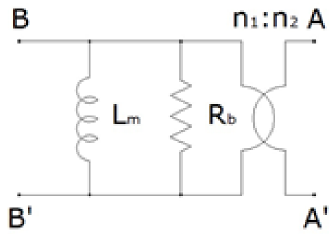

Fig.304

Therefore, the equation that links the two real parts R_L and R_b deduced from 3.0 is:

$$R_b = n^2 R_L \qquad 3.1$$

The magnetizing inductance L_m is deduced from the self-inductance of the primary winding L_p which is linked, in turn, to the building criteria of the transformer itself.

When the transformer is completed if we measure the inductance of the primary with the opened secondary, we obtain the L_{PO} value, whereas with the connecting terminals of the secondary in power surge we obtain the L_{PC} inductance. From these two inductance values we obtain the coupling factor *k.* If the transformer is well built the coupling factor turns out being around 1, so we can consider, with good approximation, the self-inductance of the primary winding to be equal to the magnetizing inductance L_m of the transformer.

$$k = \sqrt{1 - \frac{L_{PC}}{L_{P0}}} \approx 1 \qquad 3.2$$

Once we have the coupling factor *k* between primary and secondary winding, we can obtain the magnetizing inductance L_m from the following formula:

$$L_m = k \cdot L_{P0} \approx L_{P0} \qquad 3.3$$

The circuit in figure 304 shows the simplest model for a transformer. Therefore, the supply line in figure 302 at the connecting terminals level B-B' will be equivalent to the load shown in figure 305; if the transformer is well regulated, the load impedance converted to the primary R_b will correspond to the characteristic impedance of the supply line Z_0.

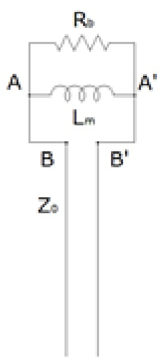

Fig.305

However, the placing of an impedance transformer between the load and the supply line produces an undesired inductance in parallel. Moreover, the said inductance causes a mismatch between the load and the line. The mismatch causes an energy reflection towards the generator and consequently the production of a stationary wave and an SWR on the line.

Hence, at the connecting terminals B-B’ we will find an impedance Z_b that is equal to the parallel resistance R_b and the inductance L_m:

$$Z_b = \frac{j\omega L_m \cdot R_b}{j\omega L_m + R_b} \quad 3.4$$

Accordingly, the coefficient of reflection Γ (Γ reads ‘gamma’) becomes:

$$\Gamma = \frac{Z_b - Z_0}{Z_b + Z_0} = \frac{\frac{j\omega L_m \cdot R_b}{j\omega L_m + R_b} - Z_0}{\frac{j\omega L_m \cdot R_b}{j\omega L_m + R_b} + Z_0} = \frac{-Z_0 R_b + j\omega L_m (R_b - Z_0)}{Z_0 R_b + j\omega L_m (R_b + Z_0)} \quad 3.5$$

Where Γ is a complex number, the modulus of which is:

$$|\Gamma| = \frac{\sqrt{(Z_0 R_b)^2 + \omega^2 L_m^2 (R_b - Z_0)^2}}{\sqrt{(Z_0 R_b)^2 + \omega^2 L_m^2 (R_b + Z_0)^2}} \quad 3.6$$

The equation 3.6 can be considerably simplified if we remember that its aim is the matching of the load R_b to the characteristic impedance of the line Z_0; so, it needs to satisfy the condition:

$$Z_0 = R_b \quad 3.7$$

Substituting 3.7 in 3.6 we obtain:

$$|\Gamma| = \sqrt{\frac{R_b^{\ 2}}{R_b^{\ 2} + 4\omega^2 L_m^2}} \quad 3.8$$

The equation 3.8 links the magnetizing impedance L_m to the modulus of the coefficient of reflection Γ. Through some passages it’s possible to obtain L_m from equation 3.8:

$$L_m = \frac{R_b \sqrt{1 - |\Gamma|^2}}{2\omega |\Gamma|} = \frac{R_b \sqrt{1 - |\Gamma|^2}}{4\pi f |\Gamma|} \quad \mathbf{3.9}$$

Equation 3.9 is very useful because, having set a maximum value of the reflection coefficient, we can obtain, at the lowest frequency of use, the minimum value that the magnetizing inductance L_m needs to have to be inserted between the line and the antenna in figure 302. Of course, a higher value makes the transformer presence less invasive.

