ON9CVD

"Communicate, and Violence Disappears"

The simple 1 : 1 balun is being know in a variety of constructions. A dedicated chapter seems appropriate

---

Select one of the following subjects:

- Introduction to baluns
- Wire baluns
- Baluns on cores
- High-Z balun

Baluns have been discussed almost from the beginning of radio-communication and articles on this

subject
are still appearing in ham-magazines. Trying 'Balun' in Google already yields million of hits.
Please contact me for more information and / or your remarks on the subjects as above, at:
on9cvd@amsat.org

Bob J. van Donselaar

Balun

(published in Electron #4, 2007)

Introduction

Baluns have been around from the beginning of wired and wireless communication. Baluns in radio antenna systems in particular have been discussed from the early radio-days and are may still being found in current radio magazines. Although the balun is a simple and straight forward circuit component, the design and application has often been surrounded by some sort of 'magic', with magistrates riding hobby-horses. Despite the long balun history, or may be just because of this, the subject of baluns is still popping-up regularly 'on the waves' and therefore it seems a good idea to put various aspects around these components a little bit into perspective in a short series of articles.

Types of baluns

Let's first take a short look into some balun applications in radio-communication.
The word 'balun' is an assembly of **bal**ance to **un**balance transformer, a transformer to connect a balanced system to an unbalanced system without influencing parameters in either system. Balanced to unbalanced transitions are part of our everyday radio-life as in balanced microphones for more dynamic range, at inputs and outputs of balanced amplifiers either base band audio or at HF, feeding balanced antenna systems etc. Baluns therefore may be discovered at various places and it is clear the same functionality to look somewhat different at different parts of the systems.

- Around a balanced microphone a low-power balancing voltage-transformer may be an optimal component.
- At the input of a balanced HF FET amplifier, both branches should be fed with a symmetrical voltage.
- At the output of the same amplifier currents should be balanced for symmetrical loading and canceling of even harmonics.
- At a symmetrical antenna, both halves should be fed with an equal and opposite phase current to obtains a symmetrical radiation pattern.

These are only a few of many situations requiring a balun at the transition point. It is clear each situation in the above examples to require a somewhat different approach and before we start thinking about baluns, it is obvious we should first recognize the particular 'problem' we are trying to solve and the voltages, currents and impedances around the application.

In almost all modern commercial transceivers we find a balanced, semi-conductor amplifier at the transmitter output stage, that is coupled to an a-symmetrical output terminal by means of a wide-band balun. Various options are open to connect this a-symmetric output to the antenna, of which one is to apply an a-symmetrical feed-line to a balanced dipole antenna, again through a wide band balun. This we recognize as a double 'transformation' of which the last part has to be constructed by the owner of the system.

This double transformation did become popular after the last world war, when much ex-army material became available to destitute European radio-hams. Part of this army dump was consisting of loads of coaxial transmission- lines since this had shown to be a reliable means for quickly rigging up transportable transmitter stations in the field. Before this period almost all transmitter stations where connected to the antenna through symmetrical feed-lines, that not only exhibited very low loss but were also easy to mach to the high(er) impedance of the tube transmitters at that time.

Pro's and con's of various antenna feeding methods have been discussed more extensively in the chapter "Where does the power go", and appear to be closely related to the impedance (range) of the antenna system.

Why unbalance

In this article-series baluns will be discussed mainly for connecting a-symmetrical feed-line to balanced antenna's, to ensure equal but opposite phase currents in each dipole half and no parasitic currents outside this circuit.

When designing symmetrical antenna's in an antenna design environment (e.g. Mmana or EZNEC), we ensure the design to be perfectly symmetrical and find radiation patterns to match this symmetry, so why should the practical antenna currents deviate from this perfect symmetry?

In practice dipole antenna's rarely are perfectly symmetrical because of all sorts of obstacles around or near this antenna, (a-symmetric) vicinity of trees, wires to couple to metal parts (roofing, drainage), different soil-types or soil humidity etc.
A second important factor is the feed-line itself. When an a-symmetric feed-line is connected to a symmetrical antenna, both antenna halves 'see' the impedance of this feed-line, with one halve to also 'see' the outside of the a-symmetric feed-line. If this additional wire exhibits a high impedance as compared to the antenna at the feed point, relatively little current is 'leaking' to this additional antenna. This parasitic antenna however may also exhibits a low impedance depending on the length of the feed-line. In this latter situation the out-side feed-line may become the main antenna with a completely different radiating pattern than designed for. This will not only be noticed at your distant communication party since the parasitic antenna is 'stealing away' communication power, but also at your neighbors when they unwillingly are listening to your radio-contacts through the stereo equipment. It may even be apparent at your position at the controls when HF feedback creeps into your equipment or make your transceiver 'hot' to the tough.

When receiving reciprocal problems arise with the unwanted additional antenna to tap into the 'electro-smog' around your house and your neighborhood, since this type of noise usually is vertically polarized.

All in all it may be clear that balanced (antenna) systems should remain balanced and should be isolated from unbalanced feed-lines by means a proper balancing device. It is also clear this device should be of the current balancing type, also known as a 1 : 1 current transformer or sleeve choke.

Current balun

In a current transformer, conductors are tightly coupled; a current in one conductor will have a current flow in the other of equal magnitude but opposite phase, a schematic example may be found in figure 1.

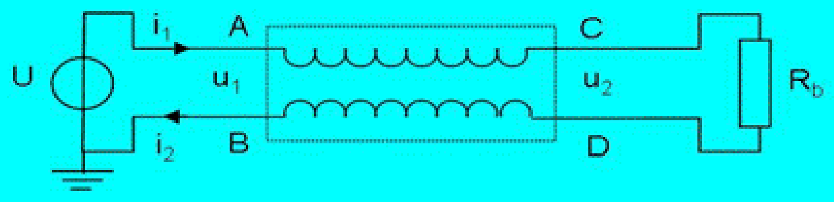

Figure 1. Schematic diagram of a 1: 1 current transformer

This 1 : 1 current transformer consists of windings of equal turns that are tightly coupled. Current 'i_1' is entering the transformer at terminal A and is leaving at terminal C. The current flowing in winding A-C, will induce a current in winding D-B of equal magnitude, that is leaving as 'i_2' at terminal C.

This current will introduce a voltage across load resistor R_b of magnitude $u_2 = i_1$ ($= -i_2$) x R_b. Since this transformer has equal number of turns for both windings, and therefore equal impedance, an equal voltage will appear across terminals A-B, making $u_1 = u_2$. This current transformer also forces currents i_1 and i_2 to be equal and equal to the current through the generator, provided no other currents than i_1 and i_2 can flow anywhere in the circuit.

This description is similar to what is happening inside a transmission-line. Again the conductors making up for this line will be carrying currents of equal magnitude and opposite phase because all of the electromagnetic filed is forced to stay 'inside' the transmission-line. Let's look at figure 2 of the transmission-line as a 1 : 1 current transformer.

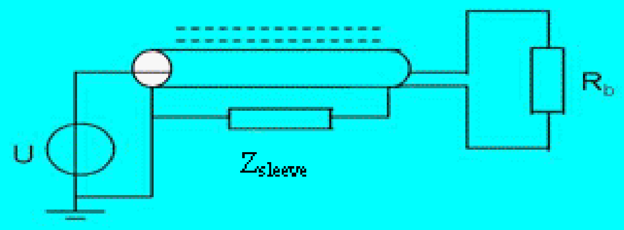

Figure 2 Transmission-line as a 1 : 1 current transformer.

In figure 2 a transmission-line is depicted. When terminated into its characteristic impedance, this will absorb all transfered power with no reflection of energy back to the generator. This also requires a voltage at the output of the line to be equal to the voltage at the input (for an ideal transmission-line). Again transmission-line currents are equal and of opposite phase and equal to the current through the generator. As in a current transformer this is true as long as no other currents will flow.
Since the output of the transmission-line is 'free floating' this may be connected to any other position in the circuit. If, for some reason the bottom part of the load resistor R_b is at a voltage U with respect to ground (e.g. when connected to the top of the generator), an additional current will flow through the outside of the transmission-line, that is no longer coupled to the currents at the inside. Since this current also has to be delivered by the generator, this leakage current is lost from the load. In terms of a transceiver circuit, power as delivered to the antenna system by the transceiver (generator) will no longer be only absorbed by load R_b, the radiation resistance, but also by a (dummy?) load Z_{sleeve}. In the receiving position energy as received by the antenna will no longer be delivered to the receiver only.

The amount of additional current will depend on the impedance of the outside of the transmission-line, depicted in figure 2 as the sleeve impedance Z_{sleeve}. At very long transmission-lines, this current will be very small, but so will the currents through the transmission-line by means of internal loss. At short lines we may enhance the ratio of transmission-line current to sleeve current by enhancing the sleeve impedance. This will make no difference to the transmission-line currents, since these are 'locked-up' inside the line and are running as a differential-mode current. So either in a straight line, curled up or made into an inductor makes no difference to the transmission-line current, because no differential current will be generated as the outside current is common-mode only. This common-mode current may be further diminished by winding around a high permeability ferrite core (the dashed line above the transmission-line). With this construction we created a common-mode choke to block off any outside current, again approaching the ideal current transformer of figure 1.

Current transformer as a balun

Let's look at an antenna system with the above transmission-line current transformer as a balun to connect a symmetrical dipole antenna to an a-symmetrical feed-line, For this model the antenna radiation resistance has been split-up in two over each dipole half. To close the loop, the ground return current has been depicted by the ground symbol at each end of the antenna and the ground symbol at the transceiver. This model may be found in figure 3.

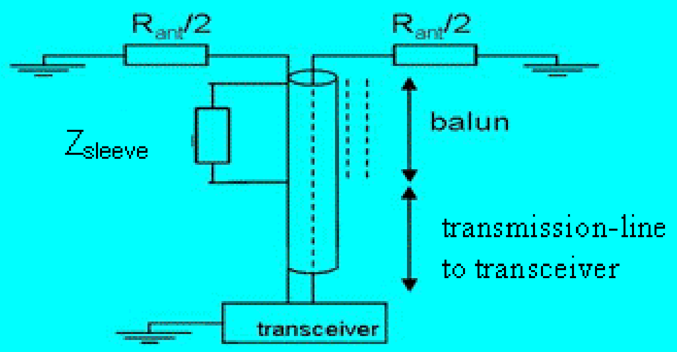

Figure 3. Current balun model

The generator of figure 2 is the transceiver in figure 3 with the load depicted as the two halve radiation resistances. Current delivered to the antenna will create voltages across each half of the dipole, creating a voltage gradient at the outside of the feed-line from the transceiver to the antenna. This voltage will create an additional feed-line current that will be 'subtracted' from the antenna current and will therefore not contribute to the dipole radiation. This parasitic current may be minimized if the sleeve impedance is maximized, presented in figure 3 as the impedance Z_{sleeve}.

The effectiveness of the sleeve impedance in minimizing this parasitic current depend on the ratio of Z_{sleeve} to R_{ant}; the higher R_{ant}, the higher Z_{sleeve} should be. This is especially important when the (dipole) antenna is driven outside resonance and R_{ant} is relatively high. To still be effective at minimizing parasitic currents, Z_{sleeve} should be higher still, a lesson that is often forgotten when applying a balun to an antenna.

When testing balun effectiveness it should be checked at the highest operational load impedance and in a maximum unbalance situation, with the Z_{sleeve} to carry the same voltage as the load. At these test conditions insertion-loss and SWR should still satisfy the specifications to qualify. In all test and measurement conditions in the following chapters, baluns will always be tested under these maximum unbalanced conditions to ensure the antenna system to operate in all practical situations.

So sleeve impedance should be high, but how high is high enough? A rule of practice is the parasitic current to be lower than 1/4 of the load current to be insignificant enough. Since power is scaling with current squared, the amount of power lost under this parasitic current regime will result in an insignificant deviation of the S-meter at the receiving side as compared to an ideal no-loss condition.

To reach this current condition, the impedance of Z_{sleeve} in practice should be in the 150 - 300 Ohm range when the balun is applied with a resonating dipole antenna. This impedance value is not hard to obtain with various types of coils (turns of transmission-line), either with or without a ferrite core.

Analyze your antenna system

Baluns will be applied at various antenna systems and frequency range . At designing or acquiring these components however, one should be very much aware of the impedance range the balun is specified for related to the range of impedances that will occur in your specific situation. Many such baluns on the market are designed for 50 Ohm system impedance and will operate according to specifications (power, frequency range) only at that impedance. Under practical conditions this specified impedance range may easily be violated, especially when an antenna is driven outside resonance with voltages to easily become much higher than the component has been designed for.
As an example: A balun that is specified for 250 W. in a 50 Ohm environment will already by overstressed by a factor of two when driven at 100 W. in a 500 Ohm environment, which is still a low value for a dipole out of resonance. Since an antenna tuner usually will translate this 500 Ohm impedance nicely to 50 Ohm again, this condition may easily go un-noticed. Burned baluns and / or damaged transceivers may be the result.

A good analysis of the tuner / balun situation at various impedance levels has been published by Kevin Schmidt, W9CF, in a paper called: 'Putting a balun and a tuner together'. In this analysis, system impedances are discussed under various load conditions, as are requirements to the balun in a balanced feed-line situation. Also a method is being presented to calculate and measure these requirements.

The next articles in this series will discuss various types of (antenna) baluns with and without core material. I would appreciate your comments, questions and remarks to this and any of the other balun articles.

Bob J. van Donselaar
mailto:on9cvd@amsat.org

Trefwoorden

Where does the HF power go

(published in Electron # 4, 2006)

Introduction

Regularly stations are reporting on signal strength as perceived by the S-meter. A long discussion could be held on S-meter calibration but on average these S-meters are showing about the same mark when a signal is received around S-9 on the scale. In Europe, many amateur stations are operating with a peak-envelop-power of around 100 Watt, with perceived signal strength at different receivers of more than one S-point apart. Even receiving stations at about the same distance of the transmitter may note these signal differences, that usually are attributed to 'conditions' that surprisingly often only apply to these weaker stations.

Next to 'conditions' the antenna system could make quite a difference as far as antenna efficiency is concerned and angle of radiation, whether azimuth and / or elevation. Many antenna books have been written on this subject and at the internet some well documented sites are being maintained on this subject.

Even the best antenna however will not be very helpful with the rest of the system between transceiver and antenna not optimized for high efficiency. Many radio-amateurs unfortunately are only vaguely familiar with loss mechanisms in this area. Furthermore many articles in radio-magazines are dealing with aspects of individual constituents (tuner, balun, transmission-line, antenna) but are rarely concerned with the complete system and the interaction between individual system parts. It therefore seems useful to take a look at output-system aspects in a more integral way.

Basic station

In this chapter we will model the complete system from transceiver up to the perceived (relative) power at the receiving station. We will try to find out how much power will be lost in a high efficient antenna system, where this loss is concentrated and what is happening when the same system is not so optimally matched anymore. To set figures into perspective for this exercise we take the basic station to be capable of delivering 100 Watt into 50 Ohm. The (transistorized) transceiver should see SWR < 2 to generate full power, and therefore should be connected to a tuner to correct non-matching situations. The system should be able to operate on all HF amateur frequencies, starting at the 80 m. band.

Antenna
Usually an antenna for the lower HF amateur frequencies is the 'hardest' component because of size for a resonant type. We will take the antenna in this article to be resonant at 3,65 MHz. When using one of the antenna design program's e.g. Mmana (free), the antenna will look something like:
antenna dimensions: 2 x 19,875 m.,
antenna height: 10 m.
above average ground type ($\varepsilon = 5$, conduction = 13 mS).

This antenna will behave like table 1.

Dipole 2 x 19,875 m. at 10 m. above average ground, designed for resonance at 3650 kHz

f (kHz)	**r** Ohm	**X** Ohm	**Z** Ohm	**SWR** re 50 Ohm
3500	44,3	-74,4	87	4,3
3520	45,2	-64,3	79	3,6
3540	46,1	-54,2	71	2,9
3560	47,1	-44,2	65	2,4
3580	48	-34,2	59	2,0
3600	48,9	-24,1	55	1,6
3620	49,9	-14,1	52	1,3
3640	50,9	-4	51	1,1
3660	51,9	6,1	52	1,1
3680	53	16,1	55	1,4
3700	54,1	26,2	60	1,7
3720	55,1	36,4	66	2,0
3740	56,3	46,4	73	2,4
3760	57,4	56,5	81	2,8
3780	58,6	66,6	89	3,2
3800	59,7	76,7	97	3,7

Table 1: Example antenna for 80 m.

The antenna indeed is living up to expectations with almost perfect SWR at the design frequency and with band-width is 140 kHz. between SWR = 2 frequencies. In between these SWR positions the antenna may be applied without an antenna tuner.
Note: The antenna input impedance is a complex number (except at the exact design frequency) so all calculations have to be dealing with these complex numbers, including SWR.

Balun
The antenna in our example has a symmetrical structure that we like to keep that way. To feed the antenna we may select a symmetrical or an a-symmetrical feed-line. We will start-off this example with the a-symmetrical type, so we will have to make a transition from a symmetrical to an a-symmetrical system, e.g. with a balun. More on this last subject may be found in a dedicated chapter on these components.

A balun should make the balance to un-balance transition without further being 'visible' in the system. This 'invisibility' will be ensured with balun impedance at least four times the system impedance:
X_l = 4 x 50 Ohm = 200 Ohm. This should already been ensured at the lowest operating frequency (3,5 MHz.), so

L = 200 / ω = 9,1 •H.
A 36 mm. toroide of 4C65 ferrite material exhibits a winding factor A_l = 170 nH., the number of turns follows from:

$n = \sqrt{(9,1 / 0,17)} = 7,3$ (8).

