A Publication for the
Radio Amateur Worldwide

Especially Covering VHF, UHF
and Microwaves

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13cm FM-ATV Exciter

Reiner Erping & Wolfgang Schneider

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A DIY Receiver for GPS and GLONASS Satellites Part-2

2.3. GPS & GLONASS Satellite Systems

GPS and GLONASS are the first satellite systems that require the simultaneous operation of a number of satellites. In other satellite systems, including earlier navigation systems, the operation of every single satellite was almost autonomous and any additional satellites only improved the capacity of the system.

In GPS or GLONASS the satellites need to be synchronised and can only perform as a constellation of at least four visible satellites for every possible user location without forgetting the GDOP requirement! Both GPS and GLONASS satellites are launched into similar orbits. A comparison among GPS, GLONASS and more popular satellite orbits like the geostationary orbit or the retrograde sun-synchronous Low-Earth Orbit (LEO) is made on the scale drawing on Fig.3. Both GPS and GLONASS satellites are launched into circular orbits with the inclination ranging between 55 and 65 degrees and the orbital period in the order of 12 hours, which corresponds to an altitude of around 20000km (one and a half Earth diameters).

The GPS system was initially planned to use three different orbital planes with an inclination of 63 degrees and the ascending nodes equally spaced at 120 degrees around the equator. Each orbital plane would accommodate 8 equally spaced satellites with an orbital period of 11 hours and 58 minutes, synchronised with the Earth's rotation rate [4].

During a 10-year test period from 1978 to 1988 only 10 such "Block I" satellites were successfully launched in orbital planes "A" and "C" as shown in Fig.4. The GPS specification was changed afterwards [5] and the new "Block II" satellites are being launched in 55-degrees inclination orbits in six different orbital planes A, B, C, D E and F, with the ascending nodes equally spaced at 60 degrees around the equator. The new GPS constellation should also include 24 satellites, having four satellites in each orbital plane, including active in-orbit spares. The orbital period of the GPS satellites should be increased to 12 hours to avoid repeat-track orbits and resonances with the Earth's gravity field. Finally, the new "Block II" satellites also include a nasty feature called "Selective Availability" (SA): the on-board hardware may, on ground command, intentionally degrade the accuracy of the navigation signals for civilian users while military users still have access to the full system accuracy.

Beginning in 1988 and up to March 1993, 9 GPS "Block II" and 10 new GPS "Block IIA" satellites have been launched using "Delta" rockets. The SA-mode is currently turned on and degrades the accuracy to between 50 and 1O0M.

The GLONASS system is planned to use three different orbital planes with an inclination of 64.8 degrees and the ascending nodes equally spaced at 120 degrees around the equator. Each orbital plane would accommodate 8 (or 12) equally spaced satellites with an orbital period of 11 hours, 15 minutes and 44 seconds, so that each satellite repeats its ground track after exactly 17 revolutions or 8 days [6].

Satellite	**Launch**	**Orbit**	**PRN#**	**Decommissioned**
GPS BI-01	78 20 A	C-?	4	Jul 85
GPS BI-02	78 47 A	A-?	7	Jul 81
GPS BI-03	78 93 A	A-?	6	May 92
GPS BI-04	78112 A	C-?	8	Oct 89
GPS BI-05	80 11A	C-?	5	*Nov* 83
GPS BI-06	80 32 A	A-?	9	Mar 91
GPS BI-07	Launch failure			
GPS BI-08	83 72 A	C-3	11	May 93
GPS BI-09	84 59 A	C-1	13	
GPS BI-10	84 97 A	A-1	12	
GPS BI-11	85 93 A	C-4	3	
GPS BII-01	89 13 A	E-1	14	
GPS BII-02	89 44 A	B-3	2	
GPS BII-03	89 64 A	E-3	16	
GPS BII-04	89 85 A	A-4	19	
GPS BII-05	89 97 A	D-3	17	
GPS BII-06	90 8 A	F-3	18	
GPS BII-07	90 25 A	B-2	20	
GPS BII-08	90 68 A	E-2	21	
GPS BII-09	90 88 A	D-2	15	
GPS BIIA-10	90103 A	E-4	23	
GPS BIIA-11	9147 A	D-1	24	
GPS BIIA-12	92 9 A	A-2	25	
GPS BIIA-13	92 19 A	C-2	28	
GPS BIIA-14	92 39 A	F-2	26	
GPS BIIA-15	92 58 A	A-3	27	
GPS BIIA-16	92 79 A	F-I	1	was #32
GPS BIIA-17	92 89 A	F-4	29	
GPS BIIA-18	93 7 A	B-1	22	
GPS BIIA-19	93 17 A	C-1	31	
GPS BIIA-20	93 32 A	C-4	7	
GPS BIIA-21	93 42 A	A-1	9	
GPS BIIA-22	93 54 A	B-4	5	

Fig. 4: Published GPS Satellite Operation

Satellite	Launch	Orbit	CHN#	Decommissioned
Glonass 34	88 43A	1-8	?	?
Glonass 36	88 43C	1-1	24	?
Glonass 39	88 85C	3-18	10	Jan 92
Glonass 40	89 1A	1-2	9	Replaced Mar 93
Glonass 41	89 1B	1-3	6	Replaced Feb 92
Glonass 44	90 45A	3-17	24	Formerly #21
Glonass 45	90 45B	3-19	3	Mar 93
Glonass 46	90 45C	3-20	15	Sep 92
Glonass 47	90110A	1-4	4	
Glonass 48	90110B	1-7	13	
Glonass 49	90110C	1-5	23	Formerly #19
Glonass 50	9125A	3-21	20	Jan 92
Glonass 51	9125B	3-22	11	
Glonass 52	9125C	3-24	14	Feb 92
Glonass 53	92 5A	1-3	22	Jan 93
Glonass 54	92 5B	1-8	2	
Glonass 55	92 5C	1-1	23	Formerly #17
Glonass 56	92 47A	3-24	1	
Glonass 57	92 47B	3-21	24	Formerly 3-18
Glonass 58	92 47C	3-20	8	Formerly 3-21
Glonass 59	93 10A	1-3	12	
Glonass 60	93 10B	1-2	5	
Glonass 61	93 10C	1-6	22	Formerly #23