For those who are more familiar with SWR I’d like to remind that:

$$|\Gamma| = \frac{SWR - 1}{SWR + 1} \quad 3.10$$

therefore, if substituted in equation 3.9 we obtain:

$$L_m = \frac{R_b\sqrt{SWR}}{2\pi f\,(SWR-1)}$$ **3.11**

For those who are more accustomed with reasoning in terms of return loss RL instead, I'd like to remind that the latter is linked to the reflection coefficient modulus as shown in the following formula:

$$|\Gamma| = 10^{-\frac{RL}{20}}$$ 3.12

It could also be helpful to refer to a table (Table 1) where the corresponding values can be found more easily (taken from [3]).

Table 1										
$\|\Gamma\|$	0.024	0.032	0.048	0.050	0.056	0.100	0.178	0.200	0.316	0.330
SWR	1.05	1.07	1.10	1.11	1.12	1.22	1.43	1.50	1.92	2.00
RL (dB)	32.3	30.0	26.4	26.0	25.0	20.0	15.0	14.0	10.0	9.6

The main effect we experience in the model in figure 302 is an increase in the SWR, even though it is not the only thing that happens. The reactive element L_m, which changes its impedance depending on the frequency, forms a high-pass filter whose cutoff frequency at -3dB depends on the other elements in the circuit too. An utmost simplification of the model gives us the circuit in figure 306:

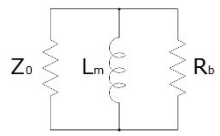

Fig.306

The resistance seen by the inductor L_m it is the parallel of Z_0 with R_b and we are going to call it R_{eg}; but given that we assumed that Z_0=R_b, we will have:

$$R_{eq} = \frac{Z_0 \cdot R_b}{Z_0 + R_b} = \frac{R_b}{2}$$ 3.13

So, the cutoff frequency of the high-pass filter will come out as:

$$f_{TL} = \frac{R_{eq}}{2\pi L_m} = \frac{R_b}{4\pi L_m}$$ 3.14

Said frequency has to result at least m times (~10) lower than the lowest frequency of use f_{min}. Therefore, L_m needs to be higher than:

$$L_m \geq \frac{mR_b}{4\pi f_{\min}}$$ **3.15**

3.15 shows how the magnetizing inductance is in direct proportion with our m factor; as a result, when the m factor increases, not only does the SWR (which is a consequence of the placement of a transformer) decreases, but we also have a drop in the cutoff frequency of the high-pass filter f_{TL}.

If we wanted to set in advance the value of the SWR we could deduce what value of m satisfies both the relations 3.11 and 3.15. From the merging of the two we then obtain:

$$m \geq \frac{2\sqrt{SWR}}{SWR-1}$$ **3.16**

Unless we have very low SWR values, the condition is easily satisfied. As I said previously, the chosen cutoff frequency of the high-pass filter is usually at least 10 times lower than the minimum frequency of use (m=10). This implies obtaining an SWR that is around 1.22, an RL of 20dB and a $|\Gamma|$ = 0.10.

If we want to consider the magnetizing impedance X_m in parallel with the load, from 3.15 we obtain:

$$4\pi f_{\min} L_m \geq mR_b$$ 3.17

Considering that $X_m = 2\pi f_{min} L_m$ we will have:

$$X_m \geq \frac{mR_b}{2}$$ **3.18**

If we take again m=10 we can find the usual rule which expresses that the impedance transformer has to have a magnetizing impedance X_m at least 5 times higher than the load one R_b.

Parasitic effects of the components

In addition to what we just explained, there are other side effects caused by the insertion of a matching transformer in the model shown in figure 302. These side effects are substantially linked to the parasitic effects deriving from the components used.

Mainly: losses in conductors, losses in the ferrite and parasite capacity. The common model of a transformer is the one represented in figure 307.

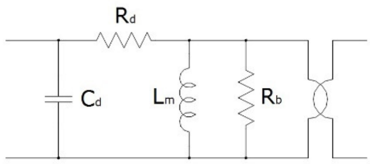

Fig.307

Conductors

Losses in conductors, both primary and secondary, can be traced back to a resistance R_d connected in series to the primary winding (figure 307). Said resistance takes into account the skin effect of the two windings and it is highly dependent on the frequency, on the conductive material used, on its section and length. The resistance R_d influences the attenuation in the transformer bandpass, which is the available one [2]; said resistance depends on the work frequency, because the skin effect experienced with conductors, both primary and secondary, causes the increase of the resistance rate by the square root of the frequency [4].

For non-ferromagnetic conductors that are solid cylinders (at 20°C, with diameter that is much bigger than the depth of penetration δ) we can estimate the conductor resistance (assumed rectilinear, not winded) with the help of the formulas in table 2; it can be found in [6] part 1, where l_w expresses the length and d_w the diameter of the wire (both expressed in mm) and f in MHz.