When constructing this balun as a trifilar flux-transformer, we need four trifilar turns since the antenna and the feed-line will be connected across two windings in series each. Although one sometimes comes across many more turns for these transformers (at 4C65 ferrite), 'more is less' in this situation since parasitic capacitance is increasing with each additional turn, decreasing maximum usable frequency for this balun.
Best 1 : 1 balun however is the sleeve-choke type, consisting of 8 turns of transmission-line on the same toroide.

Even a well-designed balun will not be ideal and will 'consume' some power when connected to the antenna. We will take this power-loss into account in our total picture. Again calculations will be of a complex nature.

Feed-line
In our example design, the antenna plus balun is situated at ten meter above ground, so we need some feed-line to connect to the transceiver in the shack. In this example we will first selected the well known RG58 type of coaxial transmission-line (Belden 8259), with a loss of 2,76 dB per 100 m. at 3,5 MHz. and at a length of 10 m. In practice this transmission-line may be somewhat longer since the cable may not be connected in a straight line from the antenna to the transceiver. The cable loss factor is showing some power will be lost in the feed-line and we will take this into account in our total picture.

Tuner
Before arriving at the transceiver, a tuner will translate line impedance to the required 50 Ohm resistive value, the transceiver needs to generate full power. For this component, a simple L-type tuner has been selected as low-component count usually also means low loss. In this example, the tuning inductor will be in the series branch and is showing a quality factor, Q = 200. This L-type will also perform as a low-pass filter, which may be useful to suppress undesired harmonics.
Depending on the type (value) of the line impedance, the tuning capacitor will be at the line side or at the transceiver side. In our example-tuner a high Q-factor has been selected which is not an impossible when constructing with a large enough wire diameter and not too small inductor dimensions. Nevertheless, some power will be dissipated in this component as well and we will take this into account.

Antenna assembly

With the complete antenna system designed as in our example system, let's find out how this will behave on the amateur frequency band it has been designed for, as in figure 1.

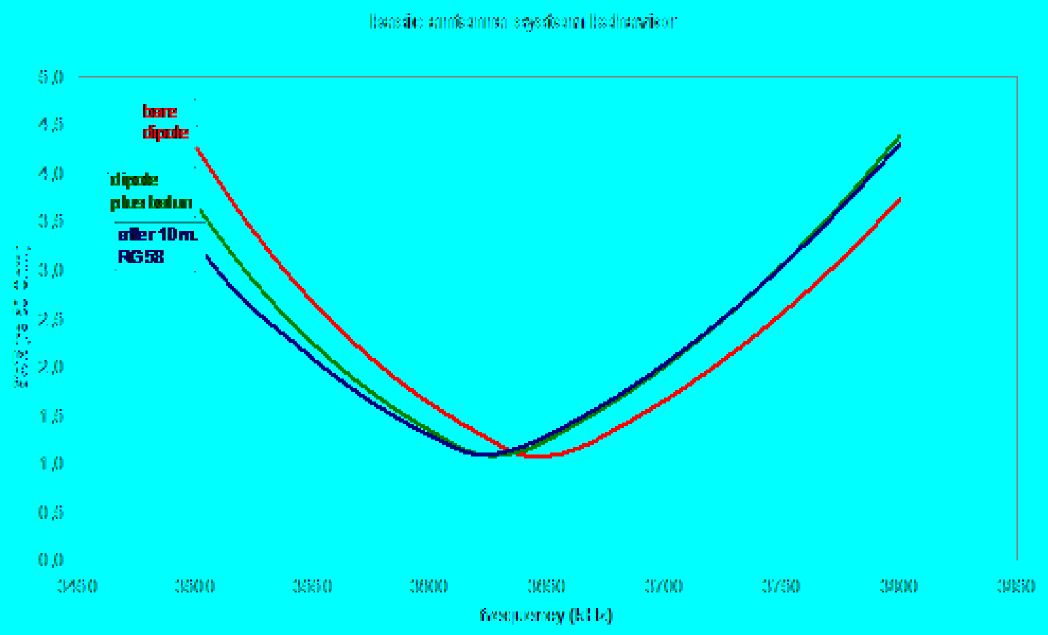

Figure 1: Behavior of the basic antenna system

In figure 1 we find the red curve representing the bare antenna as a result of the antenna design program. This antenna is behaving as required, resonating at 3,65 MHz. and showing a band-width of 140 kHz. between SWR=2 points.
When connecting the balun (green curve), the compound system has been shifted by 20 kHz. to the lower frequency side. Connecting the transmission line (blue curve), the curve is somewhat shifted again and has increased bandwidth to 145 kHz. between SWR = 2 points. When designing the antenna at one particular operating frequency, we will have to take these small frequency shifts into account.

In our design example we also connected an antenna tuner, so the 140 kHz. usable band-with between SWR = 2 positions is not relevant anymore since we are capable of matching the antenna across the entire 80 m. amateur band. To allow for this situation, the L-tuner will have to meet conditions as in table 2 (at 100 W. input power).

L-series antenna tuner

f kHz	**L** •H	**C_{par}** pF	**tuner loss** Watt
3500	1,4	1341*	1,1
3520	1,4	1176*	0,9

3540	1,3	1008*	0,8
3560	1,2	838*	0,6
3580	1,1	662*	0,5
3600	0,96	472*	0,3
3620	0,69	238*	0,2
3640	0,74	158	0,2
3660	1,22	201	0,3
3680	1,56	175	0,4
3700	1,85	127	0,5
3720	2,1	70,2	0,5
3740	2,33	7,7	0,5
3760	3,2	357*	0,9
3780	3,42	535*	1,2
3800	3,47	689*	1,5

*) C at transceiver side

Table 2: Tuner for the 80 meter amateur band at the example antenna system

In table 2 we see this tuner may be constructed with run-of-the-mill components, except maybe for the high capacitor values below 3560 kHz. although this may be accomplished by adding fixed capacitors.
Note: The '*' sign indicates the capacitor the be connected at the transceiver side; otherwise the capacitor is connected at the line-side of the tuner.

Again we find some power left behind in the tuner. This power is in a first approximation related to inductor loss (Q-value) and therefore it pays to apply high quality parts. With all components known, we may put all loss in perspective in figure 2.

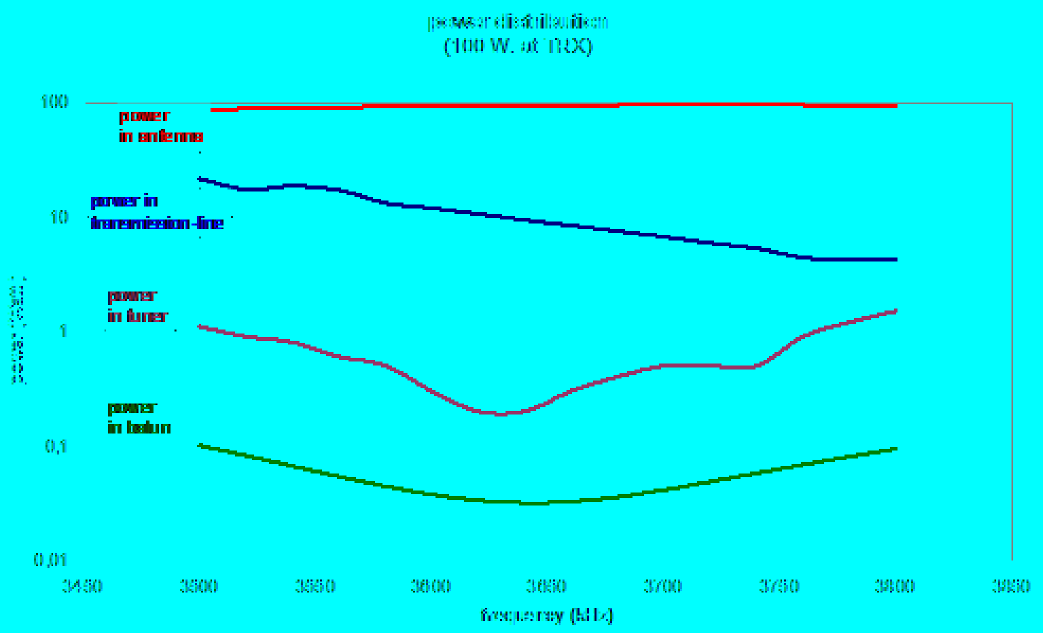

Figure 2: Power distribution in the example antenna system

We are happy to find most of the input power is radiated at the antenna and only little power is dissipated in other parts of the system. It is remarkable that despite low-loss transmission-line (2,76 dB / 100 m. @ 3,5 MHz.) this still is consuming most of the system loss-power. Figure 2 is also showing we designed a well behaved antenna system with high overall efficiency.

Same system, other amateur frequencies

Since we are satisfied with our antenna system at the 80 m. amateur frequency band, and a tuner is available to protect the transceiver from adverse operating conditions, we may be curious as to the behavior of this system at other amateur frequencies.
With this statement we are approaching the motive for this article. Many amateurs in Europe are applying their antenna system that was designed for application at one amateur band also to other frequencies, since a tuner is available and the system will show low SWR at the transceiver after tuning to new frequencies. A system that is designed around a particular impedance (range) will behave quite differently when operating in a non-matched situation as we may find in table 3.

bare dipole			**dipole plus balun**			**after 10 m. RG58**			**tuner**		**loss @**
	SWR	**radiated**		**SWR**	**balun**		**SWR**	**cable**		**tuner**	

f	**r**	**X**	**re 50**	**power**	**r**	**X**	**re 50**	**loss**	**r**	**X**	**re 50**	**loss**	**L**	**C_{par}**	**loss**	**receiver**
(kHz)	Ohm	Ohm	Ohm	Watt	Ohm	Ohm	Ohm	Watt	Ohm	Ohm	Ohm	Watt	•H	pF	Watt	S-punt
3650	51,4	0	1	94,3	49	10	1	0,0	57,6	-11	1	5,5	0,96	188	0,2	0,0
7050	5530	911	114	13,3	41	466	108	0,9	4,5	31	15	84	2,08	898	1,8	1,5

Table 3. Non matched system

Table 3 is showing behavior and loss figures for the well designed, 3,6 MHz. antenna system (red number), now operating at the 40 m. radio-amateur band (black numbers). The dipole system is now showing a high complex impedance which may be understood with the antenna at a dimension of about one wavelength for this frequency. Because of this high input impedance, SWR is high when related to the original system impedance at 50 Ohm.
Connecting the balun in parallel to the antenna, will bring the real-part of total impedance down but will leave SWR at a very high value since the imaginary part is still quite high. Further more the balun impedance (Z = 477 Ohm) will be too low compared to the antenna impedance (Z = 5600 Ohm) to perform its balancing function. The balun was not designed for this high impedance and will now consume just under 1 Watt, as compared to almost 0 Watt at 80 m.   This fortunately is still lower than maximum allowed dissipation for this component at 4 Watt (see balun calculations).

The transmission-line that is now totally unmatched, will transform the high input impedance to a (very) low value and to our enjoy we find SWR is down considerably. This however is related to very high cable loss with 84 Watt of available HF power (100 W.) to be turned into heat along this line. At the tuner side we hardly will notice anything unusual as the transceiver will show SWR = 1 when the tuner is set to: L = 2,08 •H. and C = 898 pF. Tuner loss is not unreasonable either with only 1,8 Watt burned internally.

After all system loss along the way, the antenna will radiate only 13,3 Watt of HF power since this un-matched antenna situation is turning 86,7 % of available power into heat. Our communication partner at the other end will notice signal strength at the S-meter to be down by over 1,5 S-point as compared to our neighbor with a well designed, low-loss station. Relative S points may be defined at: $(10 \log(P_{max}/P_{actueel}))/6$. The diminished signal report may be regarded as acceptable (some signal is better than not being able to operate at all), but any signal loss will be too much at adverse communication conditions like high QRM or at a DX pile-up.

The high loss figures at 40 m. are in contrast with the figures for 80 m. in the same table. It is immediately obvious that high loss may be encountered when operating an antenna system at a frequency this was not designed for. When looking at the behavior of our 80 m. antenna system at other HF amateur frequencies, figure 3 may be calculated.

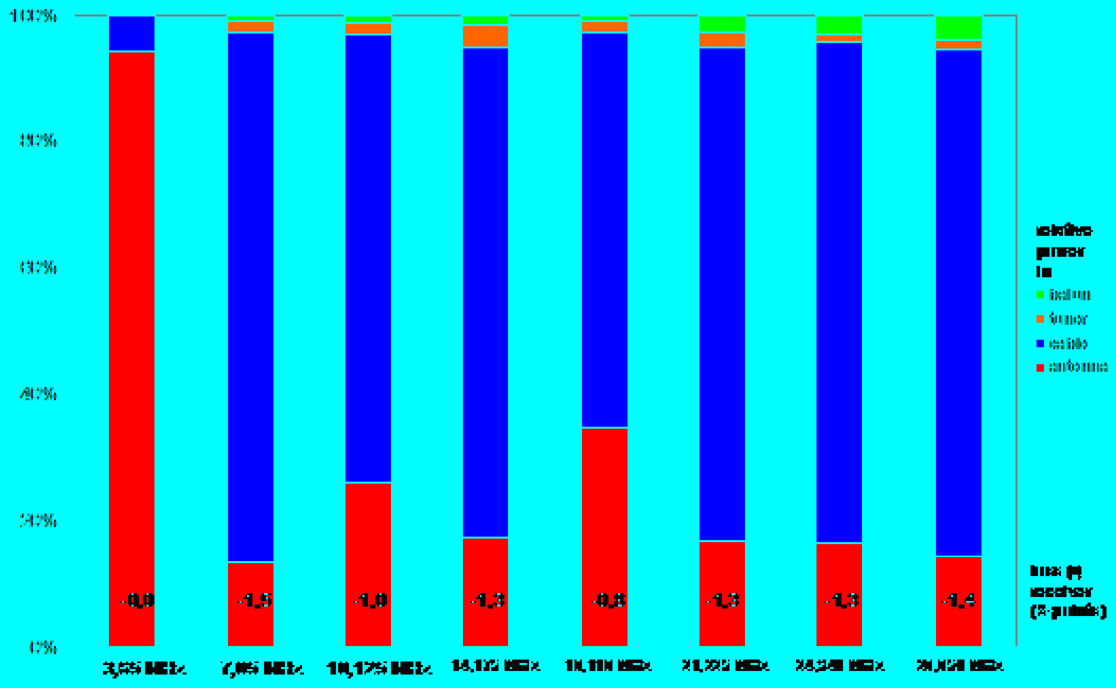

Figure 3. Relative power into example dipole at HF frequencies

Figure 3 is telling us the dipole antenna system designed for 3,65 MHz. is doing very well at this frequency with most of the input power going into the antenna (red). At all other HF amateur bands, SWR is high with low antenna power as a consequence. At all but 3,65 MHz. most of available power is lost in the feed-line (blue), with total loss of one S-point or more at the receiving station.

Balun loss (green), limited to 4 Watt maximum, is within safe limits in this example antenna system because of high cable loss.

Cable loss

Why high cable loss?
It may be difficult to imagine why this relative short stretch of feed-line is generating these high loss figures, since cable specifications show 2,76 dB / 100 m. (in a characteristically terminated cable) and in our example antenna system we find a loss of many dB's over a mere length of 10 m., with over 80 % of power transformed into heat?

Total cable loss is somewhat mystified when in cable transfer function but we may gain some insight from the models (just for insight) in figure 4 and 5.

a b

Figure 4: Transmission-line with reflected power

In figure 4a we find a transmission-line that terminated into its characteristic impedance. All energy from the generator is transported to the terminator and only very mildly absorbed by the line. No energy is reflected back.

In figure 4b we find the same situation, this time (very) uncharacteristically terminated. Part of the energy will be reflected at the load and will travel back to the generator, as in:

$\rho = E_r / E_f$

with
ρ (rho) = reflection coefficient (complex)
E_f = forward energy

E_r = reflected energy

With the bare generator being a voltage source, the transmission-line again is terminated uncharacteristically and again power will be reflected. Note. The generator in our example system is our transceiver. This transceiver is designed to deliver full power when terminated into a 50 Ohm resistive load and is protected to cut back on system power when terminated into a different load to safeguard the active elements in the output. The specified termination resistance however is not related in any way to the 'output impedance' of this transceiver. An article discussing this situation in more details may be found here.

We may be more familiar with the 'SWR' terminology and of course 'SWR' and 'ρ' are related, as in:

$|\rho| = (SWR - 1) / (SWR + 1)$ (vertical lines to denote absolute value)

With SWR = 10, ρ = 0,82 meaning that 82 % of power will be reflected. In above tables we found much higher SWR not to be unlikely. The energy will be reflected back and forward and this will go on until finally all energy has been burned. The transmission-line will therefore be passed a number of times, each time consuming a little energy. The larger the mismatch at the terminator, the larger the portion of energy that will travel up and down and the more times the transmission-line will be traversed (model!).

An other model to visualize the loss mechanisms in un-characteristically terminated transmission-lines, may be found in the SWR definition, as in figure 5.

a b

Figure 5: More loss by higher voltage and current

In figure 5a we again find a generator, a transmission-line and an end-user to terminate the line. The transmission-line has been terminated characteristically so noting is reflected and the voltage and current on the line is constant and equal at the generator and the load (minus a little resistive loss in the cable).