Fig. 5: Recently observed GLONASS Satellite Operation

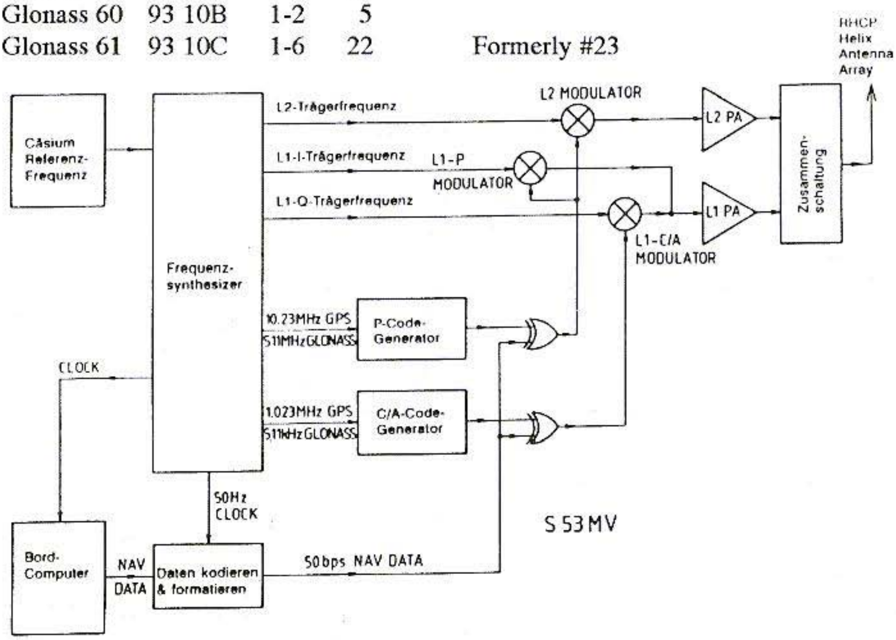

Fig.6: Block Diagram of GPS and GLONASS Satellites

Since the beginning of the GLONASS program a large number of satellites have been launched into GLONASS orbital planes 1 and 3, the orbital plane 2 has not been used yet. Some satellites never transmitted any radio signals, since the GLONASS system also includes passive "Etalon" satellites used as optical reflectors for accurate orbit determination.

GLONASS satellites are launched three at a time with a single "Proton" rocket. Due to this constraint all three satellites can only be launched in the same orbital plane. Recently observed GLONASS satellite operation is shown on Fig.5. The observed lifetime of GLONASS satellites seems to be shorter than that of American GPS counterparts.

A critical piece of on-board equipment are the atomic clocks required for system synchronisation. Although each satellite carries redundant rubidium and caesium clocks, these caused the failure of many GPS and GLONASS satellites. In addition to this, GLONASS satellites have had problems with the on-board computer. Unfortunately, the GPS or GLONASS orbit altitude is actually in the worst ionising-radiation zone, the same radiation that already destroyed the AMSAT-OSCAR-10 computer memory.

2.4. GPS & GLONASS Satellite On-board Equipment

Since the two systems are similar, GPS and GLONASS satellites carry almost the same on-board equipment as shown in Fig.6.

For the navigation function alone, the satellites could be much simpler, carrying a simple linear transponder like on civilian communications satellites. The required navigation signals could be generated and synchronised by a network of ground stations. However, both GPS and GLONASS are primarily intended as military systems.

Uplinks are undesired since they can be easily jammed and a network of ground stations can be easily destroyed. Therefore, both GPS and GLONASS satellites are designed for completely autonomous operation and generation of the required signals. Synchronisation is maintained by on-board atomic clocks that are only periodically updated by the ground stations.

Both GPS and GLONASS satellites carry a caesium atomic clock as their primary time/frequency standard, with the accuracy ranging between 10-12 and 10-13.

Much smaller and lightweight rubidium atomic clocks are used as a backup in the case the main time/frequency standard fails, although rubidium atomic clocks are an order of magnitude less accurate. Due to the stable space environment these atomic clocks usually perform better than their ground-based counterparts and any long-term drifts or offsets can be easily compensated by uploading the required correction coefficients in the on-board computer.

The output of the atomic time/ frequency standard drives a frequency synthesiser so that all the carrier frequencies and modulation rates are derived coherently from the same reference frequency. The on-board computer generates the so-called navigation data. These include information about the exact location of the satellite, also called precision ephemeris, information about the offset and drift of the on-board atomic clock and information about other satellites in the system, also called almanac. The first two are used directly by the user's computer to assemble the navigation equations.

The almanac data can be used to predict visible satellites and avoid attempting to use dead, malfunctioning or non-existent satellites, thus speeding-up the acquisition of four valid satellite signals with a reasonable GDOP. Besides the transmitters for broadcasting navigation signals, GPS and GLONASS satellites also have telecommand and telemetry radio links.

In particular, the telecommand link is used by the command stations to

Satellite Channel	L1 - Carrier	L2 - Carrier
GPS (all Satellites)	1575.420 MHz	1227.600 MHz
GLONASS Channel 0	1602.000 MHz	1246.000 MHz
GLONASS Channel 1	1602.5625 MHz	1246.4375 MHz
GLONASS Channel 2	1603.125 MHz	1246.875 MHz
GLONASS Channel 3	1603.6785 MHz	1247.3125 MHz
GLONASS Channel 4	1604.250 MHz	1247.750 MHz
GLONASS Channel 5	1604.8125 MHz	1248.1875 MHz
GLONASS Channel 6	1605.375 MHz	1248.625 MHz
GLONASS Channel 7	1605.9375 MHz	1249.0625 MHz
GLONASS Channel 8	1606.500 MHz	1249.500 MHz
GLONASS Channel 9	1607.0625 MHz	1249.9375 MHz
G LONASS Channel 10	1607.625 MHz	1250.375 MHz
GLONASS Channel 11	1608.1875 MHz	1250.8125 MHz
GLONASS Channel 12	1608.750 MHz	1251.250 MHz
GLONASS Channel 13	1609.3125 MHz	1251.6875 MHz
GLONASS Channel 14	1609.875 MHz	1252.125 MHz
GLONASS Channel 15	1610.4375 MHz	1252.5625 MHz
GLONASS Channel 16	1611.000 MHz	1253.000 MHz
GLONASS Channel 17	1611.5625 MHz	1253.4375 MHz
GLONASS Channel 18	1612.125 MHz	1253.875 MHz
GLONASS Channel 19	1612.6785 MHz	1254.3125 MHz
GLONASS Channel 20	1613.250 MHz	1254.750 MHz
GLONASS Channel 21	1613.8125 MHz	1255.1875 MHz
GLONASS Channel 22	1614.375 MHz	1255.625 MHz
GLONASS Channel 23	1614.9375 MHz	1256.0625 MHz
GLONASS Channel 24	1615.500 MHz	1256.500 MHz