This estimate is not accurate because it does not assume the conductor as winded, so it fails to take into account the interference caused by the nearby turns (known as proximity effect) that could produce a resistance even 5.8 times higher than the value calculated with the formulas in table 2 [6].

Table 2	
Material	**Resistance in Ω; with *f* in MHz; l_w and d_w both in mm.**
Silver	$R = 79.75 \cdot 10^{-6} \frac{l_w}{d_w} \sqrt{f}$
Copper	$R = 83.04 \cdot 10^{-6} \frac{l_w}{d_w} \sqrt{f}$
Aluminum	$R = 106.28 \cdot 10^{-6} \frac{l_w}{d_w} \sqrt{f}$

Hence, where the current flow is high, it is useful to minimize the losses by choosing the shortest possible conductors, that have a wide section and that are made with materials with a low resistivity. On the contrary, in signal transformers, the conductor resistance is not the main problem.

Ferrite

Even ferrite has losses that can be traced back to a resistance R_s connected in series to the magnetizing inductance L_m. Losses in ferrite can be minimized by choosing with shrewdness the type of ferrite; particularly, we need to have in the transformer bandpass a lower rate of electromagnetic permeability μ_s'' (read mu) than the electromagnetic permeability μ_s' (where the subscript $_s$ stands for series).

I'd like to remind that the parameter μ_s' is the one that contributes to the creation of the inductance in series L_m (the auto inductance of the winding) according to the following formula:

$$L_S = A_L \cdot N^2 = 4\pi \frac{N^2 A_e}{l_e} 10^{-9} \mu_s' = L_0 \mu_s' \quad \text{[H]} \qquad 3.19$$

A_L is the inductance factor in nH/sp^2 (the inductance produced by one turn), L_0 represents the inductance of the winding without the ferrite, N is the number of turns, A_e is the area in cm^2 inside the winding and l_e is the length of the magnetic circuit in cm (the last two parameters need to be taken from the datasheet).

$$L_0 = 4\pi \frac{N^2 A_e}{l_e} 10^{-9} \quad \text{[H]} \qquad 3.20$$

On the other hand, the parameter μ_s'' is the one that generates the dissipative effect in the ferrite (heat), which is formed by a resistance in series R_s, whose value is determined by this formula:

$$R_S = 2\pi f L_0 \mu_s'' \quad \text{[Ohm]} \qquad 3.21$$

As shown in 3.21 that R_s is linearly dependent from the frequency of use, just like the reactance X_s, which is produced by the inductance L_s.

In addition to that, the permeabilities μ_s' e μ_s'' vary at the varying of the work frequency in a not linear way; therefore, it is necessary to refer to the values indicated on the datasheet of the chosen ferrite. Moreover, there can be variations from batch to batch in the aforesaid parameters, which can fluctuate from +/-20%; so, it is not rare to find permeability values that differ even 40% one from the other.

Capacity

Afterwards, we have to minimize the losses that are caused by the parasite capacity which generates between the winding turns. These generate because there is an electric potential difference between the transformer turns (and also between the turns and the ferrite) and also because there's a delay in the EM wave propagation that flows through the primary and secondary winding. Parasite capacities render with a capacity in parallel (C_d)

at the input connecting terminals B-B' of the impedance transformer. Said capacity is like a low-pass filter and it lowers the performance of the transformer at high frequencies (figure 307).

It is not an easy task to reduce (nor estimate) the parasite capacity C_d. One of the ways of doing that is to keep a low tension between adjacent turns (in other words by keeping a low voltage per turn); the problem with this method is that it causes an increase in the number of turns needed and with them, also the conductors' length and eventually even the parasite capacity. Another method, which is a bit easier is to distance the turns one from the other. They can be distanced in a power transformer or, in the case of a signal transformers it would be ideal to use a wire with a thick insulating material around it or really thin wires (where the area exposed is less).

For the purpose I would like to remind that the capacity between two adjacent rectilinear wires is expressed by the following formula:

$$C = \frac{\pi \cdot \varepsilon_0 \varepsilon_r \cdot l_w}{\cosh^{-1}(\frac{D}{d_w})} = \frac{27.8 \cdot 10^{-12} \varepsilon_r \cdot l_w}{\ln\left(\frac{D}{d_w} + \sqrt{\frac{D}{d_w} - 1}\right)} \quad [F] \qquad 3.22$$

Where D expresses the distance between the wires, d_w the diameter of the wires (assumed to be identical) both in the same unit of measurement (mm) and l_w the length of the conductor in meters.