In figure 5b the line is un-characteristically terminated and part of the voltage and current will be reflected. This reflected signal will add to (or subtracted from) the incoming signal so a fixed wave pattern will be building up at the transmission-line. The peak and valleys of this voltage may be directly measured at the cable as the (Voltage) Standing Wave Ratio. This means the voltage at some position at the transmission-line may be much higher than originally envisaged, generating much higher de-electric loss than in situation 'a'. In particular situations this locally higher voltage may even generate a flash-over at below maximum (power) ratings for a well terminated transmission-line.
The identical situation also applies to currents along the transmission line, with local currents to be much higher than average. This is not just a model but will indeed be the practical situation making the transmission-line prone to local hot-spots and even melting of the central conductor at a system power below maximum rating of the well-terminated situation.

What cable loss is acceptable?
To get an idea about what SWR to allow, we could say to be not very concerned up to a level where the receiving station is receiving us with a signal strength of 0,5 S-points below optimum. This situation will arise when half of the available transmitter power is 'lost' somewhere along the antenna system. Next we calculate maximum SWR at each frequency for RG58 cable (2,76 dB/ 100 m. @ 3,5 MHz. when terminated characteristically) at different line length to just loose this amount of power.
Note these SWR figures to be measured directly at the reflecting side (at the antenna). At the input side SWR will be much lower at half or less the value at the antenna. We may therefore easily under-estimate the bad situation.

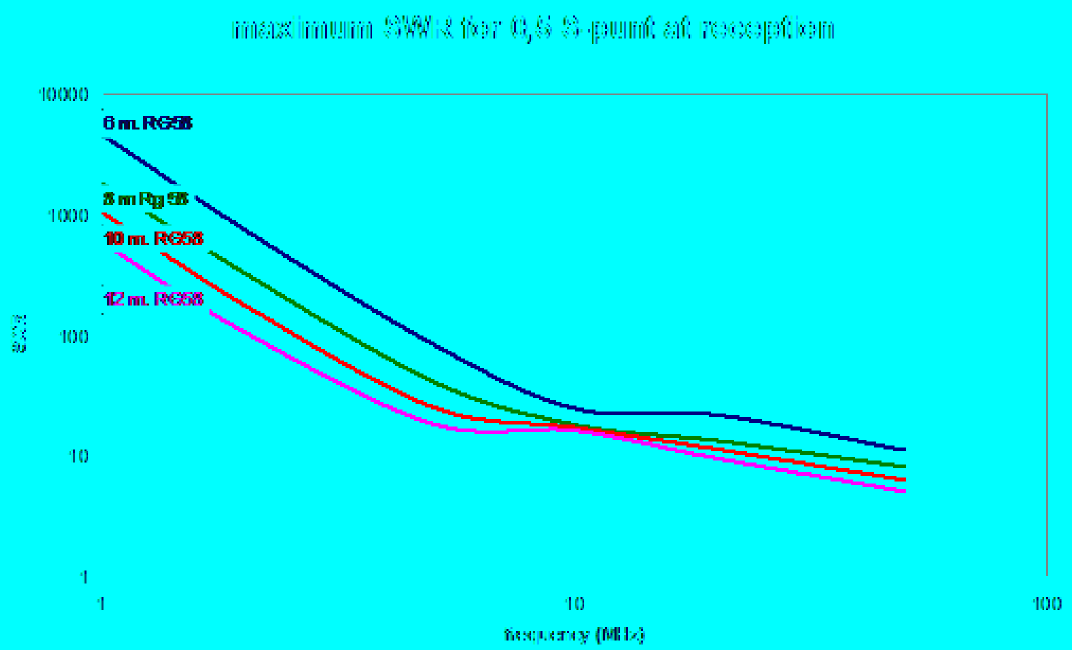

Figure 6: Maximum allowable SWR for 0,5 S-point loss at receiver

In figure 6 and when applying RG58 type of transmission-line, we may allow quite high SWR before cable loss will become excessive (SWR > 50) at frequencies below 3 MHz. Between 3 and 20 MHz. we better ensure SWR < 10 and above 20 MHz. we better be very careful with any type of mismatch as cable loss quickly leads to unacceptable loss of radiated power.

Better transmission-line

In the previous section we found transmission-line loss to be 'amplified' when SWR is high because of non-characteristic line termination. This may lead us to lower transmission-line loss as a way-out. To test this idea I changed the RG58 for lower-loss RG213 (Belden 8267) exhibiting 1,194 dB / 100 m. at 3,5 MHz., less than half the RG58 loss. Again the tuner ensures a perfect match to the transceiver, that will deliver 100 W. into the system. Total power-picture may be found in table 4.

f	**tuner loss**	**cable loss**	**balun loss**	**antenna power**	**loss@ receiver**
(kHz)	**Watt**	**Watt**	**Watt**	**Watt**	**S-point**
3650	0,2	2,4	0,0	97,4	0,0

7050	3	69,4	1,7	25,9	1,0
10125	2,5	50,7	2,3	44,5	0,6
14175	5,4	59,4	3,1	32,1	0,8
18118	2,1	40,9	1,8	55,2	0,4
21225	4,5	59,7	5,0	30,8	0,9
24940	1,4	61,1	6,1	31,4	0,8
28850	2,4	62,3	7,7	27,6	0,9

Table 4: Power distribution when applying 10 m. of RG213 transmission-line

When comparing table 4 to figure 3, we find this better transmission-line is doing some good. Since SWR did not change, we still loose quite some power that will manifest itself as a loss of around 1 S-point at the site of the receiver of our signals when compared to an optimal situation. In table 4 it is also clear we have to look again to our balun, that is loaded above its rated maximum power of 4 W. at several amateur frequencies. At least a double amount of ferrite is to be applied.

Better matching

Although we have been applying a better type of transmission-line, cable loss still is high because of high cable mismatch. As a side step it may be useful to look at a system that is designed to deliberately avoid (high) mismatch at a number of amateur frequencies. This antenna may be found at "Multiband trap antenna" and is consisting of a dipole at 2 x 12,6 m., a trap and again 2 x 5,3 m. of wire. To calculate figures, Q = 150 at the trap inductor as a realistic (loss) number for outside situations.
In this design, the antenna is coupled through a 1 : 2,25 transformer to ensure SWR < 4 throughout amateur HF frequency bands of 80, 40, 20, 15 and 10 m. Only requirement to the feed-line is the characteristic impedance of 50 Ohm, so I calculated loss figures for 10 m. transmission-line at RG58 for this comparison.

For the analysis in table 5, I set the balun to be a 1 : 1 type as in the earlier examples for a more direct view to the influence of the well behaved antenna.

f (kHz)	**tuner loss** Watt	**SWR re 50** Ohm	**cable loss** Watt	**SWR re 50** Ohm	**balun loss** Watt	**trap loss** Watt	**antenna power** Watt	**loss@ receiver** S-punt
3650	0,4	1,7	9	1,8	0,04	0,8	89,8	0,1
7050	0,1	1,6	11,1	1,7	0,09	17,1	71,6	0,2
10125	2,1	14,1	79	72,7	1,48	0,4	17,0	1,3

14175	1,2	3,1	21,9	3,8	0,42	2,5	74,0	0,2
18118	1,9	9,5	82,1	57,6	1,52	0,2	14,2	1,4
21225	0,6	3,0	26,8	3,8	0,81	0,9	70,9	0,2
24940	1,4	8,3	82	48,9	3,40	0,0	13,2	1,5
28850	0,9	3,0	33,9	4,2	1,49	0,0	63,7	0,3

Table 5. Power distribution in 'multiband trap antenna'.

In table 5 we find the system to indeed show a high efficiency at the design frequencies (red numbers) and low loss figures at the receiving side of a few tenth of an S-point. Applying the design transformer of 1 : 2,25 instead of the 1 : 1 balun, loss will even be less.

As expected efficiency is low at the non-design HF-bands, 10, 17 and 12 meters, since SWR is high with a high cable loss as a consequence. At these frequencies, receiving reports will be less by about the amount at the earlier antenna systems.

It is interesting to look what is happening at high loss situations: at 10,125 MHz. we find at the tuner SWR = 14,1 while at the antenna this is 72,7! Something analogue may be discovered at the other non-design bands: 18,118 MHz.: SWR_{in} is 9,5 with SWR_{out} at 57,6 and at 24,940 MHz.: SWR_{in} is 8,3 with SWR_{out} is 48,9. The lower SWR at the tuner position is no guarantee the system is operating at high efficiency!

Although a transformer is to be applied, the balun will also do an acceptable job with no particular requirements other than specified earlier, since power dissipation is always below 4 Watt, even at the non-design bands.

To my surprise some power is lost in the trap inductors in spite of the relatively high Q, set at 150. This is indicating high Q to be a prerequisite for an efficient (trap) antenna. This also is an indication to be careful when applying 'coaxial traps' to an antenna since these components are exhibiting a relatively low Q-factor because of distributed trap capacitance.

Still better transmission-line

In the early radio-days, coaxial transmission-lines where not yet readily available so antennas where connected through single wire or symmetrical lines. This latter type of line is not very helpful at lowering high SWR but line-loss of well maintained lines may be very low indeed at 0,104 dB / 100 m. at 3,5 MHz. when characteristically terminated.
To find-out what this type of line may bring, I calculated the same antenna as in our first system, this time connected through 10 m. of 600 Ohm symmetrical line directly connected to a symmetrical tuner, again constructed with Q = 200 inductors. Results may be found in table 6.

f (kHz)	**cable loss** Watt	**tuner loss** Watt	**antenna power** Watt	**loss@ receiver** S-point

3650	1,5	5,6	92,9	0,1
7050	1,5	6	92,5	0,1
10125	2,2	4,4	93,4	0,0
14175	1,5	4,7	93,8	0,0
18118	1,7	5,3	93	0,1
21225	1,6	4,5	93,9	0,0
24940	2,4	4,9	92,7	0,1
28850	1,7	4	94,3	0,0

Table 6. Power distribution in dipole antenna designed for 80 m. with symmetrical feed-line (600 Ohm) and tuner

In this antenna system SWR is still high, since the antenna is the original 80 m. dipole with low connection impedance at the 80 m. amateur band but high and wildly fluctuation impedances at all other radio-amateur frequencies. In spite if the high SWR, system loss is very low because of the very low transmission-line loss.

Note 1. This calculation is based on system loss in the contributing parts as mentioned. The low transmission-line loss however is valid for a well kept line that connects the antenna to the tuner in a straight line. This may be somewhat difficult in practice since the transmission-line may be strengthened mechanically by special wall-fixes, through-wall connection system and in-house suspension until connecting to the tuner. The electro-magnetic field in this type of transmission-line however is protruding a little beyond the line so everything supporting this will somehow influence field transmission and will induce some additional loss.

Furthermore, nature will also have its way with an open line. Garden debris like fallen leafs, branches etc may become stuck between the two conductors and the wall, as may be insect- / bird-nests, algae, moss etc. Also weather conditions like rain and snow will have some influence making this low-loss line not so very low loss anymore without regular maintenance. This does not apply to coaxial transmission-line that is more a set-and-forget type of material with considerably lower maintenance requirements, especially when 'vulcanizing' connectors.

Note 2. A well constructed, symmetrical antenna tuner is no simple and / or cheap device. Impedances may be very high, so keeping out environmental effects is not simple (hand-effect) as is maintaining symmetry. Also very high voltages may be found at high impedance, even at relatively low system power: at 100 Watt and a few 1000 Ohm reactive, voltage may rise to over 1 kV. at the tuning capacitor. Even when good quality symmetrical tuners are offered at ham-fests, this will be at a comparatively high price as compared to a-symmetrical tuners.

Compromising on system components

As a last exercise I modeled a system with the original 80 m. dipole connected to 10 m. open feed-line to the shack, than a balun before connecting to a 'standard' a-symmetrical antenna tuner. Results may be found in table 7.

dipole antenna for 80 m.	**impedancelevels**		**after feed-line and special balun**	**system loss**	**output**	
	cable	balun			pwr in	loss@

f	r	X	SWR	Z	Z	r	X	SWR	cable	balun	tuner	antenne	receiver
(kHz)	Ohm	Ohm	re 600	Ohm	Ohm	Ohm	Ohm	re 600	Watt	Watt	Watt	Watt	S-punt
3650	51,4	0	11,7	657	493	21,6	282	34,0	1,5	1,1	6,2	91,2	0,1
7050	5530	911	9,5	65	953	62,7	14,4	9,6	1,5	0,1	0,4	98,1	0,0
10125	119	-516	8,9	5153	1424	405	1351	9,6	2,2	5,0	6,1	86,7	0,1
14175	1828	1656	5,7	3078	2301	754	1357	6,0	1,5	4,4	4,8	89,3	0,1
18118	108	-308	7,1	339	3080	93	292	7,9	1,7	1,6	2,9	93,8	0,0
21225	908	1299	5,1	166	3647	132	-109	4,7	1,6	0,6	1	96,8	0,0
24940	190	-639	6,9	3045	4248	3393	235	5,7	2,4	13,6	4,8	79,2	0,2
28850	949	1209	4,6	2677	4898	1797	987	4,0	1,7	12,6	4	81,7	0,1

Table 7. Power distribution in dipole for 80 m. with symmetrical feed-line (600 Ohm), balun and tuner

At the dipole section we find SWR re 600 Ohm to be quite high as in the section before. After 10 m. of 600 Ohm symmetrical feed-line and connecting a double cored balun to allow for more power-loss at the core, SWR has changed to a higher value at 3,6 MHz., and about the same for all other amateur bands. System loss in each constituent is low again except for the two highest band, where the balun could use even some more ferrite to handle internal loss. This will also help raising balun impedance at 40, 30 and 20 meter, that is still too low with 8 turns at 2 x 36 mm. toroide to perform a good balancing function at the high impedances as found at the end of the transmission line (compare 5th and 6th column).
This is also showing we should be very careful when connecting a 'random' balun at the end of an a-symmetrical tuner to connect to a symmetrical feed-line. Usually the balun impedance is much too low to perform adequate balancing at the high impedance levels as found at this position after high antenna impedances have being transformed by high-impedance feed-lines. Therefore it is quite remarkable one sometimes may come across a carbonyl (low permeability material) balun build into an a-symmetrical tuner to allow for a symmetrical tuner output. It is quite puzzling what type of applications the manufacturer may have had in mind to this extend.

Taking above considerations into account, this last system set-up may be the system of choice to obtain a high efficient antenna system for all HF amateur frequencies, starting at 3,5 MHz. Even the a-symmetric tuner is quite unremarkable with series inductor 18 •H > L > 0,6 •H and parallel capacitor 280 pF < C < 15 pF, which is not extreme at these frequencies. Keep in mind that again we are dealing with high impedances at the tuner with high(er) voltages across system components than usual.
All in all we have to compare this open-line fed system with the 2 x 19,75 m. dipole antenna to the coaxial fed multiband trap antenna at 2 x 17,6 m. In both cases the multiband antenna system will be highly efficient.

Conclusions

At the end of our little discussions we may come to a few basis conclusions:

Coaxial feed-line: For high efficient antenna systems coaxial feed-lines may be applied with resonant, low impedance antenna's. This will yield low SWR at the transmission-line connecting the antenna and the transceiver (tuner). To this extend, 50 Ohm transmission-lines are no better than 75 Ohm cables and may be applied whichever is more convenient. Examples of resonant antenna systems are: monoband dipoles or monoband verticals, cluster of monoband dipoles at a single balun (cat-whisker antenna) and a multiband trap dipole.
Pro coax: easy handling, leading indoors, along walls, along metal drainage pipes etc, usually UV resistant and simple baluns.

Con coax: see pro 'balanced line'

Balanced feed-lines: For high efficient 'random impedance', usually non-resonant antenna systems. Although SWR will still be high, very-low line loss will not 'amplify' cable loss to un-acceptable values.
Pro balanced feed-lines: low loss, simple home-made construction.
Con balanced lines: see pro-coax and also high voltages at feed-line, balun and tuner, already at relatively low power.

In general **when designing multi-band,** complexity of the antenna system (more tuned antenna's, constructing traps) is to be traded against complexity of the feed system (handling balanced wires) and tuners (symmetrical tuners, complicated balun)

Whatever your design, it is always important to calculated the entire antenna system at all operational frequencies to selected adequate components in order to avoid unpleasant surprises like loss of power (bad report) , flash-over (tuner, transmission-line) or burned components (balun).

Bob J. van Donselaar
mailto:on9cvd@amsat.org

Wire Balun

(published in Electron #4, 2007)

General

This chapter is the second in a series of articles on baluns for antenna applications. The first article is an introduction with some background on balun types. It is advisable to read the articles in the above order especially since each next chapter is building on information and formula's already explained earlier and referencing to this.

Introduction

At many radio-amateur sites one may observe the coaxial feed-cable to be tied up into a coil with a few number of turns next to the antenna. To the casual observes this may look like some spare wire, to be used later-on when constructing the antenna at a higher position. At closer examination however this 'spare wire' is a carefully constructed inductor to keep antenna current away from the (outside of the) feed-line. The construction is hereby acting as a sleeve choke or 1 : 1 current balun.