Fig.7: Carrier Frequencies for GPS and GLONASS Satellites

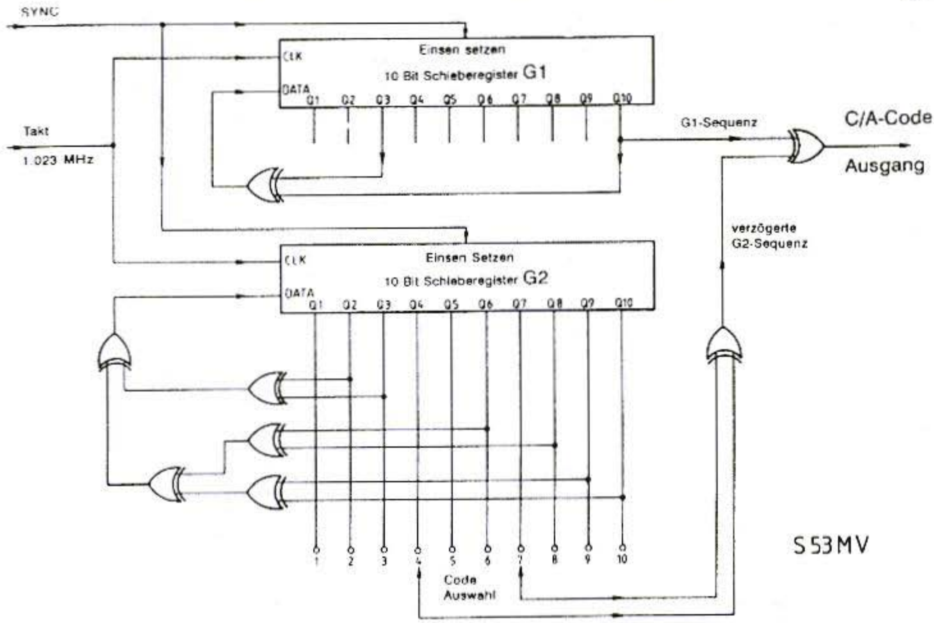

Fig.8: The GPS C/A Code Generator

regularly upload fresh navigation data into the on-board computer. Usually this is done once per day, although the on-board computer memory can store enough data for several weeks in advance.

In addition to dedicated telemetry links, part of the telemetry data is also inserted in the navigation data stream.

2.5. GPS & GLONASS Satellite Transmissions

GPS and GLONASS satellites use the microwave L-band to broadcast three separate radio-navigation signals on two separate RF channels usually called Ll (around 1.6 GHz) and L2 (around 1.2 GHz)- These frequencies were chosen as a compromise between the required satellite transmitter power and ionospheric errors. The influence of the ionosphere decreases with the square of the carrier frequency and is very small above 1 GHz.

However, in a precision navigation system it still induces a position error of about 50m at the L1 frequency during daylight and medium solar activity. On the other hand, GPS and GLONASS were designed to work with omnidirectional, hemispherical-coverage receiving antennas. The capture area of an antenna with a defined radiation pattern decreases with the square of the operating frequency, so the power of the on-board transmitter has to be increased by the same amount.

Both GPS and GLONASS broadcast two different signals: a Coarse/ Acquisition (C/A) signal and Precision

(P) signal. The C/A-signal is only transmitted on the higher frequency (LI) while the P-signal is transmitted on two widely-separated RF channels (LI and L2).

Since the frequency dependence of ionospheric errors is known, the absolute error on each carrier frequency can be computed from the measured difference between the two P-transmissions on LI and L2 carriers.

The LI C/A- and P-carriers are in quadrature to enable a single power amplifier to be used for both signals, as shown on Fig.6. The LI and L2 transmitter outputs are combined in a passive network and feed an array of helix antennas. These produce a shaped beam covering the whole visible hemisphere from the GPS/GLONASS orbit with the same signal strength.

All three GPS or GLONASS transmissions are continuous, straightforward BPSK modulated carriers. Pulse modulation is not used. The timing information is transmitted in the modulation: the user's receiver measures the time of arrival of a defined bit pattern, which is a known code. If desired, the modulation code phase can be related to the carrier phase in the receiver to produce even more accurate measurements, since both the carrier frequency and the code rate are derived coherently from the same reference frequency on-board the satellite.

The GPS C/A-code is 1023 bits long and is transmitted at 1.023 Mbps. The C/A-code repetition period is therefore 1 ms. The GLONASS C/A-code is 511 bits long and is transmitted at 511 kbps, so it has the same repetition period as the GPS C/A-code. The P-code is transmitted at 10 times the speed of the C/A-code: 10.23 Mbps for GPS and 5.11 Mbps for GLONASS. The transmitter power level for the P-code on L1 is 3dB below the LI C/A-code and the P-code on L2 is 6dB below the LI C/A-code. The P-code repetition period is very long, making an autonomous search for synchronisation not practical. All P-code receivers first acquire lock on the C/A-transmission, which also carries information that allows a quick P-code lock. Both C/A- and P-codes are generated by digital shift-registers with the feedback selected to obtain pseudo-random codes. The navigation data is modulo-2 added to the pseudo-random codes. Since the navigation-data rate is very low, only 50 bps, it does not affect significantly the randomness properties of the codes used.

The navigation data at 50 bps is synchronised to the C/A-code period to resolve the timing ambiguity caused by the relatively short 1 ms C/A-code repetition period. GPS "Block II" satellites may encrypt the published P-code into the secret Y-code. This process is called "Anti-Spoofing" (AS). Its purpose is to prevent an enemy from jamming the GPS with false GPS-like signals.

Details of the .GLONASS P-code are not published. In fact, the GLONASS P-code is even not mentioned in [6], although these transmissions can be easily observed on a spectrum analyser. The GPS and GLONASS RF channel carrier frequencies are shown on Fig.7.