Formula 3.22 is rather complex, but it tells us that, if we assume the type of insulating material to be the same (ε_r read as epsilon-r), the capacity decreases with the increasing of the distance between the conductors D, but it increases with the increasing of: the diameter of the conductor d_w and the length of the conductors l_w. To reduce the capacity, if any choice is possible, 3.22 advices us to use conductors that are short, thin and spaced out. The use of short wires (in other words with few turns) lowers the propagation delay of the wave that flows through them.

Whoever was interested in further delving into the subject, I recommend checking [6].

It is always advisable, when possible, to wind half of the coil with the highest number of turns, then, over that one, winding the one with the lowest number and ultimately, on top of the previous two, winding the remaining half of the coil with more turns. In this way you create a sort of 'sandwich' that lowers the parasite capacity [2].

Power transformers

Lastly, when a significant amount of power flows through the transformer we need to verify: firstly, that the flux density B_{max} does not exceed the admissible amount at the temperature of use; secondly, that the heat generated by losses in the conductor and in the ferrite does not make the temperature in the transformer rise to the point of reaching the Curie temperature.

In bibliography [2] you can find a relation which links the volume of the ferrite V_e to the magnetizing inductance L_m, the real material permeability ($\mu_e \approx \mu'$ in a toroid), the density value of the chosen flux B (usually half of B_{max}) and the tension E to which the primary winding is dependent (leaving out the drop in the loss resistance of conductors R_n).

$$V_e = A_e l_e = \frac{20 \mu_e E^2}{\pi \cdot f^2 L_m B^2} \quad [cm^3] \qquad \mathbf{3.23}$$

With the frequency f expressed in MHz, the primary magnetizing inductance L_m in microhenry and the flux density B in Gauss.

Beyond the result that we obtain, formula 3.23 is very revealing, because it links the different parameters to the quantity of ferrite that's needed. From this we can conclude that the volume of the ferrite increases with the square of the chosen tension and, given the power P is:

$$P = \frac{E^2}{R_b} \quad \text{[W]} \qquad 3.24$$

it follows that the usable ferrite volume increases linearly with the power involved in the system. The presence of μ_e in the numerator can be misleading, but note that it is also present in the denominator, even though it is hidden in the magnetizing inductance L_m (see formula 3.19).

For what concerns losses in conductors it’s useful to remind that they’re caused by the flowing of the current (by joule heating) in a layer of skin of the conductor that can be more or less thin. In other words, the conductor will have a variable resistance R_d=R_1+R_2, that is in direct proportion with the highest frequency of use (see [4], [6] and table 1).

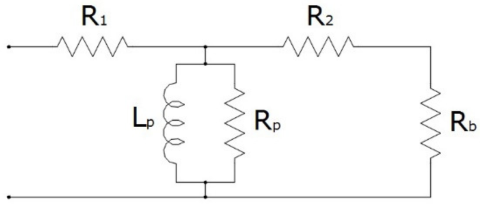

Fig.308

Taking as a model the transformer in figure 308, where the resistances R_1 and R_2 belong to the primary and secondary respectively, at a given frequency, R_b is the load frequency seen by looking into the primary, L_p is the parallel magnetizing inductance and R_p is the parallel loss resistance in the ferromagnetic material.

L_p and R_p (the subscript stands for parallel) can be obtained by transforming the series impedance of the ferrite (figure 309) in its parallel equivalent (figure 310) [5]. I would like to mention that the series permeabilities μ_s' e μ_s" depending on the frequency, are the ones that can be found in the datasheet.

Fig.309

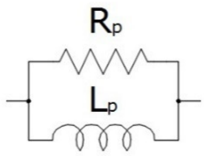

Fig.310

$$Tan\delta = \frac{1}{Q} = \frac{R_s}{2\pi f L_s} = \frac{\mu_s''}{\mu_s'}$$

$$Q = \frac{1}{Tan\delta}$$

$$L_0 = \frac{4\pi N^2 A_e}{l_e} 10^{-9} \text{ [H]}$$

$$R_p = R_s\left(1+Q^2\right) \text{ [}\Omega\text{]}$$

$$R_s = 2\pi f L_0 \mu_s'' \text{ [}\Omega\text{]}$$

$$L_p = L_s\left(1+\frac{1}{Q^2}\right) \text{ [H]}$$

$$L_s = L_0 \mu_s' \text{ [H]}$$

To minimize the losses, it is necessary that the ones in the conductors are equal to the ones in the ferromagnetic material. This happens when:

$$\frac{R_d}{R_b} = \frac{R_b}{R_p} \text{ that is } R_p = \frac{R_b^2}{R_d} = \frac{R_b^2}{R_1 + R_2} \qquad \textbf{3.25}$$

Failing to respect 3.25 will cause an increase in the losses in the power transformer [2].