Calculation

To obtain some feeling for numbers we will calculate this current balun for a system impedance of 50 Ohm, that may be found in a tuned dipole antenna, depending on antenna height and ground type. From the introductory article we already know the sleeve impedance to be at least four times system impedance to be effective at 'adverse' operating conditions. This impedance should be calculated at the lowest of operating frequencies since impedance will go up with frequency. At a lower frequency of 2 MHz. and a sleeve impedance of 4 x 50 Ohm = 200 Ohm, we calculate:

$L = X_L / \omega = 200/ (2.\pi.\ 2.10^6) = 16\ \mu H$

Note: The sleeve impedance has been calculated for a system impedance of 50 Ohm, to accomodate the example dipole and the antenna feed-line. Outside resonance, the antenna impedance will be much higher, so also the sleeve impedance should, to still be effective. In a practical situation therefore, sleeve impedance should be calculated at all operating frequencies together with the antenna impedance. At the most demanding frequency (highest antenna impedance) the sleeve impedance should still be at least four times the antenna impedance to be effective. We should be aware of this situation whenever operating an antenna outside resonance.

General formula for calculating an 'air' inductance is:
$L = n^2 \,.\, \bullet \,.\, Q / l$,   (H)

with 'n' for the number of turns, 'Q' for the area of one turn and 'l' for the inductor length. This (fundamental) formula is accurate for all inductors with a length to diameter ratio of three or more. A bunched coil as in the antenna situation does not comply to this ratio, so has to be approached in a more practical way. In the ARRL handbooks it is advised for the frequency range 3 - 30 MHz. to construct this sleeve choke with 6 - 10 turns of transmission-line, with a diameter of 10 - 15 cm. For a quick test I constructed this wire balun from 6 turns of RG58 coax with a diameter of 12 cm. as depicted in figure 1.

Figure 1. Wire balun for 3 - 30 MHz. according to ARRL.

Measurements to the 6 - turn wire balun

Transmission
The balun as in figure 1 has been tested for current balancing properties by measuring transmission and reflection when connected into 50 Ohm. For each measurement, the balun was to operate at maximum unbalance, i.e. with the center

conductor grounded. Transfer qualities have been depicted in figure 2.

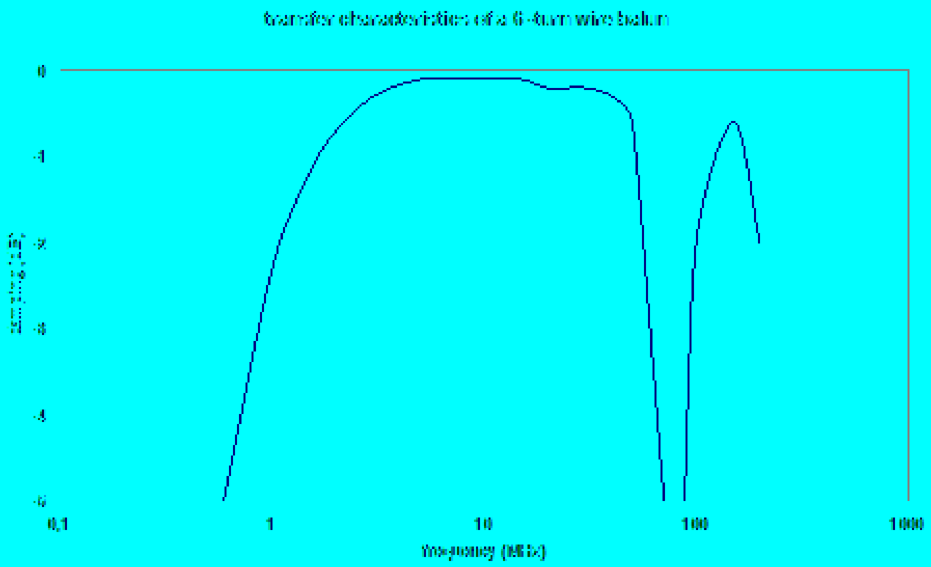

Figure 2. Wire balun transfer characteristic

Figure 2 depicts the transfer characteristic (insertion loss) of the wire balun of figure 1.
Insertion loss is lower than 1 dB starting at 2 MHz. and stays that way to over 50 MHz. Low frequency behavior is dictated by the coil inductance, that effectively is measured in parallel to the input. At the high frequency side, the parasitic parallel capacity is limiting sleeve impedance. This parasitic capacitance is an incidental value, that may easily be higher or lower depending on the way the coil is 'bunched up'. The returning of the curve above 100 MHz. is related to this particular sleeve inductor only and will be different (value and position) for a different 'bunch'.

This transfer function is quite acceptable although one may be tempted to enhance performance at the low frequency side by increasing the number of turns. When increasing, the parasitic capacitance will also increase, lowering the cut-off frequency at the high frequency side. Although some room may for play is available, margins are small.

Reflection
Equally important is the behavior of this balun as seen from the input position. We therefore measured input reflection which is depicted in figure 3 as a SWR graph.

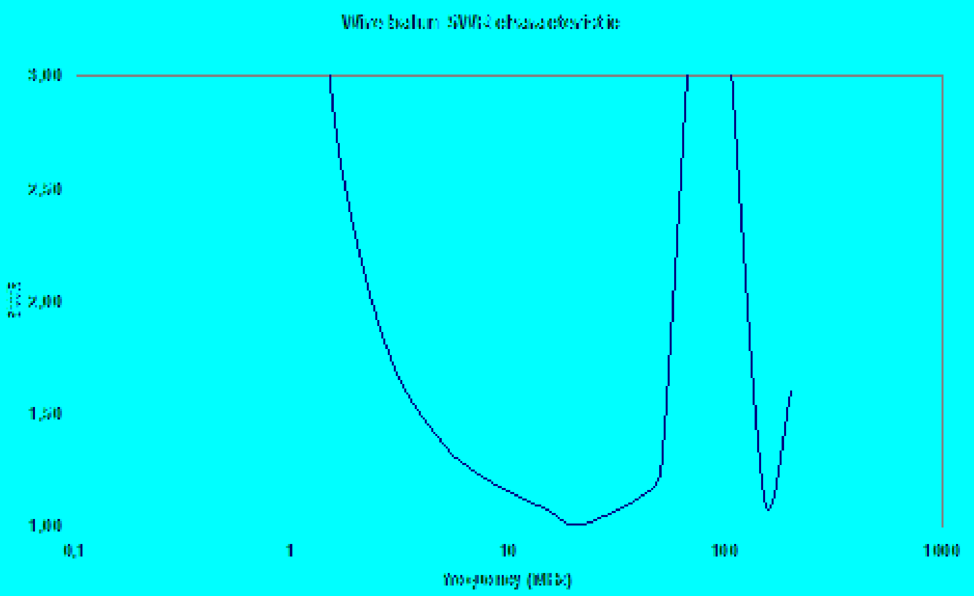

Figure 3. Wire balun reflection characteristic

In figure 3 reflection is measured for the wire balun of figure 1, when terminated into 50 Ohm and in a maximum unbalancing situation. We find SWR below 1,5 between 4 - 50 MHz. which is acceptable for most HF applications. Comparing figure 3 with figure 2, we find SWR to go up where transmission is going down, at the low frequency side as well as at the high frequencies. Also the 'return' of the curve around 150 MHz. is visible again.

System power
Taking both graphs together this certainly is an acceptable component that will serve its purpose in a 50 Ohm system environment for a wide number of HF amateur bands. This is even more so when looking at the maximum allowable system power; at this component entirely determined by the characteristics of the transmission-line. When applying RG 58 :
$V_{max.}$ = 1900 V. (600 V. in the foamed RG58 variation)

$I_{max.}$ = 2,6 A

According to the manufacturer of 'standard' RG58 cable, this type may be operated up to 350 Watt at 30 MHz., to de-rate with increasing frequency

More on wire baluns

Around the end of last century Steve Steltzer, WF3T, has performed a series of measurements using an automated test bench with a HP vector voltmeter. He started-of by constructing his wire baluns on drainage pipes with a diameter of 4 1/4 " (10,8 cm.) and 6 5/8 " (16,8 cm). Each time a series of measurements has been performed with different number of turns of RG58 transmission-line. The results may be found in figure 4.

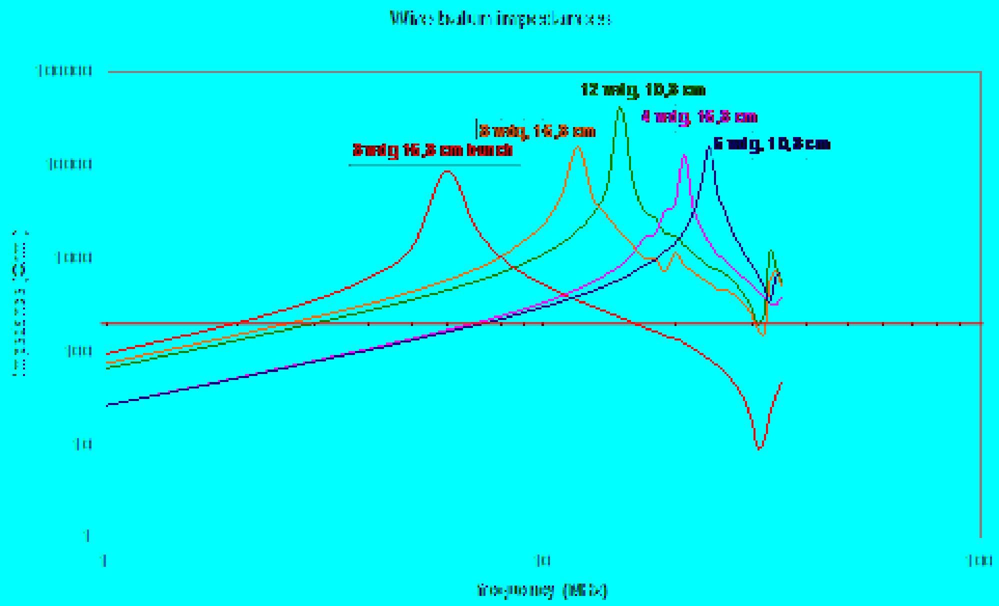

Figure 4. Wire balun impedances

In figure 4 all baluns have been constructed at the drainage pipes of the indicated diameter except the graph marked 'bunch'. This last balun is the balun on 16,8 cm drainage pipe, of which the pipe has been carefully removed and the turns been tied together in a way as depicted in figure 1.

As we have noticed before, the balun should have an impedance that is at least four times the system impedance. In figure 4 this is depicted by the red line at 200 Ohm, implicitly comparing baluns for application in a 50 Ohm system. Taking this line it is clear only 8 wdg or more will qualify with the tested diameters for frequencies starting at 3 MHz.

Next interesting part is the resonance peak for all of the constructions. At these frequencies the inductor is resonating with the parasitic parallel capacitance making the choke impedance really high indeed. Taking the same diameter, but more turns, total impedance at the low frequency side will go up with resonance frequency to go down, making this higher-turns choke even less interesting at the high frequency side (compare blue and green curve).

After resonance, the impedance will still be high for some time, later to drop off sharply when this parasitic capacitance is the dominant circuit reactance. The only chokes to still qualify at 30 MHz. are those that do not qualify at 3 MHz. for too low impedance (lower number of turns).

The 'bunch' graph is showing nicely what is happening when windings are close together. When comparing the red and orange curves, it is clear the bunch type is showing higher impedance at the low frequency side since inter-winding coupling is much better with close-winding. The parasitic capacitance is higher because in the bunched-up situation first and last winding, carrying highest potential difference, are much closer together now although the exact extent of the phenomenon is highly dependant on the accidental way of constructing the bunch. On the other hand, this effect may be applied to good use when a choke balun should exhibit a particularly high impedance at a desired frequency; the bunch may be 'fondled' until the desired effect has been reached.
Because of the higher parallel capacitance, parallel resonance is occurring at a lower frequency, making this choke fall-of earlier at high frequencies too. Steve notes this effect to also showing at the other choke baluns.

Some conclusions

From the above series of measurements it may be concluded wire baluns are useful over a number of HF frequencies and at the specific resonance frequency even very good. In general we may safely notice that wire baluns may be applied over a frequency range of about 1 : 3, but it will be hard to construct a wire balun that will effectively operate over the entire 3 - 30 MHz. field of frequencies, let alone 1,8 - 30 MHz.
Fortunately other solutions exist to solve this problem and may even be applied over a much higher bandwidth as may be seen in the next chapter.

Please contact me for your remarks and questions at:

Bob J. van Donselaar
mailto:on9cvd@amsat.org

aspects of wide-band, linear HF power amplifiers

Introduction

When designing wide-band, linear HF power amplifiers a small number of basic rules should be considered. These are not very complicated but should be followed rather carefully when the object is to maximize power and efficiency. With these basic rules firmly into position, further details will follow quite naturally.
These basic design rules apply to LF and HF amplifiers although each of these fields also exhibit additional requirements that are specific to the particular range of frequencies. With basic design rules comparable, still many radio-amateurs have no problems designing LF amplifiers and will shy away from their home-brow linear HF power equivalents. The short notes in this article will discuss these basic design rules for radio-hams to start thinking about designing their own amplifiers from available components, probably already present in the junk box.

System impedance

In tuned, single frequency amplifiers the choice of the system impedance is less important as most internal parasitic impedances will be part of a designed-in impedance that will be tuned to the desired operating frequency.
In wide-band systems, system impedance Z_0 is one of the primary selections. This impedance in principle is a freely selectable quantity but somewhat restrained in practice.

As a general rule, the higher Z_0, the lower system currents and therefore the lower the loss in all parasitic series inductances, resistances and further 'additional', series impedances. This premium is lost however in all parasitic (parallel) capacitances that will be unavoidable and will limit maximum frequency.

At a low system impedance, system currents will be high(er) and therefore all series-loss will increase. Because of the low impedance, parallel capacitance will be less important allowing for higher maximum frequencies.

From the above it follows system impedance selection to be connected to the desired frequency and frequency range of the amplifier, with the lower impedance range related to the higher operational frequencies. Figure 1 is presenting a general idea about the position of the system impedance, that does not have to be represented by an actual component and in practice usually isn't either.

Figure 1: System impedance

System impedance also is an entity of choice and therefore be closely related to practical considerations e.g. availability and price of components designed for a particular characteristic impedance. To this extend one should think of transmission-lines, filters, sub-systems etc., but also that high-impedance inductors may be constructed from lower diameter, un-silvered wire and low-voltage capacitors usually also mean lower price. Although a system impedance of 50 Ohm for wide-band systems is a well known 'standard', 75 Ohm is a very good option and other impedances may be even better depending on local requirements.

At wide-band power amplifiers based on vacuum tubes it may be profitable to select a higher system impedance. A vacuum tube usually operates under a high-voltage, low(er) current 'regime' making matching impedances also of a higher level. Further more, (dipole) antennas only exhibit a rather low termination impedance (around 50 Ohm) when resonating and a much higher impedance outside resonance. Selecting a high 'system impedance' when matching an antenna to a tube amplifier therefore usually will require a lower transformer ratio which in general is more profitable from an efficiency point of view and also simpler to construct.

At wide-band power amplifiers based on semi-conductors a low impedance system impedance may be more profitable since transistors usually operate in a low-voltage high current regime. To match to a high(er) impedance antenna, an in-between step to an 'intermediate' system impedance usually is a practical solution to avoid high-step-up ratio's.

Matching to system impedance

At power amplifiers a wide range of voltage-current regimes (impedance-level) may be encountered, amongst others depending of the type of active component. From a system point of view it is efficient to arrive at system-impedance level as soon as this is practical and to this extend a matching transformer usually is the component of choice. Parasitic effects are always a point of concern in wide band amplifiers. Therefore impedance matching ratio's should not be too high to avoid parasitic effects of the low impedance regime to influence the circuit at the high-impedance site and vise versa. In this respect one could think of (serial, leak) inductance as a factor of concern in the low-impedance regime and parallel and inter-winding capacitive effects at the higher impedance site. In practice and for high(er frequency wide-band amplifiers in particular, impedance ratio's of 1 : 16 already are very high, while 1 : 25 for LF amplifiers are run of the mill and very low when using vacuum tubes.

What parallel impedance?
A matching transformer is to match the lower impedance regime to the higher environment, preferably without being a 'matching factor' in itself. The transformer therefore should be more or less 'invisible' at system level, meaning to introduce as little loss and/or phase shift as possible. A practical rule of thumb to accomplish this is the transformer to exhibit four times the system impedance at the lowest operational frequency:

$Z_t = 4 \times Z_0$,

At this parallel impedance, standing-wave ratio will deteriorate to SWR 1,28 but will improve at higher frequencies to become even more 'invisible'. The number of turns to arrive at this impedance depends on the type of transformer core material, for which the manufacturer will supply winding factor A_L.

Note! Most manufacturers define winding factor A_L in nano-Henry per turn squared (nH/n^2). Some distributors however prefer a local definition, specifically to make low-permeability materials to look higher. Their definition is A_L in micro-Henry per 100 turns (•H/100). The difference between these definitions is a factor of 10^4! An example:
TN36/23/15-4C65 toroide (36 mm. 4C65 (61) material), permeability = 125, is specified at A_L = 170 (nH/n^2).
T157-2 toroide (1.57" (39,9 mm), grade 2 material), permeability = 10, is specified at A_L = 140.000 (•H/100)
Recalculating the latter to the basic definition will yield: A_L = 14 (nH/n^2). It is better to immediately re-calculate these local definitions to mainstream numbers to avoid confusion.

How many turns?
To calculate the number of turns for a specific application, we may rework the equation: $Z_t = \omega \cdot A_L \cdot n^2 = 4 \times Z_0$, into:

$$n = 0{,}8 \sqrt{ (Z_0 / (f \cdot A_L))}.$$

with:
n = number of turns (at system impedance level)
Z_0 = system impedance

f = lower operating frequency (Hertz)
A_L = winding factor (in Henry!)