C/A Code Number	G2 Register Taps	G2 Clock Cycles
GPS PRN 1	2 & 6	5 clks
GPS PRN 2	3 & 7	6 clks
GPS PRN 3	4 & 8	7 clks
GPS PRN 4	5 & 9	8 clks
GPS PRN 5	1 & 9	17 clks
GPS PRN 6	2 & 10	18 clks
GPS PRN 7	1 & 8	139 clks
GPS PRN 8	2 & 9	140 clks
GPS PRN 9	3 & 10	141 clks
GPS PRN 10	2 & 3	251 clks
GPS PRN 11	3 & 4	252 clks
GPS PRN 12	5 & 6	254 clks
GPS PRN 13	6 & 7	255 clks
GPS PRN 14	7 & 8	256 clks
GPS PRN 15	8 & 9	257 clks
GPS PRN 16	9 & 10	258 clks
GPS PRN 17	1 & 4	469 clks
GPS PRN 18	2 & 5	470 clks
GPS PRN 19	3 & 6	471 clks
GPS PRN 20	4 & 7	472 clks
GPS PRN 21	5 & 8	473 clks
GPS PRN 22	6 & 9	474 clks
GPS PRN 23	1& 3	509 clks
GPS PRN 24	4 & 6	512 clks
GPS PRN 25	5 & 7	513 clks
GPS PRN 26	6 & 8	514 clks
GPS PRN 27	7 & 9	515 clks
GPS PRN 28	8 & 10	516 clks
GPS PRN 29	1 & 6	859 clks
GPS PRN 30	2 & 7	860 clks
GPS PRN 31	3 & 8	861 clks
GPS PRN 32	4 & 9	862 clks

Fig.9: GPS C/A Codes and their correlation with the Registers

All GPS satellites transmit on the same L1 and L2 carrier frequencies: 1575.42 MHz and 1227.6 MHz, which are held in the exact ratio 77/60 and are integer multiples of the fundamental GPS clock frequency of 10.23 MHz.

Every GPS satellite transmits its own set of C/A- and P-codes that have good cross-correlation properties with the codes used by other GPS satellites. Since a GPS receiving antenna is omnidirectional and receives many satellites at the same time, the receiver is using Code-Division Multiple Access (CDMA) techniques to separate signals coming from different satellites.

GPS satellites are therefore identified by the Pseudo-Random-Noise code number (PRN#). The GLONASS satellites use 25 different RF channels. Channel 0 is reserved for testing spare satellites while channels 1 to 24 are dedicated to operational GLONASS satellites.

All GLONASS satellites transmit the same C/A-code and are usually identified by the Channel Number (CHN#)The L1 and L2 carrier frequencies are in the exact ratio 9/7 and the channel spacing is 562.5 kHz at L1 and 437.5 kHz at L2. Although there exist civilian P-code receivers, the majority of civilian GPS or GLONASS receivers are C/A-only receivers. Since the

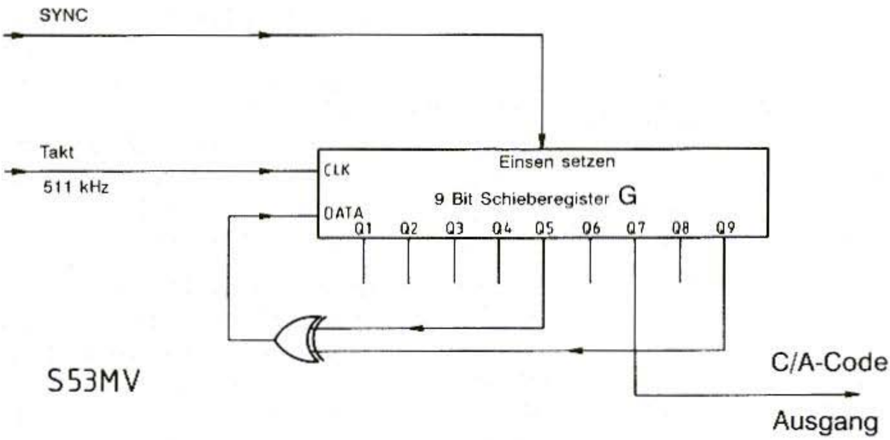

Fig.10: The GLONASS C/A Code Generator

advantages of using the P-code are limited, especially with SA, AS or both active, only the C/A-code transmission will be discussed in detail here.

2.6. GPS C/A-Transmission Format

GPS satellites use code-division multiplexing on both C/A- and P-transmissions. Since C/A-codes are relatively short sequences (only 1023 bits), the codes have to be carefully selected for good cross-correlation properties. GPS C/A-codes are Gold codes (named after their inventor Robert Gold) that can be generated as a modulo-2 sum of two maximum-length shift-register sequences. The GPS C/A-code generator is shown on Fig.8.

It includes two 10-bit shift registers Gl and G2, both clocked at 1.023 MHz, each with a separate feedback network made of exclusive-or gates. Both feedback networks are selected so that both generated sequences have the maximal length of 1023 bits. Both shift registers are started in the "all-ones" state and since both sequences have the same length, the shift registers maintain the synchronisation throughout the operation of the circuit.

Gold codes are obtained by a modulo-2 sum (another exclusive-or operation) of the outputs of the two shift registers Gl and G2. Different codes can be obtained by changing the relative phase of the two shift registers. Instead of desynchronising the shift registers it is easier to delay the output of one of them (G2). This variable delay is achieved with yet another modulo-2 sum (exclusive-or) of two G2 register taps.

Exclusive-or feedback shift-register sequences have the property that a modulo-2 addition of a sequence with its delayed replica produces the same sequence, but delayed by a different number of clock cycles. Choosing two G2 register taps, 45 different delays can be generated yielding 45 different Gold codes with good auto-correlation and cross-correlation properties. Out of these 45 possible codes, 32 are allocated to GPS satellites as shown on Fig.9.

The cross-correlation properties of GPS C/A-codes guarantee a cross-talk smaller than -21.6dB between the desired and undesired satellite signals. The 50bps navigation data stream is synchronised with the C/A-code generator so that bit transitions coincide with the if all-ones" state of both shift registers Gl and G2. At 50bps one data bit corresponds to 20 C/A-code periods.

The navigation data is formatted into words, subframes and frames. Words are 30 bits long including 24 data bits and 6 parity bits computed over the 24 data bits and the last two bits of the previous word. Parity bits are used to check the received data for errors and to resolve the polarity ambiguity of the BPSK demodulator. 10 words (300 bits) form a subframe which always includes a subframe sync pattern "10001011" and a time code called "Time-Of-Week" (TOW). One subframe is transmitted every 6 seconds. Five subframes form one frame (1500 bits) that contains all of the information required to use the navigation signals. One frame is transmitted every 30 seconds.

The first subframe in the frame contains the on-board clock data: offset, drift etc. The second and third subframes contain the precision ephemeris data in the form of keplerian elements with several correction coefficients to accurately describe the satellite's orbit. Finally, the fourth and fifth subframes contain almanac data that is not required immediately and are subcommutated in 25 consecutive frames, so that the whole almanac is transmitted in 12.5 minutes.