In the end, it is not possible to avoid losses in a transformer but we can at least minimize them. They produce heat, which will cause the temperature of the magnetic material to rise; this must never reach the Curie temperature, because if that was the case, the magnetic material would lose all its magnetic properties and it would become unusable.

Example

After all the theorical explanation, let's try to put the above things into practice and draw from it some useful formulas to build a ferrite transformer.

For example, we want to build a transformer that is capable to adapt the impedance of a coaxial TV line Z_0=75 Ω to a purely resistive impedance R= 1200 Ω, which is typical for a receiving antenna for 160m, just like Double Half Delta Loop (DHDL). For the purpose we are going to use a binocular ferrite, Amidon type BN-73-202 or Fair-Rite 2873000202 and we are going to set the Return Loss at 20dB (corresponding to an SWR of 1.22).

From 3.16 we can draw that in order to obtain an SWR=1.22 we need to have an m value that is at least equal to:

$$m \geq \frac{2\sqrt{SWR}}{SWR-1} = \frac{2\sqrt{1.22}}{1.22-1} \approx 10$$

We are going to take m=10 for convenience.

Combining 3.15 with 3.19, after a few passages, we obtain:

$$n_P \geq \sqrt{\frac{mR_b}{4\pi f_{\min} A_L}} \qquad \textbf{3.26}$$

Where n_p are the turns in the primary (the one attached to the line at 75 Ω) and A_L is the inductance factor in nH/sp^2 of the ferrite (it is the inductance produced by just one coil turn), which we do not know, but we can easily determine by measuring the minimum frequency of work with the method shown in [7]. In my case I found A_L=5540 nH/sp^2 at 1.8MHz; so, from 3.26 we will have:

$$n_P \geq \sqrt{\frac{mR_b}{4\pi f_{\min} A_L}} = \sqrt{\frac{10 \cdot 75}{12.57 \cdot 1.8 \cdot 10^6 \cdot 5540 \cdot 10^{-9}}} = \sqrt{5.983} = 2.44 \text{ turns in the primary.}$$

For convenience we have chosen n_p=3 turns; the formula above satisfies the conditions in 3.26 and at the same time it limits the number of turns needed (therefore also the length of the conductors) in order to minimize the losses. We can also obtain from 3.1:

$$n_S = n_P \sqrt{\frac{R_L}{R_b}} \qquad \textbf{3.27}$$

3.27 allows us to find the number of turns of the secondary winding n_s:

$$n_S = n_P \sqrt{\frac{R_L}{R_b}} = 3\sqrt{\frac{1200}{75}} = 12 \text{ turns.}$$

The result is not always an integer, we can often find decimals. If that is the case the number of turns can be increased to reach the closest integer or half.

In the case of a binocular ferrite it is possible to wind even the half turns because you count one on a turn only when the wire has gone through both the holes; if it goes through only one of them, we have a half turn. This way we can approximate more easily the number found to the real number of turns.

Let's now move on to the making of the transformer. First of all, we need to identify the conductor and its diameter. The most accessible conductor is the one used for electric engine windings, which is nothing more than enameled copper wire. The section needs to be small to avoid the capacitive effects. Keep in mind that the available area (the diameter of a BN-73-202 is 3.8 mm) can only fit 20 wire turns that have a diameter of 0.50mm because the load factor of a handmade winding never exceeds 80% of the available area. Personally, for signal transformers, I usually use an AWG30 wire wrap wire because it is easier to peel. Its section measures 0.50mm while its core measures around 0.25mm and it can be found in different colors.

Now we need to find the average length of the turn. The easiest and quicker method is to wind a turn with the chosen wire, then we need to mark the intersection point and measure the distance between the two resulting marks (figure 311). In our case the average length of the turn is l_{wm}=40 mm. Hence, from this we can calculate the length of the primary and secondary conductor, remembering to lengthen the pieces by adding a turn for the connecting wires.

$$l_{WP} = l_{wm} \cdot (n_P + 1) = 40 \cdot (3+1) = 160 \quad \text{[mm]} \qquad 3.28$$

$$l_{WS} = l_{wm} \cdot (n_S + 1) = 40 \cdot (12+1) = 520 \quad \text{[mm]} \qquad 3.29$$

At this point the only thing left to do is to wind the coil. Let's begin to wind half of the secondary (six turns) because it is the one with the highest number of turns. Once that is done, let's wind all the primary one (three turns). Next, we need to complete the secondary winding by adding the remaining 6 turns. This way the parasite capacity is minimized and the band is broadened.