Selecting 50 Ohm as the system impedance and a 36 mm. toroide, 4C65 ferrite core (A_L 170 nH/n^2):

$n = 13{,}7 \sqrt{(1/f)}$ (f in MHz. = lower operating frequency)

which would lead to 10 turns for a lower operating frequency of 2 MHz.
Note. More turns is not necessarily the better. At more turns than the required number the parasitic parallel capacitance will increase lowering the maximum operating frequency.

What power?
Depending on the actual operating frequency, the transformer core will be more or less lossy. Depending on the voltage across the transformer, some power will be dissipated in this core-loss and will heat-up the core. Several mechanisms will limit the maximum core temperature.

Maximum temperature

* Curie temperature. This number is related to the maximum ferrite temperature, above which the material will lose all permeability. This implies impedance to drop to very low values which usually will destroy the component and / or the system it was embedded in. When the system is going first, temperature will drop and material will return to the usual permeability. This Curie temperature is very much related to the specific ferrite material, e.g. 350 °C for 4C65 (61) materials and 125 °C for 4A11 (43) materials.
* Mechanical degradation. This is related to the maximum core temperature of non-sintered materials e.g. electrolytic-iron (hydrogen reduced) iron powder and Carbonyl iron powder. According to some manufacturers, these materials will be permanently degraded when exposed to temperatures above 75 °C. although other manufacturers may alow somewhat higher temperatures.
* Winding materials temperature. Some transformers will be constructed using transmission-lines, as in transmission-line transformers. These lines are usually constructed from a soft, easily deformable material that will loose mechanical integrity above 80 °C. Since these lines are being applied at some tension, care should be taken not to make temperatures go high.
In general maximum temperature rise from internal heating for transformers and baluns is restricted to 30 K to also allow for high environmental temperature conditions as found inside power amplifier cabinets and at the antenna at a hot summer day.

Maximum voltage
A transformer at a 36 mm. toroide of 4C65 (61) ferrite material allows for a maximum of 42 Volt per turn at 2 MHz., going down to 24 V. per turn at 30 MHz. just because of this core loss.

In the example transformer above, that required 10 turns for reasons of system impedance, maximum allowable voltage is: 10 x 42 Volt = 420 Volt at 2 MHz., leading to over 3,5 kW of system power in a 50 Ohm system or maximum 10 x 24 Volt = 240 Volt at 30 MHz., leading to around1,15 kW. When constructed from RG58 coaxial transmission-line, it will be this latter that limits maximum system power to 350 W. at 30 MHz. for reasons of maximum current handling capacity.

We prefer to discuss maximum voltage across the transformer over specifying maximum allowable system power. A well designed transformer may still be destroyed at much lower system power when system impedance is higher than designed for, as in a dipole antenna outside resonance. Within power limits, maximum voltage across the transformer may become easily too high leading to much higher internal core dissipation and finally destruction of this component and even the system driving this.

In table 1 an impression may be obtained of the maximum allowable voltage across a one-turn inductor at the well known 36 mm. toroide (36,9 OD x 21,9 ID x 15,7 H), at various ferrite materials and frequencies. For more turns, impedance of the one turn inductor should be multiplied by the number of turns squared as in:
$Z = n^2 \times Z_t$.

Maximum voltage for more turns may be obtained by multiplying the voltage of this one turn inductor by the number of turns as in:
$V = n \times V_{m.diss}$.

Maximum allowable system power may be obtained by taking the maximum, more turns voltage squared and dividing by the system impedance as in:
$P_{trafo\ max.} = (n \times V_{m.diss})^2 / Z_0$.

Ferrite materials; impedance (Z_t) and maximum voltage ($V_{m.diss}$)

of a 1 turn inductor at a 36 mm. toroide

ferriet type	fer.mag. res. freq.(MHz)	1,5 MHz		4 MHz		10 MHz		30 MHz		50 MHz	
		Z_t	$V_{m.diss}$	Z_t	$V_{m.diss}$	Z_t	$V_{m.diss}$	Z_t	$V_{m.diss}$	Z_t	$V_{m.diss}$
3E25	0,5	32,1	11,4	20,3	9,0	20,3	9,0	22,8	9,6	25,4	10,1
3F3	2	35,8	19,2	61,4	15,8	18,8	8,7	23,3	9,7	25,8	10,3
3S4	3,5	22,7	14,2	32,3	13,3	45,8	14,1	56,0	15,8	67,2	17,5
4A11 = 43F	5,5	11,6	15,8	28,6	14,1	44,8	14,5	65,4	16,5	73,9	17,4
4C65 = 61F	45	1,6	37,5	4,2	56,0	11,0	53,5	39,8	23,5	65,9	18,5
68F	250	0,2	10,7	0,5	17,4	1,3	27,6	3,8	47,8	6,3	61,7

F = FairRite / Amidon type

Table 1. Application numbers for various ferrite materials

The second column is the ferri-magnetic resonance frequency (f_r) of the specific ferrite material. This is a materials parameter and is related to the reactance (•') and loss (•") parameters as the frequency where: •' = •" (Q = 1). It is inversely related to permeability, so 3E25 type of material representing the highest permeability and is therefore a low frequency material.

From table1 it may be seen two ferrite parameters to compete when determining maximum voltage across the inductor, i.e. permeability and loss, that both are frequency dependent, but in different ways. When finally loss will be the determining factor at a frequency above f_r, maximum allowable voltages tend to arrive in the same ball-park.
Also, total impedance will still continue to rise long after this ferri-magnetic resonance frequency, which is very useful when designing choke type of applications.

The number of turns at the 'non-system side' of the transformer, and so the transformer ratio 'T', is still un-determined, but will follow from the system to match to. We will attend to this in the next paragraph. Independent of T, most important transformer parameter already have been determined like the core material, number of 'system-side' turns and maximum allowable voltage (system power).

The transformer will match one impedance to an other, but will not influence the power ratio from input to output, apart from a small amount of power loss. This is a 'no-brainer' but will often be forgotten when designing transformers. We will use this simple 'input power is equal to output power' rule quite often.

Background to various ferrite calculations may be found in the series of articles on this subject as in: Ferrites in HF applications' on this web-site.

What transformer type?
We discussed various transformer aspects, that may take different shapes depending on materials and shapes available. Also we discussed 'turns' and transformer ratio's in terms of the classic flux transformer, i.e. transformers where input is coupled to the output by means of the

electro-magnetic field in the core. This transformer type will serve its purpose if well designed but is also suffering from inherent limitations, related to the mechanism of flux coupling and parasitic capacitance across and between 'windings'.

Flux transformers
When core loss is becoming noticeable as a system parameter, i.e. around and after ferri-magnetic resonance of the core material, inter winding coupling is going down. This will be noticeable in the increasing of leakage flux, translating into a series inductance with the transformer which will limit high(er) frequency response. Together with the intrinsic transformer loss, this type of transformers will find their application limit around the ferri-magnetic resonance frequency, i.e. when •' = •" or Q = 1. More on this subject may be found at HF transformers.

Transmission-line transformers
A different type of transformers is consisting of one or more transmission-lines, that are combined in series or in parallel at the input and / or at the output. Transformer ratio is determined by directly adding currents and / or voltages and different transformer ratios may be created although somewhat less flexible than with flux transformers. Because of the inherent wide-band characteristics of these transmission-lines, this type of transformer may be applied over a much wider range of frequencies than the flux type.

Critical point at transmission-line transformers is the input to output separation. For this application, again core materials will be used to create a high input to output impedance. It is virtually unimportant how this high impedance is created, by loss, by reactance or a combination, which is leading to a much wider frequency application range of the same core material, as also may be appreciated from table 1.

As far as 'parallel impedance' at the transformer input is concerned, the same formula's apply as with the flux transformers. Because of a different circuit lay-out, a different strategy has to be followed for feeding DC-power to the active components.
More on transmission-line transformers may be found in: Transmission-line transformers.

The active components

As discussed before, power amplifiers may be designed with vacuum-tubes or solid-state devices or even a combination of the two. At the higher power range vacuum-tubes will usually be the preferred component as it is less difficult to remove the internally dissipated power that results from a less than 100 % system efficiency. Modern higher power broadcast stations may currently be equipped with solid-state output stages, although these have been design as a combination of a number (tens) of lower power solid-state amplifiers 'building bricks', combined to deliver the required high power output. This will also ensure higher system reliability when only one of these building bricks is failing. Whatever the active component, the design procedures are very much comparable.

As an example, we will start designing a power amplifier based on a real life transistor, BUZ308, although this type will not likely be applied in a final HF wide band-width power amplifier. The design procedure however will not be different.

Factory information
The transistor manufacturer will supply a great number of parameters and specifications. To start-off we will look at some limiting values we

should obey to exploit the life expectance for the component. For BUZ308 we find:

- maximum drain - source voltage: 800 Volt (V_{max}),
- maximum drain current: 2,6 Ampere (I_{max})
- maximum dissipation internally: 75 Watt (P_{max}).

Other important information may be supplied on the internal thermal conditions. According to specifications, maximum silicon temperature is T_j = 150 °C (j for junction, the source, gate, drain contacts) with a thermal resistance (R_{th}) between junction and mounting base of $R_{th\ j\text{-}mb}$

= 1,67 K/W. This information will be needed when designing this power amplifier.

Next we will look at transistor characteristics as in figure 2.

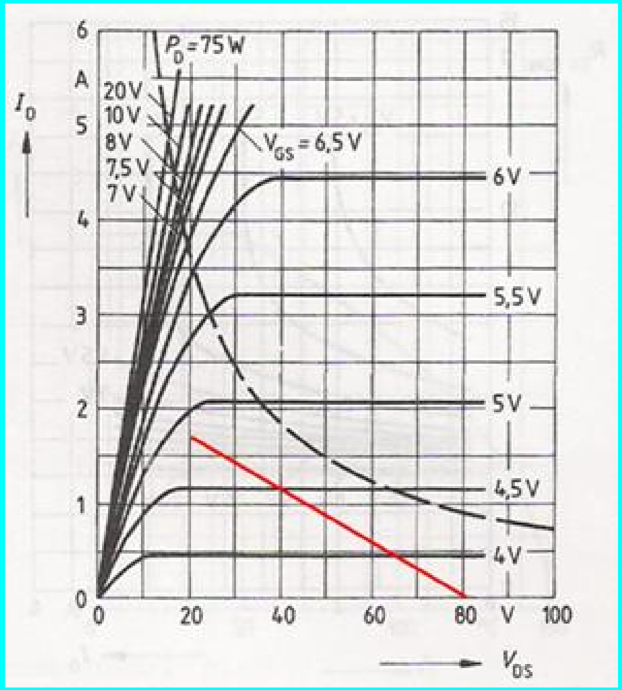

Figure 2: I_D-V_D characteristics of BUZ304

To our surprise we find numbers (far) above the maximum figures for this component (current) or not far enough (voltage). Usually this is meaning the component is (also) being designed for a number of pulse-type applications or specific high-voltage low-current circuits as in inductive switching. Further more this manufacturer has been so kind to also supply the maximum internal power rating as in the dotted line.

Knee voltage
In figure 2 we find a number of horizontal lines for various gate voltages. When we follow the line for Vg = 4,5 V from high to low drain voltage, we find the curve to deviate from the horizontal line at around 15 V., curving back to 0V thereafter. This 15 V. point is called 'knee voltage' and is marking an area we should stay out for linear applications.

Calculation values
Whatever choices we like to make at designing this transistor power amplifier, we should take care to keep some margin to allow for adverse operating conditions e.g. switching to full power at less than ideal matching conditions (SWR > 1). Therefore we prefer to stay away by at least a factor of 1,5 from the earlier mentioned absolute limiting values. Our limiting values for the amplifiers we are designing therefore will be:
$I_{d\ max.} = I_{max}$ / 1,5 = 1,7 Ampere, $V_{d\ max.} = V_{max}$ / 1,5 = 530 Volt, P_{max} = 75 / 1,5 = 50 Watt.

Class A amplifier

Different types of amplifiers may be designed around the selected transistor. We may remember class A amplifiers to operate at low distortion, so this might be our first try. In a class A amplifier, the active element is adjusted at the control port (at the BUZ308, the gate) so half the maximum current is flowing in the output terminal.

Load line
When selecting the maximum (calculation) drain current: I_p van 1,7 Ampere, at the knee voltage: 20 Volt, we may draw a straight line to the point where drain voltage is 80 Volt at drain current is 0 Ampere. This line is always below the dotted line for maximum allowed power so may be regarded as an allowed and safe load line. In figure 2 this is the red line.

With this load line, the power amplifier is completely determined. This may be appreciated since:
- At I_d = 0 Ampere, no voltage may be found across an impedance in the drain circuit, so this voltage may be regarded as the supply voltage: V_b = 80 Volt.

- Maximum output voltage as delivered to the load is equal to the supply voltage minus the knee voltage:
$V_p = V_b - V_{kn}$ = 80 - 20 Volt = 60 Volt.

This is also the top-top value of the output voltage to the load, with an effective (rms) value of:
$V_{out} = 60 / 2\sqrt{2}$ V. = 21,2 Volt rms.

- Maximum current through the transistor (and the load) is 1,7 Ampere and minimum current is 0 Ampere. This is the top-top value of the

output current, with an effective (rms) value of:
I_{out} = 1,7 / 2√2 A. = 0,6 Ampere rms.

- Out of the rms values for voltage and current we may calculate maximum power as delivered to the load:
$P_{out} = V_{out} \times I_{out}$ = 21,2 x 0,6 = 12,7 Watt

- Class A amplifier at no drive is set in the middle of the load-line, so at a drain current of:
$I_{DC} = I_p / 2$ = 1,7 / 2 = 0,85 Ampere

which leads at the load line to a drain voltage of 50 V.

- Total delivered DC input power to the transistor is:
$P_T = V_{DC} \times I_{DC}$ = 50 x 0,85 = 42,5 Watt.

- The efficiency (•) of the power amplifier may be calculated in a percentage of the DC input
• = (P_{out} / P_T) x100 % = (12,7 / 42,5) x 100 % = 29,9 %

This is considerably less that what we expect from a class A amplifier (50%) and is mainly due to the relatively high knee-voltage as compared to the supply voltage. The area below the knee is un-usable (outside linear voltage - current behavior) but is contributing to the total DC input power.

- The load-line may be directly translated into a load resistance, which may be calculated from AC peak voltage and current:
$R_b = V_p / I_p$ = 60 / 1,7 Ohm = 35,5 Ohm,

This in turn is determining the output transformer, that will have an impedance transformer ratio of
35,5 : 50 = 1,4 : 1 , of which the turns ratio is following from the square route: 1,2 : 1.
With 10 turns at the primary side as calculated before, the secondary number is 10 / 1,2 = 8 turns.

Thermal requirements of the class A amplifier
The thermal requirements of a power amplifier are an important part of the design.
From the manufacturers specifications we learn maximum internal temperature: Tj = 150 °C and thermal resistance from junction to mounting base: $R_{th\ j\text{-}mb}$ = 1,67 K/W.

Difference between the amplifier input and output power (see also efficiency) is turned into heat, we may calculate:
$P_W = P_T - P_{out}$ = 42,5 - 12,7 Watt = 29,8 Watt

Because of this power, a temperature gradient will exist across the thermal resistance of:
$\Delta T = P_W \times R_{th\ j\text{-}mb} = 29{,}8 \times 1{,}67 = 49{,}8$ K

Since maximum internal temperature is 150 °C , the allowed maximum mounting base temperature will be:
$T_{mb\ max} = 150 - 49{,}8$ °C = 100,2 °C.

When we would like to operate this amplifier at a maximum ambient temperature of: $T_{amb} = 35$ °C, we may calculate the thermal resistance form the mounting base to free air as:
$R_{th\ heatsink} = (T_{mb\ max} - T_{amb}) / P_W = (100{,}2 - 35) / 29{,}8 = 2{,}12$ K/W, and this is more than a simple cooling flap.

Some conclusion on class A amplifier
By selecting a class A amplifier, we obtained a solution with a relatively low output power and a low power efficiency. Furthermore the power supply will have to supply a relatively high average current of 0,85 A. that will have the output transformer already spend part of the saturation budget and will bring it up to a different linearity regime. Distortion therefore may be less than we expected by selecting the class A set-point in the first place.
It is clear the class A choice to not be an optimal one, so we better investigate a different set-point (amplifier class).

Class B amplifier

Load line
In a class B amplifier, the active element is adjusted at the control port (at the BUZ308, the gate) so (just) no current will flow in the output terminal. When increasing the drive, drain current will flow in the positive half of the drive cycle and no current during the negative half. Since our amplifier is to be operated in the linear mode, we need two active elements that will each operate at the other half of the input drive. Both halves will be combined in the output transformer to generate the full output cycle. This also means each active element to deliver full power at half the time, or on average, half power during the full period. In the off half cycle, the active element will not dissipate any (DC or AC) power.

We may again operate each active element according to the red load-line of figure 2, since this has been selected for maximum drive and power as related to the absolute maximum ratings.
Note the transistor this time should be rated for double the drain voltage since the output voltage of the active device is added to the drain voltage of the non-active device, that is at supply voltage level at no drive.

At the selection of the load line, the complete amplifier is determined.

- At $I_d = 0$ Ampere, supply voltage: $V_b = 80$ Volt.

- Maximum output voltage swing:
$V_p = V_b - V_{kn} = 60$ Volt.