The allocation of the single data words is completely described in [5]. Most numerical parameters are 8-, 16-, 24- or 32-bit integers, either unsigned or signed in the two's complement format. Angular values that can range from 0 to 360 degrees are usually expressed in semi-circles to make better use of the available bits. GPS is also using its own time scale. The units are seconds and weeks. one week has 604800 seconds and the week count is incremented between Saturday and Sunday. GPS time starts on the midnight of January 5/6, 1980.

GPS time is a continuous time and therefore it differs by an integer number of leap seconds from UTC. The difference between UTC and GPS time is included in the almanac message.

2.7. GLONASS C/A-Transmission Format

GLONASS satellites use the more conventional frequency-division multiplexing at least for the C/A-code transmissions. All GLONASS satellites use the same C/A-code, generated by a 9-bit shift register G as shown on Fig.10. The GLONASS C/A-code is a maximum-length sequence and thus has an ideal auto-correlation function.

Frequency-division multiplexing allows a better channel separation than code-division multiplexing. The separation between two adjacent GLONASS channels should be better than -48dB. A large channel separation is useful when the signal from one satellite is much weaker because of reflected waves and/or holes in the receiving antenna radiation pattern.

On the other hand, the GLONASS satellites require a wider RF spectrum and a GLONASS C/A-receiver is necessarily more complex than a GPS C/A-receiver.

The GLONASS navigation data stream is synchronised with the C/A-code generator so that level transitions coincide with the "all-ones" state of the shift register. The navigation data stream is formatted into lines of the duration of 2 seconds. Each line contains 85 information bits, transmitted at 50bps for 1.7 seconds and a "time mark" sync pattern "111110001101110101000010010110", which is a pseudo-random sequence of 30 bits transmitted at 100bps for the remaining 0.3 seconds.

The 85 information data bits always start with a leading "0" , followed by 76 bits containing navigation information and 8 parity-checking bits, computed according to the (85, 77) Hamming code. After computing the parity bits, all of the 85 bits are differentially encoded to resolve the phase ambiguity in the receiver.

Finally, the 85 differentially-encoded bits are Manchester encoded, so that a "10" pattern corresponds to a logical "one" and a "01" pattern corresponds to a logical "zero". The additional transition in the middle of the data bits introduced by the Manchester encoding speeds-up the synchronisation of the receiver. 15 navigation data lines form one frame of the duration of 30 seconds. The allocation of the single data bits in the frame is completely described in [6]. The first four lines of a frame contain the time code, on-board clock offset and drift and precision ephemeris data of the satellite orbit in the form of a state vector (position vector and velocity vector). To simplify the computations in the user's receiver, the corrections for the Sun- and Moon-gravity forces are also supplied. The almanac data is transmitted in the remaining 11 lines of the frame.

Almanac satellite ephemeris is in the form of keplerian elements and is transmitted in two consecutive lines in a frame. The whole almanac is transmitted in five consecutive frames also called a superframe of the duration of 2.5 minutes. The various numerical parameters are transmitted as different size, either unsigned or signed integers. Signed integers are transmitted in the form of a sign bit followed by an unsigned integer representing the absolute value of the number (this is different from the two's complement notation!). Angular values are usually expressed in semi-circles.

The GLONASS time is kept synchronised to UTC. GLONASS uses more conventional time units like days, hours, minutes and seconds. The day count begins with a leap year (currently 1992) and counts up to 1461 days before returning back to zero.

(To be continued)

(References overleaf)

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Richard A. Formato, Ph.D., K1POO

Improving Impedance Performance of the Extended Double-Zepp Antenna

The Extended Double Zepp (XDZ) has been a popular amateur antenna since the early days of shortwave radio. It can be used from HF well into the UHF range, where it is especially attractive because shorter wavelength offsets the XDZ's increased physical length. The XDZ provides exceptional gain for a simple antenna, but its impedance properties can create matching problems [1]. This note describes a simple technique for controlling and improving the XDZ's impedance performance.

1. THE XZD ANTENNA

XDZ geometry is shown in Fig.1. The antenna consists of two collinear, end-fed, electrically long radiating elements ("Zepp" elements). The element spacing at the feed is S, the diameter D, and the overall physical antenna length is L. If the length-to-diameter ratio is large (L/D > 1), the antenna is "thin"; otherwise it is "fat". The electrical length of the original Zepp element is half-wave, but an XDZ element is somewhat longer, approximately 0.64 wave. When the spacing S small (S/L < 1), the XDZ is essentially a long centre-fed dipole, and its performance is accurately analysed using a dipole model. It is assumed that S/L < 1.

The reason for the XDZ's popularity is apparent from an examination of Fig.2, which plots the directivity of a free-space centre-fed dipole vs. end-to-end electrical length. The widely used thin, half-wave dipole provides 2.15 dBi gain (dB relative to an isotropic radiator) with a well-behaved input impedance of approximately $77+j44\Omega$. But longer dipoles provide much better performance. As the length increases, the directivity also increases, reaching a

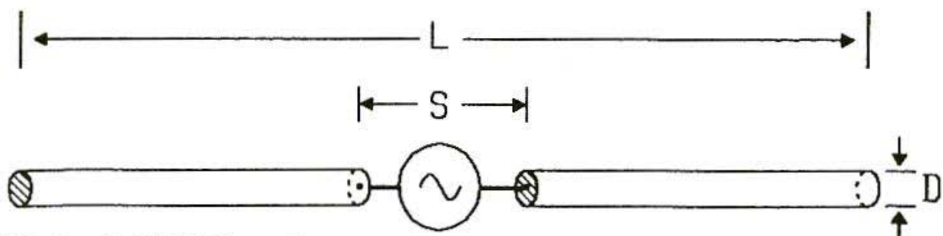

Fig.1: Fat XDZ Geometry

maximum of nearly 5.2 dBi at a length of 1.27 waves. This electrical length is optimum for the XDZ. More than 3 dB gain over a half-wave dipole is obtained by simply making the antenna longer.

The pattern factor (normalised radiation pattern) for the 1.27-wave XDZ is shown in Fig.3. Three lobes appear because the antenna is electrically long. The main lobe, with a maximum gain of 5.2 dBi, is oriented broadside to the antenna axis. Its -3 dB beamwidth is 31.5 degrees. By comparison, the half-power beamwidths for half and full-wave dipoles, respectively, are 78 and 47 degrees. The XDZ's two sidelobes are almost 10 dB down, and, for practical purposes, can be ignored.

Note that Figs. 2 and 3 are based on an ideal thin radiator (infinite L/D ratio) having a sinusoidal current distribution.