When the transformer is done (figure 311) let's remember to: to apply a sticky label on the right side of the ferrite that states at least the impedance of the primary and the secondary and to scrape away the enamel layer from the connecting wires if we used an enameled wire.

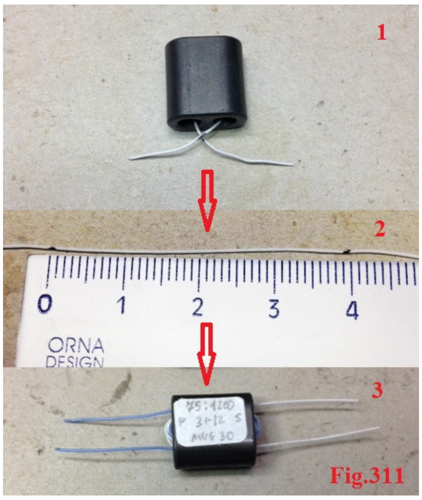

Fig.311

The transformer is now completed, it just needs to be checked to see if theory corresponds to reality.

The first parameter to check is the coupling ratio. To do so we need to connect a small trimmer to the secondary connecting wires and an antenna analyzer to the primary (calibrated) [8] which has a typical impedance of 50 Ω. So, we need to set the trimmer according to:

$$R_L = R_b \cdot \left(\frac{n_S}{n_P}\right)^2 = 50 \cdot \left(\frac{12}{3}\right)^2 = 800\Omega \qquad \mathbf{3.30}$$

The turns ratio can be verified at 50 Ω.

The antenna analyzer will show an SWR rate around one when it is connected to a resistive load of 50 Ω. In our case there's an impedance in parallel caused by the magnetizing inductance L_m, so, the SWR at the lowest frequency cannot be worse than 1.22. The value of the minimum frequency of work has an SWR=1.07 at 1800kHz.

The second parameter to determine is the magnetizing inductance L_m. To do so we're going to measure the inductance of the primary with the secondary open L_{PO}=75.5µH and then in power surge L_{PC}=0.4µH [9]. So, from 3.2 we obtain the coupling factor *k:*

$$k = \sqrt{1 - \frac{L_{PC}}{L_{P0}}} = \sqrt{1 - \frac{0.4 \cdot 10^{-6}}{75.5 \cdot 10^{-6}}} = \sqrt{0.9947} = 0.997 \approx 1$$

As expected, it is around one. Consequently, the magnetizing inductance L_m obtained from 3.3 is:

$$L_m = k \cdot L_{P0} = 75.3 \mu H$$

The cut frequency of the high-pass filter, from 3.14 will be:

$$f_{TL} = \frac{R_b}{4\pi \cdot L_m} = \frac{75}{12.57 \cdot 75.3 \cdot 10^{-6}} = 79.24 kHz$$

The value turned out to be much lower than the lowest frequency of use f_{min}=1.8MHz, therefore, from 3.15, the value of L_m measured is over double the necessary:

$$75.3 \mu H \geq \frac{mR_b}{4\pi \cdot f_{\min}} = \frac{750}{12.57 \cdot 1.8 \cdot 10^{6}} = 33.15 \mu H$$

Even though my instruments are not very accurate all the measurements have been carried out at 50 Ω and I am rather positive that the transformer that I built turned out to be acceptable.

Considerations

In this paragraph I would like to show to the reader the electric characteristics that I draw from three ferrite transformers that I built as indicated in the above paragraph, but using different wires. The ferrite that I chose is always an Amidon BN-73-202, where we have 3 turns for the primary and 12 for the secondary, therefore with the same volt per turn; The measurement is carried out with the primary connected to the VNWA of DG8SAQ, having an input impedance of 50 Ω and the secondary to a trimmer set at 800 Ω. As you can see in figure 312 the three transformers are made respectively with: 0.50mm enameled wire, an AWG 30 wire wrapping wire and 0.25 enameled wire.

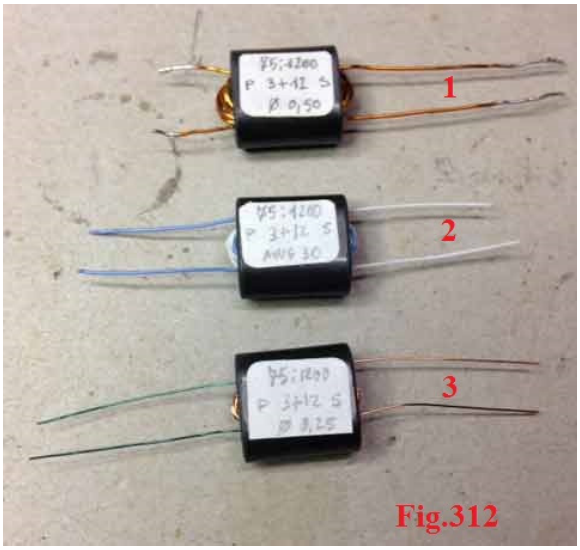

Fig. 312

As showed by the following graphics the return loss (RL) is the module of S11 in dB (which is equal to the S11 taken with positive sign) and it is represented by the red line; the marker #4 is set at RL=20dB. In addition, the measurements are made at 50 Ω and not at 75 Ω, as stated in the project.