This is the peak value of one drive period, the other half is generated by the second transistor. Effective voltage of this half cycle is:
$V_{out} = 60 / \sqrt{2}$ Volt = 42,4 Vrms.

- Maximum current is again 1,7 Ampere peak value, with an effective value of
$I_{out} = 1{,}7 / \sqrt{2}$ Ampere = 1,2 A.rms.

- During the half period cycle each transistor is generating
$P_{out} = V_{out} \times I_{out} = 42{,}4 \times 1{,}2 = 50{,}9$ Watt, and no power during the next half period when the other transistor takes over. Total output power for this class B amplifier therefore is also 50,9 Watt.

- Through each transistor a current is flowing only during the active cycle. Average value of this current is:
$I_{DC} = 2 \times I_{max} / \pi = 1{,}08$ Ampere.

DC voltage at the transistor is equal to the supply voltage, so total DC input power per transistor, per half cycle is:
$P = I_{DC} \times V_{DC} = 1{,}08 \times 80 = 86{,}4$ Watt, which is also total DC input power for the full drive cycle of this class B amplifier.

- Total efficiency of this class B amplifier again is calculated as a percentage of total DC power as delivered to the amplifier:
• $= (P_{out} / P_T) \times 100\ \% = (50{,}9 / 86{,}4) \times 100 = 58{,}9\ \%$.

This again is lower than the theoretical maximum for this class of amplifiers (68%) again because of the relatively high knee voltage.

- Drain load (load-line) per transistor is the same as in the class A amplifier, making also turns ratio (per transistor) identical. The output transformer therefore will have a turns ratio of (1+1) : 1,2 or (8 + 8) : 10 in turns.

Thermal requirements of the class B amplifier
Calculations are running in parallel to those in the class A amplifier. Power as delivered to the amplifier that is not delivered to the output, will be generated in heat and should be drained away to below permissible temperatures. For this class B amplifier and per transistor:
$P_W = P_T - P_{out} = 43{,}2 - 25{,}5$ Watt = 17,8 Watt.

This will generate a temperature gradient in the transistor from junction to mounting base:
$\Delta T = P_W \times R_{th\ j\text{-}mb} = 17{,}8 \times 1{,}67 = 29{,}6$ K

The maximum allowable mounting base temperature therefore will be
$T_{mb\ max}$ = 150 - 29,6 °C = 120,4 °C.

When we would like to operate this amplifier at a maximum ambient temperature of: T_{amb} = 35 °C, we may calculate the thermal resistance form the mounting base to free air as:
$R_{th\ heatsink} = (T_{mb\ max} - T_{amb}) / P_W$ = (120,4 - 35) / 17,8 K/W = 4,8 K/W.

When we prefer one mounting base for the two transistors, thermal resistance will be halved to 2,4 K/W.

Comparing class A to class B amplifier.
To compare the class A to the class B amplifier, table 2 has been generated:

	efficiency (%)	**output power (Watt)**	**heat generated (Watt)**	**heat sink (K/W)**	**supply power (Watt)**
class A	29,9	12,7	29,8	2,12	42,5
class B (2 tr)	58,9	25,5 (50,9)	17,8 (36)	4,8 (2,4)	43,2 (86,4)

Table 2. Comparing the class A to the class B amplifier

In table 2 it is immediately clear what the efficiency difference between the two amplifier classes is bringing. At about double the input power by the power supply, the class B amplifier is delivering four times the output power and will require about the same heatsink as the class A amplifier.

Since each transistor in this class B amplifier is taking DC current per cycle, no net DC current will flow though the output transformer so full drive capabilities are available for HF-power with inherent lower distortion.
This example is showing it pays to go for amplifiers with high efficiency because of the positive effects at various system components.
Selecting a low 'knee-voltage' transistor will further improve efficiency.

Note, all calculations have been performed at full output power. At lower (average) HF output power as in SSB operation, average efficiency will also be lower.

Some practical remarks

Rest current
In a practical class B amplifier each transistor usually will not be adjusted to a no-current at no-drive situation. Starting from zero drive, the first part of the transistor characteristic is not very linear, generating harmonic content in the output current. By adjusting each transistor to a small 'resting current' this non-linear drive area may be avoided. This amplifier-setting is called a class AB amplifier since the set-point is somewhere in-between class A and class B. At this set-point however, transistors are operating in class B mode for most of the time.

The class AB setting is requiring some additional DC power that will not be delivered to the output and so will make total efficiency of this class somewhat lower then full class B amplifier, with additional (small) requirements to the heatsink.

Feedback
A second practical point concerns the transistor un-equality. Even from the same manufacturing batch, transistors will always exhibit small differences as to their characteristics. This may result in one transistor to still be completely switched-off with the other transistor already fully at the class AB drain current for the same set-point. A second effect may be noticeable at the output with one transistor delivering full power and the other just keeping a 'toe in the water'. These undesired situations we of course very much like to avoid.

An effective way of dealing with these effects is to apply some form of feed-back, for instance by having a small portion of total drain current to deliver a voltage across a source resistor. This voltage will effectively be in series with the input drive and set-point, but in the opposite phase. This feed-back voltage will have a positive and equalizing effect to transistor drive characteristics to the DC (set-point) as well as to AC (linearity). As a rule of thumb, source voltage should be in the range of 10 - 30 % of the gate voltage at full drive swing, this to be applied even when transistors are selected for pairing.
With these feed-back resistors, linearity will be improved and so output harmonic content will be lowered. Since this a an HF amplifier, these source resistances should be of a 'no-inductance' type and also of sufficient power handling capacity since source currents may be high.

Paralleling transistors
To enhance current drive capabilities, more transistors may be connected in parallel. To ensure good tracking of characteristics, the same principles of feed-back apply.

As for the calculations (load-line, power delivery and handling, heat considerations), the same principles apply. Because current are doubling (for two transistors in parallel) the same characteristics of figure 2 apply, this time with doubled current figures on the vertical axes. With three transistors in parallel the figures obviously triple, etc.

Note, the matching transformer also to match to a lower impedance, with half the primary impedance for two transistors in parallel one third at three transistors etc. All further calculations will follow the, by now familiar patterns as above.

Some conclusions

- With the above method it should be not too difficult to design a power amplifier with components already available from the junk-box (transistors, ferrite cores, heatsink etc.).

- Although high linearity is always a design requirement for wide-band amplifiers, this should be balanced to other practical requirements as system efficiency and available drive and DC input power.
- High system efficiency usually translates into simpler requirements to the heatsink, power supply and will result in more output power.
- The higher the amplifier class (A, AB, B, C, currently also D and E are practical), the higher system efficiency although not all classes may be applied in wide-band linear amplifiers.

Closing remarks

The next project phase is concerned with practical system construction. To this phase, practical experience is an important requirement and usually means battling with al sorts of parasitic and non-ideal effects, that are becoming more important as frequency rises. Therefore it is important for a starting designer to study the designs of experienced people, with the principles as explained in this article as a basis. In this way the practical learning curve will be less steep and my help starting your own amplifier design

Bob J. van Donselaar.
mailto:ON9CVD@AMSAT.org

Core Baluns

(published in Electron #4, 2007)

General

This chapter is the third in a series of articles on baluns for antenna applications. The first article is an introduction with some background on balun types. The second article is on wire baluns. It is advisable to read the articles in the above order especially since each next chapter is building on information and formula's already explained earlier and referencing to this.

Introduction

When going from an a-symmetrical to a symmetrical system some form of adaptation / matching should be applied to uphold the integrity of both systems. This may be accomplished by ensuring no current to flow anywhere except for the desired signal path. A simple and direct way to prevent other currents to flow is blocking-of all other current path'. A sleeve choke, 1 : 1 current transformer or current balun will perform this system function. The better this choke, i.e. higher impedance, the better this current balancing function in the signal path.

Calculations

With the current balun requirements as settled, we checked the performance of a wire-balun in the previous chapter. Here we discovered this component could well be constructed for a limited frequency range (1 : 3), but was increasingly more difficult to design for higher bandwidth. Main problems were leakage inductance at lower frequencies and high parasitic capacitance to limit the high frequency side. Both problems may be attacked when applying a ferrite core at the balun since high permeability will enhance inter-winding coupling (prevent leakage inductance) while at the same time allowing for a lower number of turns (diminishing parasitic capacitance).

As an example we will calculate a 1 : 1 current balun around a 36 mm. toroide of 4C65 (61) ferrite material. For this function to be effective, sleeve impedance should be at least four times system impedance which we will set at 50 Ohm, so sleeve impedance should be 4 x 50 Ohm = 200 Ohm. At the lowest operating frequency of 2 MHz. this translates to a reactance of 16 •H. as we noticed in the previous chapter.

At the manufacturers web-site we will find for this material and core shape the winding factor A_L = 170 nH/n^2, so we will need a number of turns at this core:
n = √16 / 0.17 = 9,6 (10) , which leads to the construction as in figure 1.

Figure 1. Current balun 1 : 1

Note: The balun in figure 1 has been constructed using RG58 coaxial transmission line. Since the electromagnetic field is 'locked-up' inside the transmission line, the integrity is being preserved independently of the 'handling' of the line, whether in a straight line, curled up or around some core material. This applies to coaxial transmission-line as well as symmetrical line types. Since the latter is not as 'closed' as coax, turns should not be wound too closely and preferably not to overlap.

Measurements

Transmission
The balun as above has been constructed and measured in a condition of maximum unbalance, as discussed before. Results for transmission (insertion loss) may be found in figure 2.

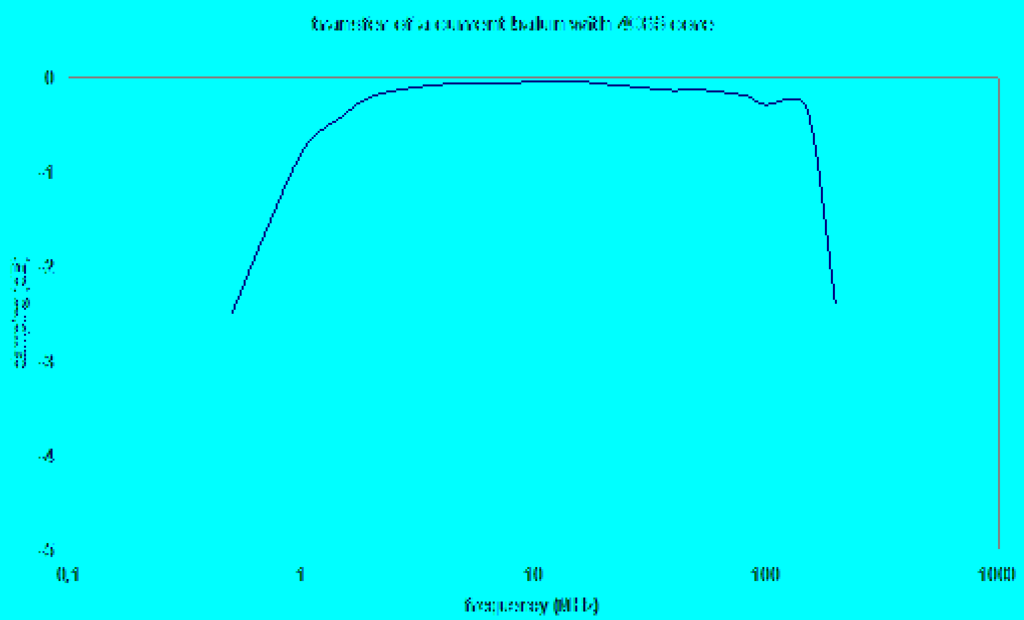

Figure 2. Insertion loss of a 4C65 current balun

The insertion-loss for the 1 : 1 current balun on 4C65 may be directly compared to the wire balun in the previous chapter. It is easily noticed the 4C65 construction to perform significantly better, with the 1 dB loss position at the lower frequency side to have shifted from 1,8 MHz. to 0,9 MHz. and the high frequency side from just over 50 MHz. to over 150 MHz. In the interesting frequency range 3 - 30 MHz., insertion loss has improved from 0,16 dB on average to 0,07 dB and this may be regarded as almost 'ideal'.

Reflection

As with the wire balun we also measured reflection with the balun again terminated into 50 Ohm. The results of this measurement may be found in figure 3.

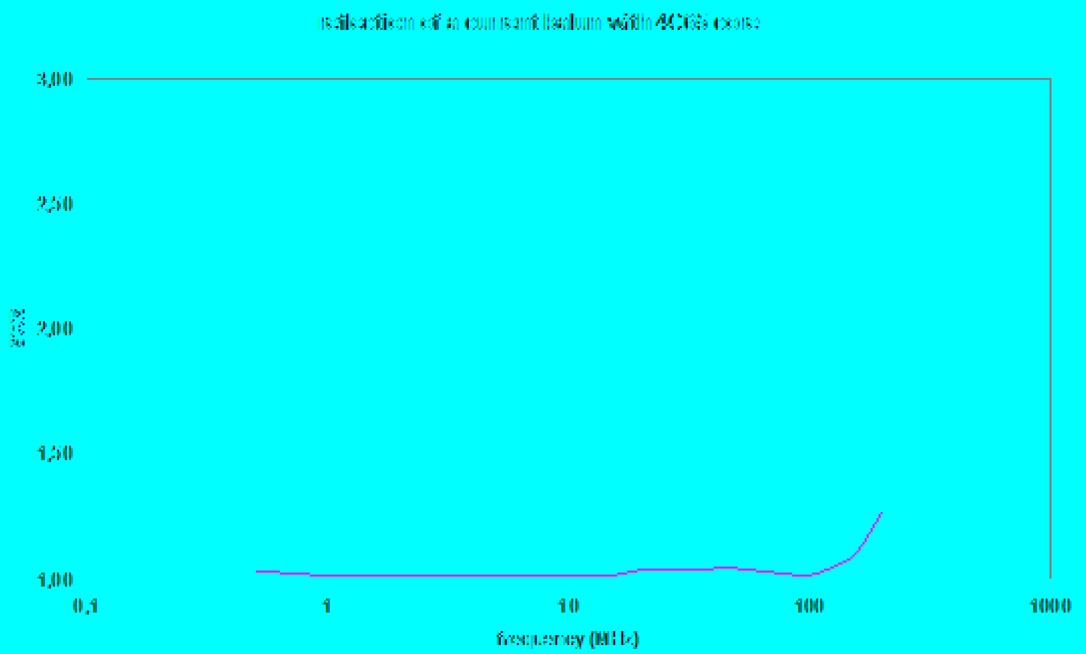

Figure 3. Reflection of a 4C65 current balun

Again the improvement as related to the wire balun is clearly visible. SWR is very low (1,03) from the first measuring frequency at 0,5 MHz. (compare to SWR = 1,5 @ 4 MHz. for the wire balun), up to SWR = 1,25 at 200 MHz. (compare to SWR = 1,5 @ 50 MHz.).

Power
There is not such a thing as a free lunch, so we may be curious as to what price these excellent properties?
As it happens, price to pay is very low. In the chapter on [HF ferrite applications]() we found that above 3 MHz. maximum core load is determined by power dissipation inside the core, with highest stress at the highest operational frequency. In the same chapter we derived a formula for the system voltage that will make core temperature rise by 30 K. at the maximum internal power dissipation of P_{max}.

$$U_{dissipation} = 2{,}8 \times 10^{3} \times n \times \mu_r \times \frac{B_{max} \times f \times A}{l \times \mu''}$$

to be simplified to

$U_{L(dissipation)} = \sqrt{(P_{max} \cdot X_L (Q + 1/Q))}$

with P_{max} = 4 Watt for this 36 mm. ferrite core. At a highest operational frequency of 30 MHz. we find
$U_{L(dissipation)}$ = 250 V, allowing for a system power of 1,25 kW in a 50 Ohm system.

When using RG58 it again is the transmission-line specification that will set the limit at 350 Watt.
We should be aware (again) that above calculation is for a system impedance of 50 Ohm, e.g. about the impedance of a resonating dipole antenna. Outside resonance, impedance will be much higher, even though the antenna tuner in the shack will have transformed all odd impedances back to 50 Ohm for the transceiver. The balun may therefore be stressed beyond allowable limits without the operator being aware of this.

Different cores

In the first chapter on baluns we discovered the minimum choke impedance of four times the system impedance. It is not important how this impedance is created, as long as this is high which is the same is we found in Ferrites for EMC applications.
One therefore may be wondering whether only HF ferrites will qualify and what would happen when applying LF ferrite materials?
To test this, I made the same balun (10 turns on 36 mm. toroide) on 3E25 type of ferrite, a LF type of material with a high initial permeability of 6000, as compared to 4C65 (125).

Transmission
Measuring set-up is identical to the other tests (under maximum unbalance conditions) and results may be found in figure 4.