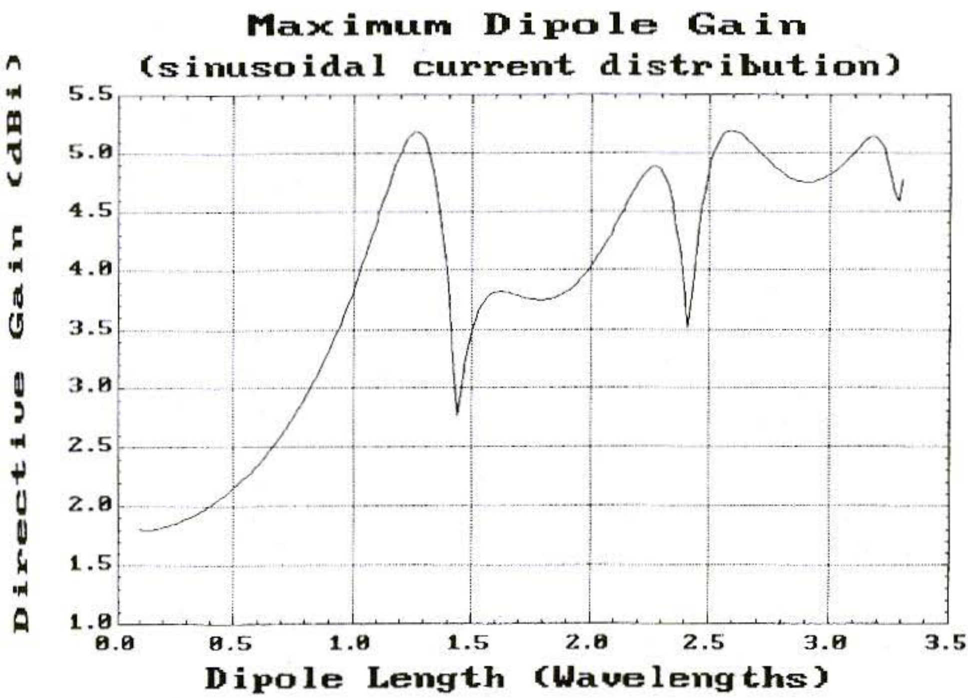

Fig.2: Directivity of a Free-Space Centre-Fed Dipole vs Electrical Length

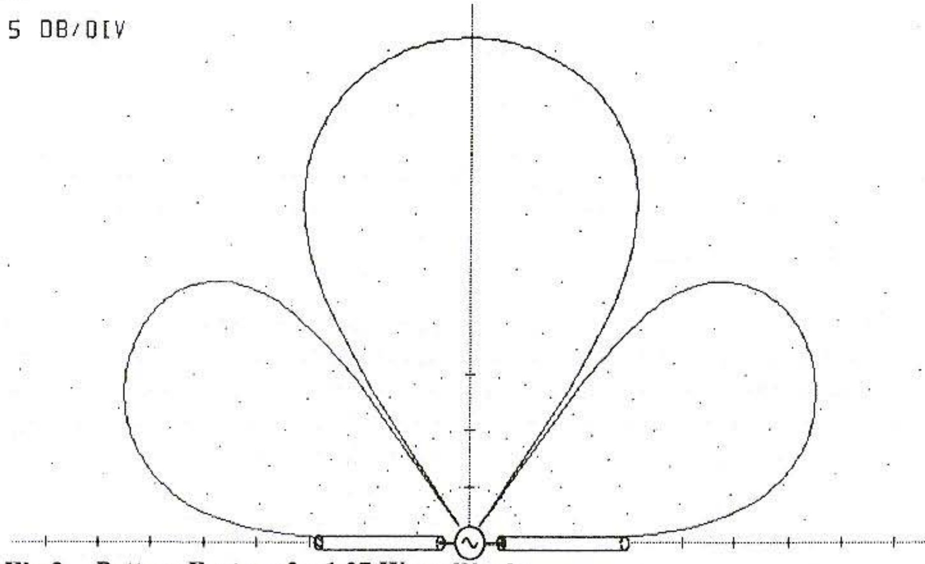

Fig.3: Pattern Factor of a 1.27 Wave Dipole

This approximation provides accurate directive gain and the general pattern shape even for "fat" radiators (small values of L/D). The major pattern effect for a fat element is that the nulls begin to "wash out". The sinusoidal current approximation, however, is not accurate for impedance calculations, especially for fat elements.

Considering only the data in Fig's.2 and 3, the XDZ might seem to be the ideal antenna, one that provides excellent gain in a very simple, easy-to-build structure. Unfortunately, the XDZ has a serious drawback. Because it is a full-wave antenna, its input impedance is high, possibly thousands of ohms, which can create matching problems. An XDZ with more moderate input impedance would be a better antenna. Fortunately, there is a simple solution to this problem, and it lies in choosing the optimum XDZ L/D ratio.

Fig's.4 and 5 plot dipole input resistance and reactance vs. electrical length. At its full-wave resonance (X=0), a thin dipole (L/D=5000) has a very high resistance (approximately 1,800Ω) . In contrast, "fatter" elements (smaller L/D ratios) exhibit more moderate impedance levels. By properly choosing L/D, the XDZ input impedance can be controlled while still achieving maximum directivity from its increased electrical length. The data in Fig's.4 and 5 are based on a non-sinusoidal current distribution for improved accuracy at small L/D values.

The optimum L/D ratio for a 50Ω feed is 30.5, since this value results in a driving point impedance of 50-*j*123Ω for a 1.27 wave XDZ. These theoretical values provide a starting point for an improved XDZ design. The only matching required is an inductor to tune out the 123Ω capacitive reactance. At most frequencies, the matching