1) with 0.50mm enameled wire

Fig. 313

2) with AWG 30 wire wrapping wire (ϕ_e=0,50 ϕ_i=0,25)

Fig.314

3) with 0.25mm enameled wire

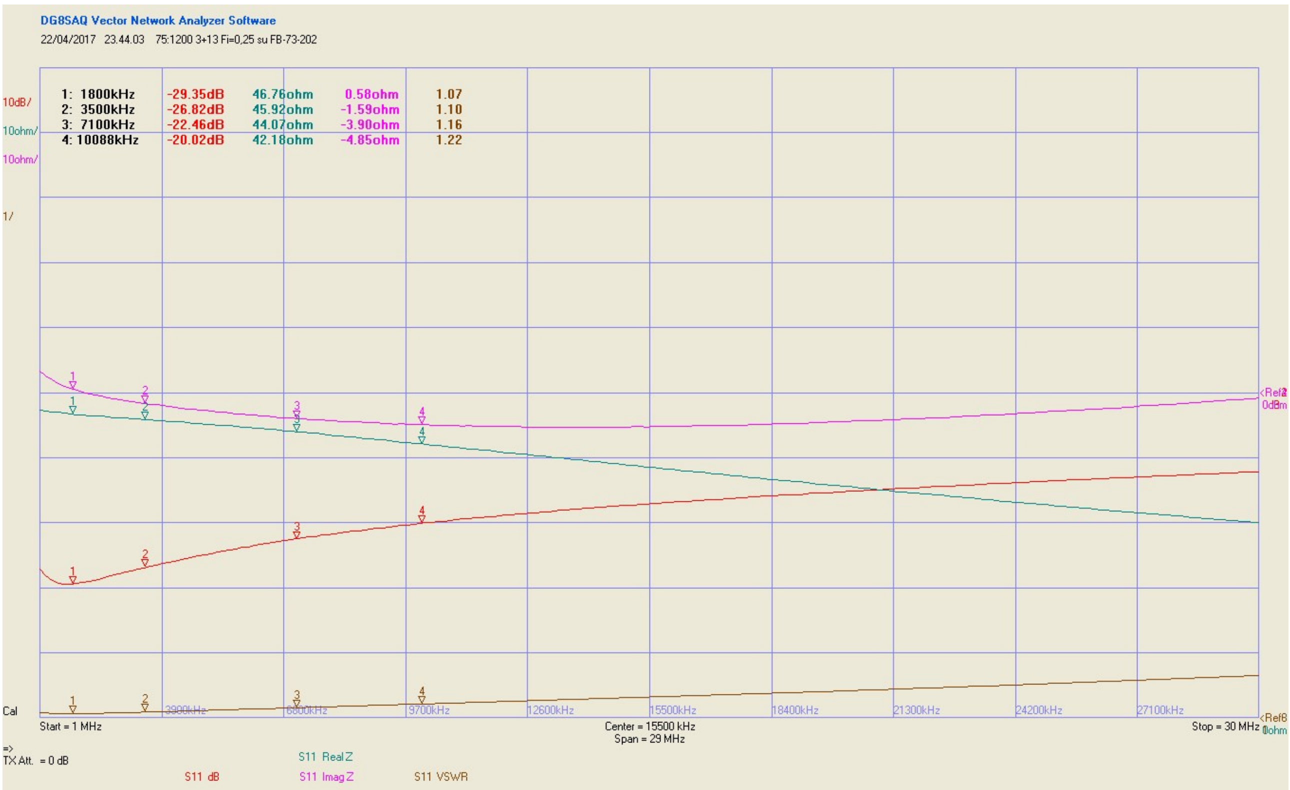

Fig.315

It is clear that the frequency where the RL=20dB (underlined by marker #4) in the first case is around 1.67MHz, in the second one 27MHz and in the third 10MHz. This confirms that the best choice between the three is the AWG 30, even though I have nothing more that speculations on why that is.

Lastly, I made two identical transformers with AWG 30 wire wrap wire. I connected the secondary ones together and the primaries one to the RX input and the other to the VNWA output TX, properly calibrated. Then I measured the attenuation of the two transformers in series (figure 316).