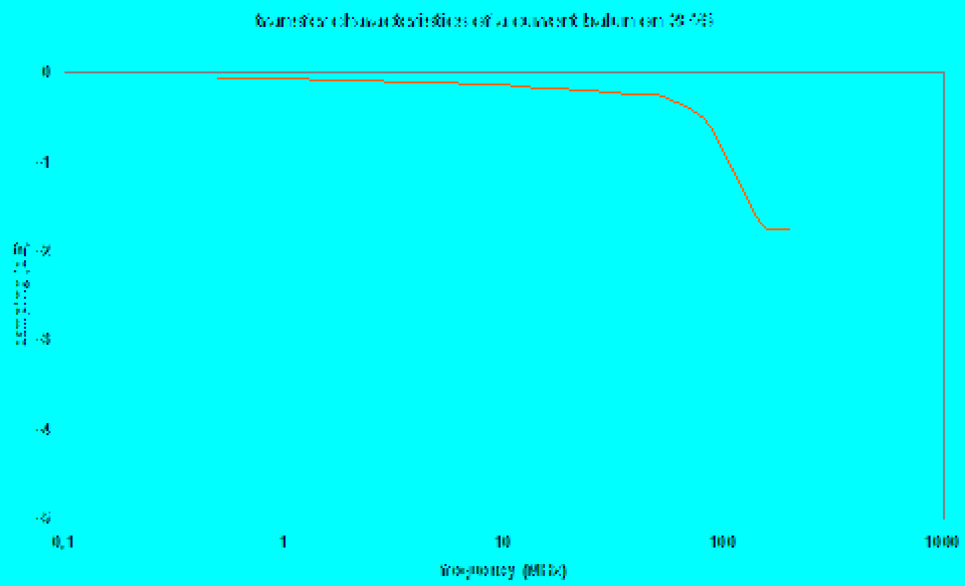

Figure 4: Insertion loss of a 3E25 current balun

As may be expected the low frequency behavior is even better than with the 4C56 core, and will probably extnd to far below the lowest test frequency of 0,5 MHz., although we do not need this for HF applications. At the high frequency side we find a 1 dB loss at 100 MHz. because this LF material really was designed for low frequencies. At 100 MHz. even the material's loss-factor is at the end of its operational life.

Reflection
Reflection properties of this 3E25 balun may be found in figure 5.

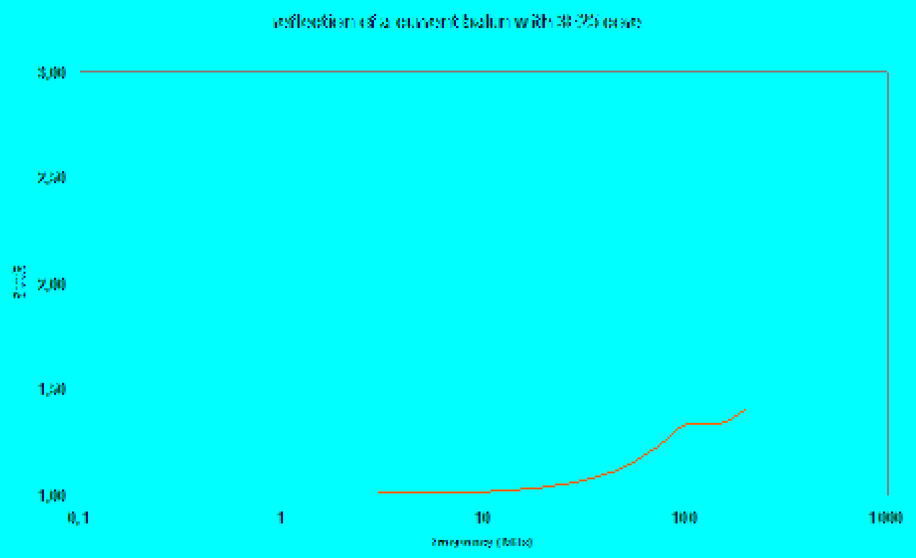

Figure 5. Reflection of a 3E25 current balun

Measurements extend to 0,5 MHz. but are so close to the X-axis that this is not showing in this graph anymore.
Again we find a very useful component for HF frequencies that may be applied to over 100 MHz.

Power
Since parameters are different again for this core material, we calculate maximum voltage across this choke with the formula above, at highest operational frequency, to find:
$U_{L(dissipation)}$ = 203 V.

This allows for a system power of over 800 Watt in a 50 Ohm system, still more that allowed for when using RG58 as the transmission-line material.

Although exhibiting somewhat less bandwidth and power than the balun on 4C65, also this 3E25 balun will perform an excellent job over the entire HF frequency range. Since this is a (very) low frequency material it is save to say that all ferrite material from the junk box will qualify as a core for a current balun.

Parasitic capacity
As we have discovered, these current baluns will perform well as long as the sleeve impedance is high. In practice we find a parasitic capacitance in parallel to the choke with a diminishing impedance with frequency. Since the value of a capacitor is

related to the voltage across, we like to keep first and last turn of the choke as far apart as possible. When measuring the effect of distance it appears that for these current baluns a distance of about one centimeter (two RG58 diameters) is sufficient for this capacitance effect to be low.
An alternative winding technique has been tested to keep this first to last turn far apart. This technique first puts half the number of turns on the core and then crosses over to complete the other half of the turns in the opposite direction. First and last turn will end up half a core diameter away. As it happens, cut-off frequency was lower than with the straight winding method, so this allegedly 'better technique' is no improvement.

Bead balun

In the last chapter and also before the role of the parasitic capacitance at the high frequency side was clear and was a limiting band-width factor. To diminish this capacitance we may consider to not coil-up the transmission-line around the core, but 'coil the core' around the transmission-line. The latter situation arises when stringing ferrite beads on the transmission-line since we also know that every time the line passes through the center hole of the toroide, this accounts for a full turn. The string of beads therefore acts like many reactances in series, as also applies to parasitic capacitances.

When constructing this 'bead-balun' at RG58 coaxial cable (diameter is 5 mm.), a well fitting toroide shape is TN10/6/4. Using ferrite type 4C65, winding factor A_L = 52 nH/n^2, so for a total inductance of 16 •H we would need over 300 beads. This balun would be 1,35 meter long, which is not really practical anymore. When selecting 4A11 type of material (A_L = 286 nH/n^2) we would need 56 beads with a total length of 25 cm., or when selecting 3E25 (A_L = 2250 nH/n^2) only 3 beads at a length of 3 cm.

Since maximum dissipation of these beads is scaling with the root of volume (see Ferrites in HF applications) , maximum power dissipation of a 4A11 bead of this size is 30 mW. Maximum voltage per bead is:
$U_{L(dissipation)}$ = 1,6 V.

allowing for maximum voltage of 90 V. for the 56 bead balun, which translates to maximum system power of 160 Watt in a 50 Ohm system. Although all ferrite batches are created equal, some beads may be more equal than others making up for a different voltage distribution, so a maximum 100 Watt would be a safer limit for total system power.

When preferring somewhat larger RG213 cable with outside diameter of 10 mm., the bigger size ferrite bead TN23/14/7 would fit and would have a winding factor A_L = 485 nH/n^2 when 4A11 type of ferrite will be applied, requiring only 33 beads. The balun will now have a total length of 24,8 mm. and would allow for a maximum voltage of 2,9 V. / bead, or in total 95,7 V. allowing for 183 Watt in a 50 Ohm system.
Except for using a lower number of beads, everything else did not really change that much.

Also the route along the 3E25 road is not really better since the number of beads, the •' and •" will be lower making maximum allowed system power even lower still. It is clear bead baluns will be superior for very-wide bandwidth applications (EMC measurements) but will be less practical in HF (only) applications since the simple 10 turn, 36 mm., 4C65 toroide is already quite near 'perfect' for these applications.

Voltage balun

Up to now we have been discussing current baluns since we found in the first chapter this to be the device we would need to match a symmetrical dipole to an a-symmetric transmission-line. Still one may regularly find voltage baluns recommended for this type of applications in ham magazines. The usual construction of this voltage balun is a trifilarly wound transformer, wired-up such that the balanced connection is taken around a 'ground' connection and the a-symmetrical side with reference to this ground terminal as in figure 6.

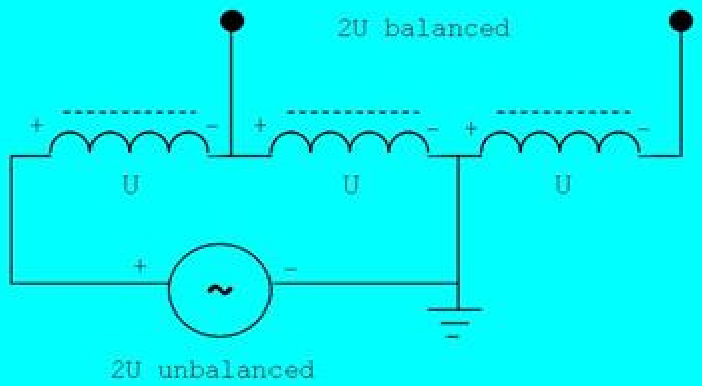

Figure 6. Example of a 1 : 1 voltage balun

In figure 6 we find the generator connected across two windings of the transformer with the balanced side also taken from two windings. Since all windings consist of the same number of turns (trifilar) and are closely coupled, input and output voltages are equal.

In the diagram the balanced output is situated around a 'grounded' terminal, which may not be so very much 'grounded' anymore at the end of the feed-line. According to 'Reference data for Radio Engineers', standard hook-up wire is showing an inductance of 1,6 mH/mile (1 •H/m) for frequencies between 3 - 30 MHz. For a feed-line length of 10 m., inductance is 10 •H or around 600 Ohm at 10 MHz. It is clear this 'ground' terminal is at low impedance any more.

The model of figure 6 may be constructed as in figure 7, when applying the same principles as in our earlier designs (10 turns on 36 mm. toroide of 4C65 material to obtain 200 Ohm at 2 MHz.). The generator is connected across two windings in series, so effectively across 10 turn.

Figure 7. Trifilar 1 : 1 voltage transformer

Measurement at ideal operating conditions

To measure this transformer at ideal operating conditions, I constructed two transformers and connected these 'back-to-back' to keep the balanced side in-between the transformers at a perfectly balanced level. Results of the measurements have been 'halved' to represent each of the transformers.

Transmission
Results of the transmission measurement may be found in figure 8.

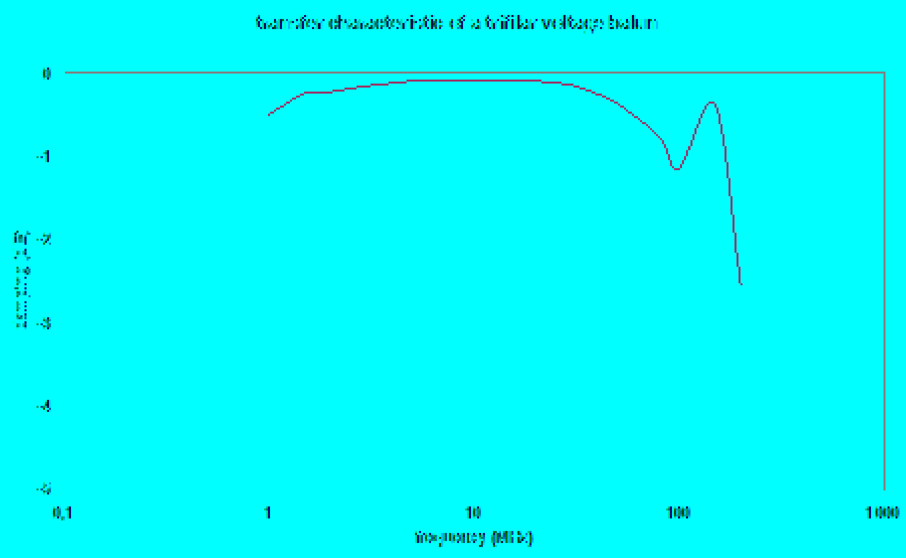

Figure 8. Insertion loss of a trifilar voltage balun

The insertion loss as in figure 8 is showing a fairly good transformer that starts dropping-off below 1 MHz. and above 60 MHz. In between the graph is flat and is showing low loss.

Reflection
Reflection characteristic of the trifilar voltage balun may be found in figure 9.

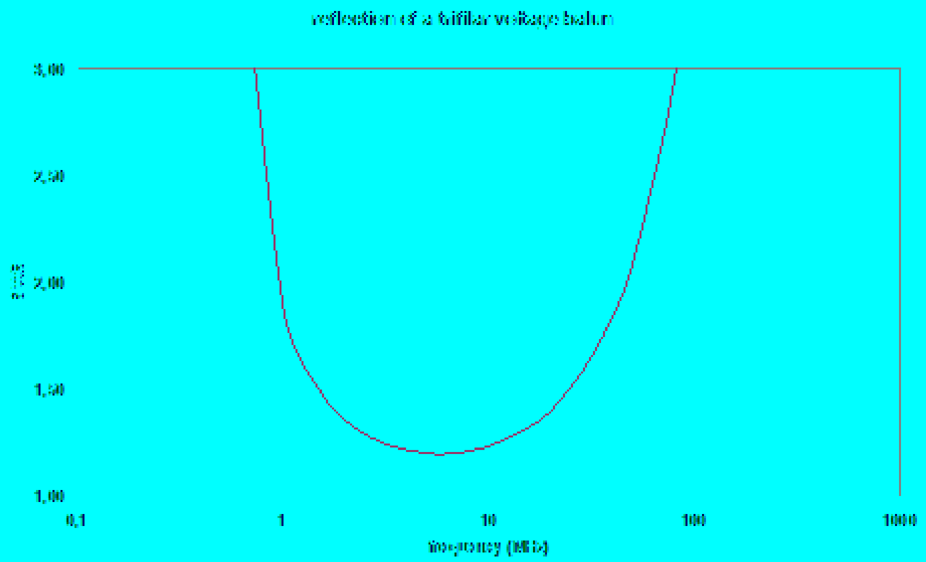

Figure 9. Reflection of a trifilar voltage balun

The graph of figure 9 is showing SWR which is somewhat better than the first wire balun in the previous chapter at the low frequency side but definitely worse at the high frequency end. Also in-between cut-off frequencies the SWR is not really low at 1,25. All in all this is a marginally performer.

Measuring under practical conditions

In the above situation the voltage balun has been tested under 'ideal' conditions, i.e. a condition of perfect balance. In a practical situation there usually is some form of unbalance, up to total unbalance in an adverse situation. Therefore all earlier baluns have been tested in a fully unbalanced situation to prevent surprises when in an operational situation. To compare balancing qualities of various balun types, also this voltage transformer has been tested under the same conditions with one output terminal grounded, to simulate a condition of maximum unbalance. The voltage balun however is not a symmetrical device as may be appreciated when regarding figure 6. One output terminal is connected to a winding to the generator, while the other output terminal is left open. For a complete impression we therefore have to measure this voltage balun two times, with different output terminals grounded.

Note, this unequal situation also exists for the current baluns. In testing with the 'center-conductor' grounded we always have been measuring under most unfavorable conditions. With the other terminal grounded in a current balun, we would have measured just an other transmission-line with only transmission-line damping as a non-ideal condition. Also when measuring in reflection only a 'perfect' transmission-line would have been found.

Transmission

Two measurements, each with a different output terminal grounded may be found in figure 10.

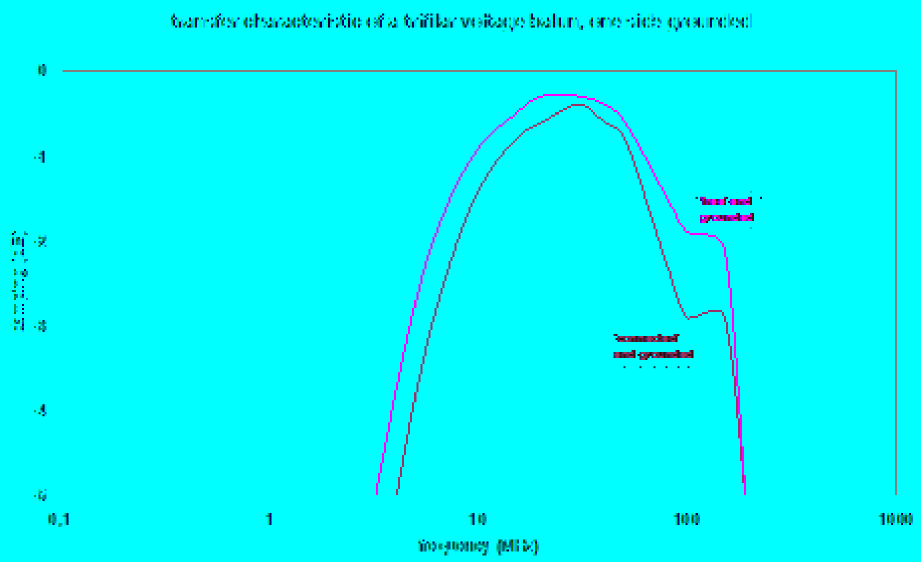

Figure 10. Insertion loss of a trifilar voltage balun, one side grounded

In figure 10 we indeed see different behavior depending on the grounded terminal. Even the best of the two graphs only is showing insertion loss below one dB over a limited HF frequency, and only above 10 MHz. This is not the type of behavior we would like to see.

Reflection

To complete this series, we also have measured the reflection characteristics, as shown in figure 11.