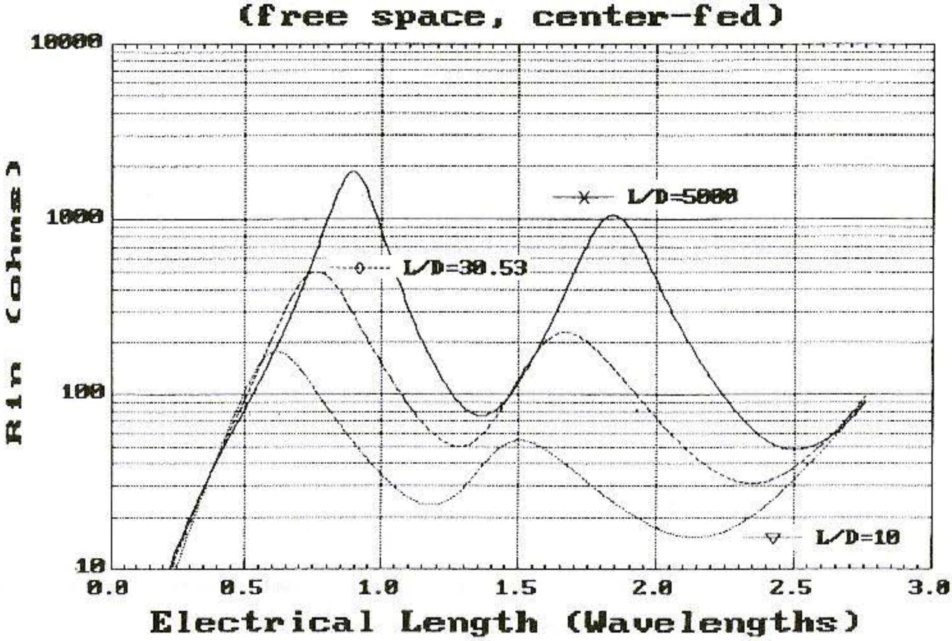

Fig.4: Dipole Input Resistance

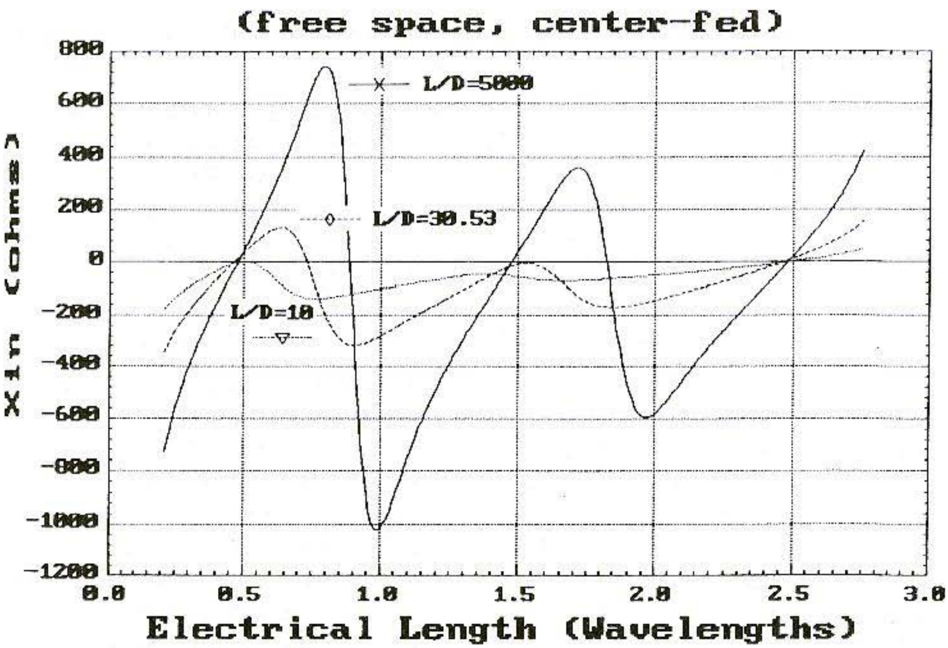

Fig.5: Dipole Input Reactance

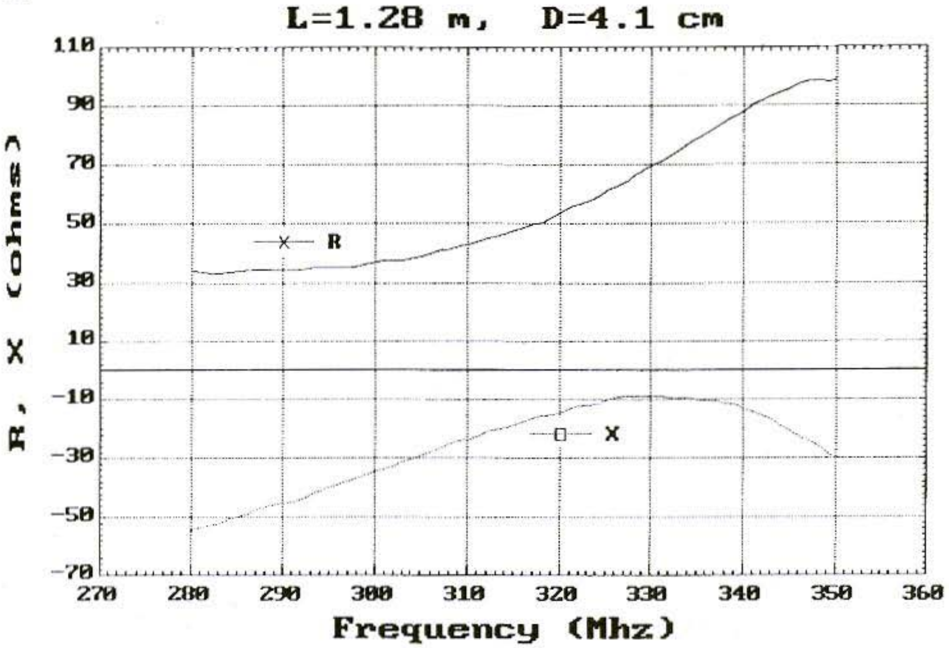

Fig.6: Measured XDZ Input Impedance

inductance is small. At frequencies in the high VHF-UHF ranges, the feed system may well contain enough stray inductance to virtually eliminate the need for adding any.

A simple "plumber's delight" dipole was built and measured as a crude validation of this technique. Even though no effort was made to perform a controlled experiment, the data clearly illustrates the viability of this approach.

The test antenna consisted of two 24-3/4" x 1-5/8" O.D. copper tubes separated 1" at the feed point. These radiating elements were strapped to a 12"x1"x3/8" Plexiglas support using 4 nylon cable ties on each element. A female type N chassis connector was soldered to the elements using straight #14 AWG solid copper wire pigtails (no balun was used, although normally one would be).

The antenna was mounted vertically in a 10" diameter pine tree about 16" from the trunk. The feed point was approximately 8' above the ground. The RG-8 coax feed cable was tied horizontally along a branch for a distance of about 4' from the antenna feed, then dropped to the ground.

Measured values of R and X appear in Fig.6. The test XDZ was approximately 1.27 wavelengths long at 300 MHz, where the input impedance was 36.7-$j35\Omega$. Without matching, the corresponding VSWR is 2.34:1 (0.76dB mismatch loss). By adding inductance to tune out the -35 ohms reactance, the VSWR could be reduced to 1.36:1 (0.1dB loss).

Without matching, the minimum measured VSWR was 1.3:1 (0.07 dB loss) at 322 MHz (I/P impedance $56.3\text{-}j12.4\Omega$). These moderate values of R and X show how effective L/D can be in controlling XDZ input impedance.