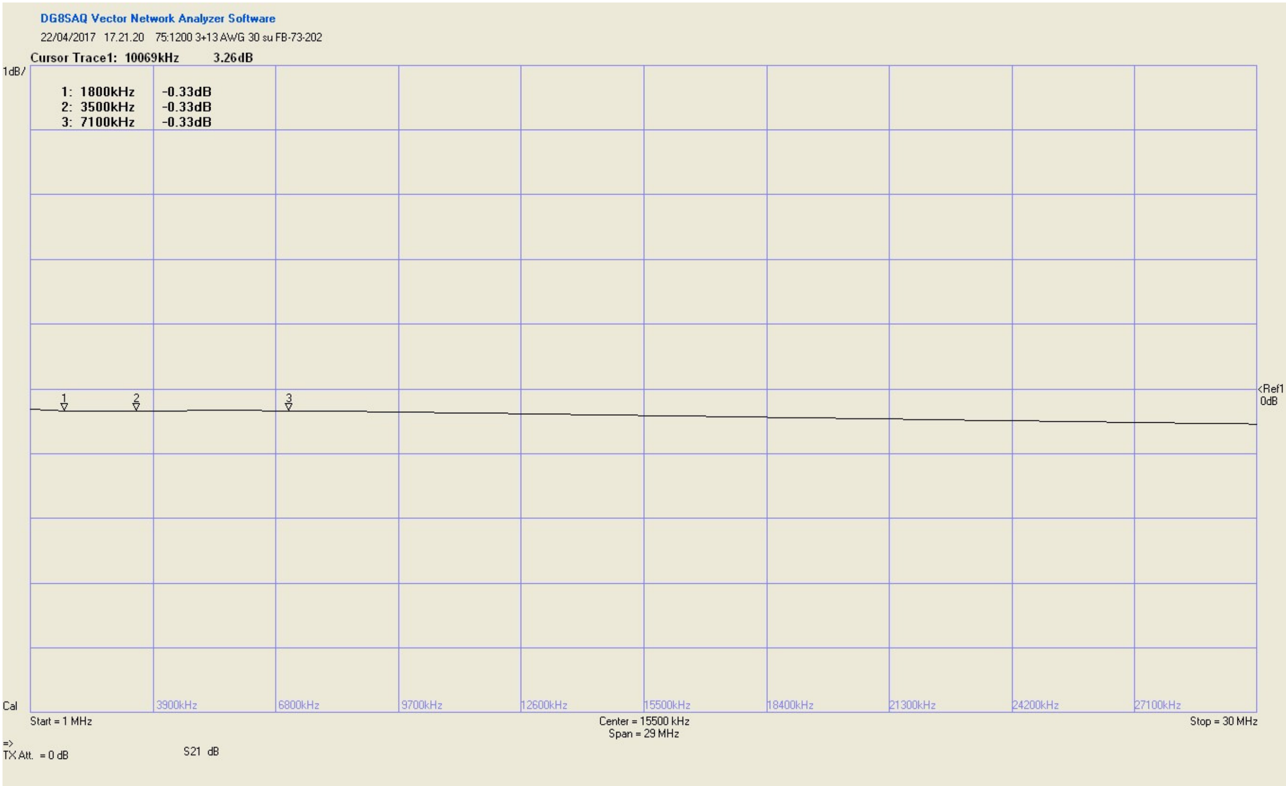

Fig.316

As you can see it is fixed at around -0.33dB, that means that one transformer attenuates -0.16dB, which is a pretty considerable result.

Bibliography

[1] C. A. Balanis; Antenna Theory analysis and design, 2nd ed. John Wiley & Sons, 1997.

[2] E. C. Snelling; Soft Ferrites, ed. Ilieffe Books Ltd, 1969.

[3] D. M. Pozar, Microwave Engineering, 3° ed. John Wiley & Sons, 2005.

[4] S. Ramo, J.R. Whinnery, T. Van Duzer; Campi e onde nell'elettronica per le telecomunicazioni, 11° ed. Franco Angeli, 2016.

[5] H.P., Operating and service manual 4261A digital LCR meter, ed. Yokogawa-Hewlett Packard Ltd, 1976.

[6] David Knight, G3YNH Components and materials http://g3ynh.info/zdocs/comps/

[7] http://www.iw2fnd.it/sites/default/files/docs/Misura_di_Ferriti.pdf, pubblicato su RR7/8-2014.

[8] Misura eseguita con RigExpert AA-230 Zoom.

[9] Misura eseguita con LCR Meter HP 4261A @1kHz.