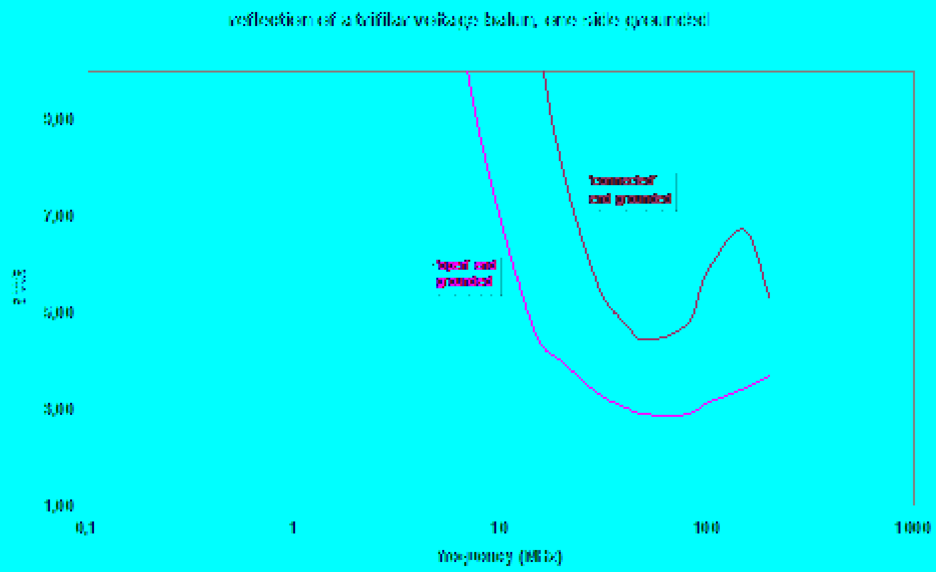

Figure 11. Reflection of a trifilar voltage balun, one side grounded

Again in figure 11 we find a component that will not qualify in an antenna system and for which reflection is un-acceptably high at all frequencies. Note the vertical axes of the diagram has been changed with respect to earlier and comparable diagrams to show anything at all on this voltage transformer.

This 'weird' behavior may be better understood when looking again at figure 6; by grounding one of the output terminals the voltage transformer is partly shortened out. The transfer characteristics may be affected less because of capacitive coupling 'across' the transformer.

When adding all up it is obvious this voltage transformer is not the ideal component (to say the least) to connect a balanced antenna to an unbalanced transmission-line and it is surprising (this type of) voltage baluns are still to be found in some antenna system and / or be recommended in magazines for this application.

Some conclusion about 1 : 1 baluns on ferrite cores

From the above it may be concluded that a **current balun** in general is the component of choice to connect a symmetrical antenna to an a-symmetrical feed-line. When constructing this current balun, any ferrite material from the junk-box will do. When applying a 36 mm. ferrite toroide, ten turns will be sufficient when applying 4C65 material and 6 - 8 turns are optimal for other (less HF) type of material. For this type of current baluns, more turns usually will make a lesser quality component since

increasing parasitic parallel capacity will decrease maximum usable frequency.
Since the 1 : 1 current balun is based on isolating the input from the output terminals, the same quality may be applied to good use to stop undesired HF currents to arrive at the transceiver. It's therefore good practice to also apply this current balun / sleeve choke at the other side of the feed-line at the receiver or at any other electrical appliance that is picking up HF currents (audio / video equipment).

Voltage baluns should be avoided around balanced antenna systems because of sensitivity of these components to any type of unbalancing (trifilar balun) or because of diminishing internal coupling (other flux-transformer types).

Current baluns on powder-iron (carbonyl) cores

So far we discussed sleeve chokes / current baluns in the form of larger coils 'on air' or on ferrite cores.
Sometimes also baluns on powder-iron / carbonyl cores are being offered, which should be regarded with some care. To compare with ferrite materials, let's look at the 36 mm. toroide shape. Using 4C65 ferrite with an initial permeability of 125, we needed 10 turns to make an inductance of 16 •H, that will provide a parallel inductance of 200 Ohm at 2 MHz. When we were to apply e.g. Amidon grade 2 carbonyl core (permeability = 10), of the same shape and dimensions, we would require 3,5 x as many turns (√ (125 / 10)) to obtain the same impedance. Since the center core hole is 22 mm. 35 turns of RG58 will not pass and even using miniature (teflon) coaxial cable will be challenging, next to boosting parasitic capacitance to an un-acceptable level.
For current baluns / sleeve chokes, powder iron / carbonyl type of materials are less suitable materials.

Please contact me for your remarks and suggestions at:

Bob J. van Donselaar
mailto:on9cvd@amsat.org

High impedance Baluns

(published in Electron #4, 2007)

General

This chapter is the fourth in a series of articles on baluns for antenna applications. The first article is an introduction with some background on balun types. The second article is on wire baluns. The third article is discussing baluns around core materials. It is advisable to read the articles in the above order especially since each next chapter is building on information already explained earlier and referencing to this.

Introduction

Many types of balun are designed for application in a low impedance environment e.g. for matching to the input and output of (balanced, power) amplifiers, resonant antenna's etc. System impedance will be between a few Ohm to around 100 Ohm which requires the (current) balun to exhibit a sleeve impedance of hundreds of Ohm that is not too difficult to realize over a wide band-width. Low frequency behavior is determined by the sleeve reactance which will quickly reach the required values when being constructed around a ferrite core. High frequency drop-off is mainly determined by the parasitic capacitance across the (sleeve) reactance, this in turn depending on the number of turns and the proximity of the first to last winding.
At antenna systems outside resonance, system impedance may be considerably higher than 100 Ohm, with comparably higher demands to the design and construction of baluns. An interesting analysis of such situation may be found in the article of Kevin Schmidt, W9CF, Putting a balun and a tuner together. It is shown that for high impedance antenna systems, balun (sleeve) impedance should be above 1000 Ohm and preferably even higher.
To ensure these high impedances over a wide bandwidth is not trivial anymore.

A few system requirements

A balun to connect to a high impedance, balanced feed-line is not a simple component.
- It should force equal currents in the balanced connections with as low a current as possible to ground. This is always a problem at 'flux'- transformers that usually will not be constructed in perfect symmetry (see e.g. trifilar transformer in previous chapter). Furthermore, any form of un-balance will make the feed-line radiate / receive in an undesired way as your neighbors will tell about your transmissions and you will discover yourself tapping into the (usually vertically polarized) electro-smog in your environment.

- It should exhibit high transfer efficiency at all operating frequencies. With comparably low allowable internal power dissipation (e.g. maximum 4 Watt in a 36 mm. ferrite toroide for a temperature rise of 30 K), maximum allowable system power is very much related to this transfer efficiency.

Especially the first requirement differentiates a balun for a symmetrical feed line from coaxial feed-line applications.

Symmetrical voltage balun

Diagram
Even to date we may find see proposals for antenna tuners at with a voltage balun in the output circuit to take care of a balanced output (-currents). To this extend the voltage balun is looking something like figure 1.

Figure 1. Design of a symmetrical voltage balun.

For this voltage balun, the primary and secondary winding usually have equal turns although step-up ratios may be seen as well. In the 1 : 1 variation, the generator voltage 'U' will also appear across the load and with proper winding techniques the secondary winding may be balanced around the ground connection in order to also balance the parasitic capacitance.

Construction
A practical solution to this balancing problem may be found in figure 2. In this transformer the secondary turns have been put in-between the primary turns, like a screw within a screw. Input terminals are on one side of the toroide, the output terminals at the other side to balance each output terminal with respect to ground.
All turns have been spaced equally at the inside of the coil-former, to minimize inter-winding capacity.

Figure 2: Voltage balun with a 1 : 1 transformer ratio

When designing this balun for a 50 Ohm system, with a lower operating frequency of 2 MHz., at a 36 mm. 4C65 (61) ferrite core, a number of 10 turns will suffice to obtain a reactance of 200 Ohm, as we have calculated in previous chapters. The inter-winding capacitance has been measured at 1 kHz. to be 15 pF.

Measurements
As in previous chapters this transformer has been measured for transmission and reflection. Because of the construction details, a small unbalancing may be expected so the measurements have been performed two times, each time with a different output terminal grounded. For all measurements, the balun was terminated into 50 Ohm. Figure 3 presents the transfer measurements.

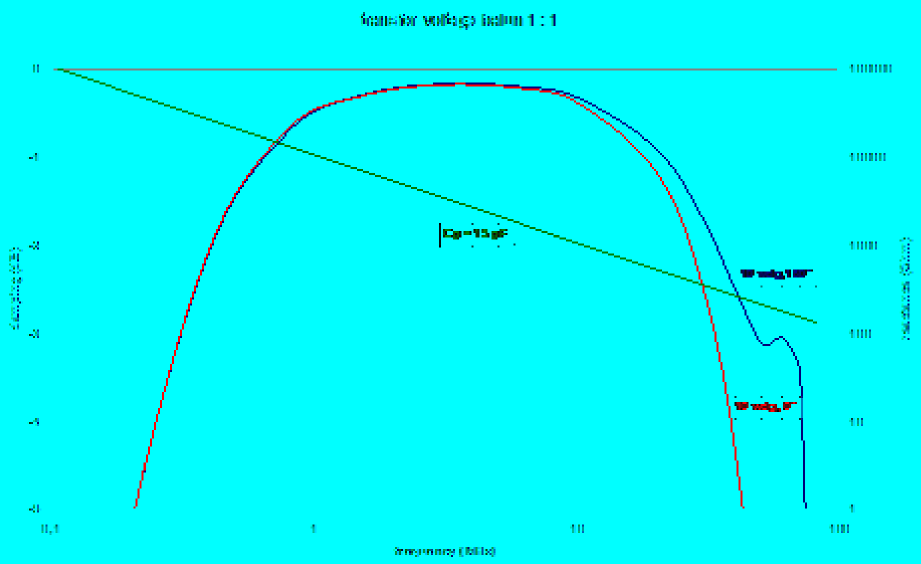

Figure 3: Insertion loss of a 1 : 1 voltage balun

In figure 3 it is shown that indeed some transfer difference may be noticed . When measured in counter phase (blue curve) transfer is improved at 20 MHz. by 0,3 dB and at 30 MHz. by 0,8 dB. This different transfer behavior most probably is due to some capacitive unbalance in this voltage transformer

Apart from some unbalancing, transfer characteristics are not really ideal either with transfer loss of 1,7 dB and 2,5 dB at 30 MHz. This will limit application in a high(er) power system at this operating frequency. Never-the-less, this 1 : 1 voltage transformer is doing a better job than the trifilar 1 : 1 transformer we discussed earlier in the chapter on baluns with cores.

Reflection measurements
We also performed some reflection measurements, identical to earlier transformers. Results may be found in figure 4.

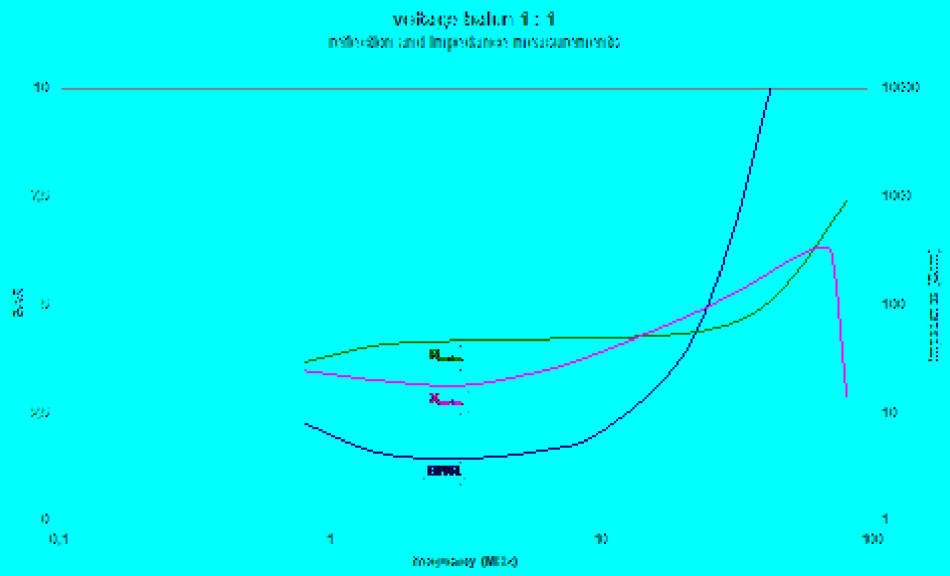

Figure 4: Reflection / impedance measurements

The reflection graph (SWR) is showing some interesting facts:

- SWR never drops below 1,5
- Below 2 MHz. SWR is going up because of diminishing impedance
- Above 10 MHz SWR is going up sharply to take impractical values above 20 MHz.

To investigate this disappointing behavior we measured the real and imaginary impedance values that make-up the reflection curve. In figure 4 we notice reactance (X_{series}) to not drop to insignificant values, which accounts for the not too low SWR in the frequency range up to 10 MHz. Above this frequency, the reactance rises further due to loss of inter-winding coupling. This effect is showing up as a leakage induction, making (series) reactance to go up again with frequency.

For 4C65 materials, ferrimagnetic resonance frequency (f_r) is at 45 MHz., meaning permeability and loss to have the same value (Q-factor is 1). This is showing in figure 4 in the rising of X_{series} (lowering permeability, loss of coupling) and in the rising of R_{series} (rising ferrite loss).

Around 65 MHz. rising of X_{series} breaks off to fall off sharply and become negative. This is the effect of the parasitic parallel capacitance becoming the dominant impedance.

We might consider applying higher ferrimagnetic resonance materials, e.g. 4D2, with f_r at 150 MHz. Unfortunately the higher the ferrimagnetic resonance, the lower the permeability (Snoecks law) so we would need more turns to comply with reactance requirements at the low frequency cut-off. More turns also would mean higher parasitic capacitance and this would lower the high frequencies cut-off.
Because of these effects this voltage balun may not be improved too much beyond current, already not ideal performance.

To operate as a balun for a symmetrical tuner, impedance to ground should be 1500 Ohm or higher. Going back to figure 3 one may notice the inter-winding reactance to be below this value already at 10 MHz. and at 30 MHz. even below 350 Ohm. This again makes this voltage balun less fit for the job.

Current balun for a symmetrical tuner

Let's look again to the earlier current transformer, this time specifically for application in a symmetrical tuner. All earlier requirements still apply and so we again arrive at the 4C65 current balun at a 36 mm. toroide with ten turns of RG58 as in figure 5.

Figure 5. Current balun 1 : 1

The transfer characteristic for this component may be found in figure 6, this time with the graph for the sleeve impedance to ground added.

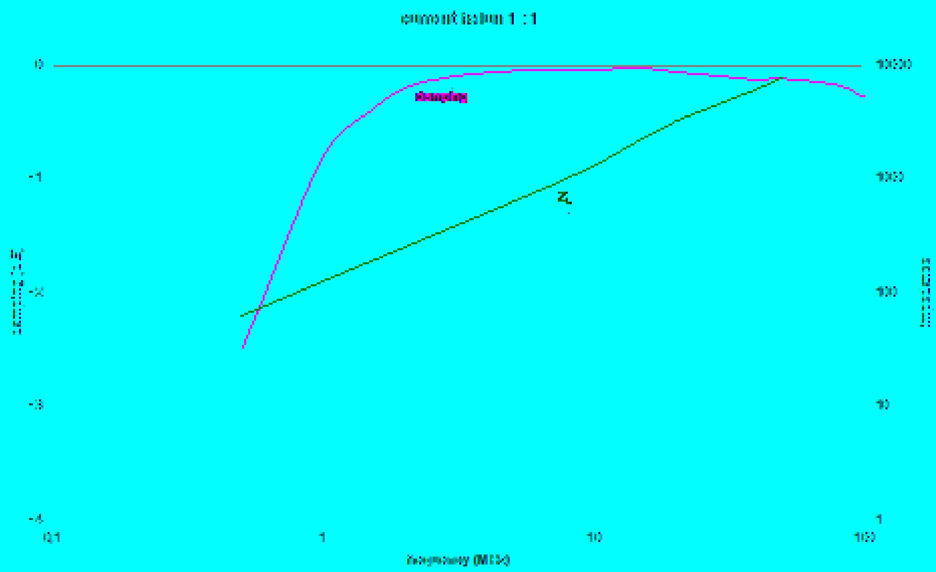

Figure 6. Current balun transfer plus sleeve impedance

In figure 6 we find this current balun already better performing than the voltage balun at 2 MHz. with 0,2 dB of damping versus 0,25 dB. In the rest of the HF frequency range, insertion loss is much lower for the current balun again with a damping of 0,1 dB at 30 MHz. as compared to 1,7 - 2,5 dB for the voltage balun. Also beyond this frequency, the current balun is still in full service for quite some time.

Impedance to ground

Using this component in a symmetrical tuner, the impedance to ground is especially important. Although permeability of 4C65 ferrite is going down at higher frequencies, the vector summation of permeability and materials-loss keeps on rising for a very long frequency range so total sleeve impedance will keep going up. This is opposite the impedance in the voltage transformer that is determined by the parasitic inter-winding capacitance, generating a falling impedance graph.

Starting at 10 MHz. the current balun is close to the minimum impedance requirement (Z > 1500 Ohm). At lower frequencies impedance is still too low. This is easily enhanced by adding ferrite to the core. For each additional ferrite ring, impedance is added by the same factor. With three toroides in a stack and the same ten turns of RG58, we easily reach to over 1000 Ohm at 2 MHz. which is close enough for this lower operational frequency.

Stacking cores is the better method over constructing three individual ten turn current baluns and putting these in series. In this latter situation, parasitic capacitance of one sleeve coil will inevitably resonate with the inductance of the next transformer, creating a low impedance instead.

A last but useful feature of the current balun over the voltage balun is the direct DC-path of the first type to ground; static charge may not build up making this component safer to operate and more 'silent' in dry seasons (no discharge noise).

Before or after the tuner?

A frequently recurring discussion to put the balun before of behind the tuner has found a definitive answer in the article by Roy Lewallen: 'The 1 : 1 current balun' . He proves that there is no difference in operation and the balun will have to exhibit an equally high 'sleeve impedance' in both situations.

Bob J. van Donselaar,
mailto:ON9CVD@AMSAT.org