The bandwidth of a "fat" XDZ is also surprisingly good because the input impedance varies gradually with frequency. VSWRs less than 2.5:1 are achievable over more than 10% of the design frequency, which is enough bandwidth to cover most amateur bands with one antenna without matching.

The XDZ's directivity, however, falls off quickly on either side of the frequency at which the antenna is 1.27 waves long. Nevertheless the gain is still better than that of a half-wave dipole.

Designing an optimum XDZ thus consists of three steps:

❶ Choosing the electrical length to provide the desired gain.

❷ Choosing L/D ratio to achieve the desired antenna input resistance.

❸ Adding components at the feed to tune out any reactance at the design frequency*.

* *Inductance or capacitance should be added symmetrically to the radiating elements. For example, if a total of 1μH is required, then 0.5μH should be added in series with each radiating element to maintain the XDZ's electrical balance as a symmetrical radiating system.*

The plots in this article provide a starting point for these design steps, but design details will vary depending on the specific antenna. For example, resistance and reactance introduced at the feed point will modify the XDZ input impedance. Such effects are difficult, if not impossible, to predict in advance because they depend on exactly how the feed is built. As with any antenna design, some "tuning" will be necessary after the basic system is fabricated.

XDZ implementations using solid copper or aluminium tubing are feasible at VHF/UHF. But at HF the element diameter needed to obtain the desired L/D ratio is too large for tubing. The "cage" structures described in [1], Ch. 9, can be used instead of a large diameter conductor. As a general rule, the radiating element should consist of at least 8 wires parallel to the element axis to adequately simulate a continuous conducting surface.

The technique of varying L/D to control antenna input impedance and impedance bandwidth is not restricted to the XDZ. Similar considerations apply to monopoles on ground planes, active and parasitic arrays of Zepp elements, and, in fact, any wire antenna structure. Input impedance and impedance bandwidth can be significantly modified by changing L/D. Of course, the optimum L/D ratio depends on the specific antenna geometry and the design objective; there is no universally "best" value.

This article illustrates how useful the L/D ratio can be in designing wire antennas. It is hoped that this information will encourage experimentation with easily constructed antennas, in particular dipoles and monopoles.

Simply changing the element diameter often produces a much better antenna!

2. LITERATURE

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Angel Vilaseca HB9SLV and Serge Riviere F1JSR

A Solid State Broadband 80W Amplifier for 24cm

This circuit produces 80 Watts of RF on the 23cm band, with 4 Watts of drive. It is made with four coupled hybrid amplifiers M 57762 from Mitsubishi.

1. DESCRIPTION

One of the most interesting points about the unit is that it can be used for mobile or portable use, because it only needs a 12 volt power supply. It is rather compact and can be fitted into a standard 19" rack. It weighs about 10kg with its switching power supply, so it can advantageously replace a 2C39 tube amplifier. For the owners of QRO tube amplifiers (F 6007, TH328, 338, ..., types), this circuit can be used as a driver. Unlike tube amplifiers, this amplifier is suitable for all modes. It is broadband, which allows for ATV use, and does not drift, which means that no retuning is required during operation. The broadband qualities of the individual amplifiers are retained in the global circuit, because we used broadband power splitters.

Multi-brick amplifiers have been described before, using splitters made up of four coax stubs. Although much cheaper, such an approach requires extreme care to be used in the realisation to avoid unbalances. Unbalance produces losses, thus decreasing both gain and power output capabilities.

Furthermore, a coupler made from coax has a much narrower bandwidth than with hybrid couplers, which means that the broadband capability of the hybrid amplifiers is also lost in the global amplifier. This can be a disadvantage for FM ATV operation for example.

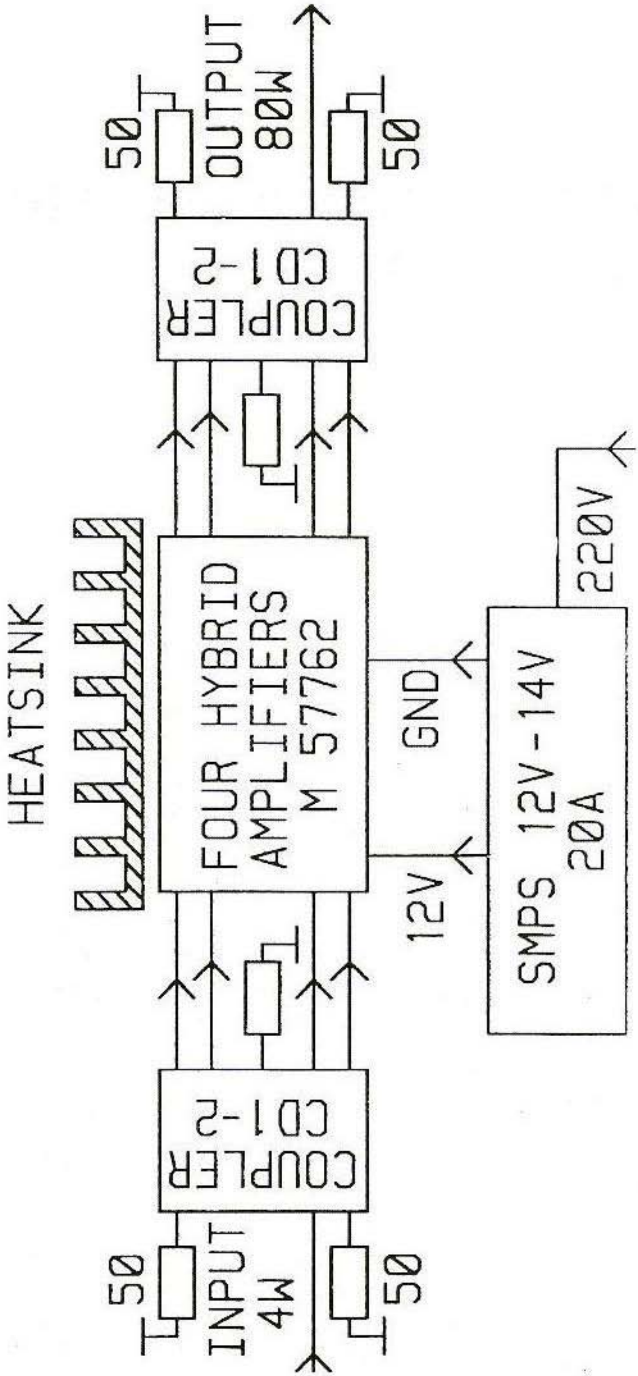


Fig.1: Circuit Connections

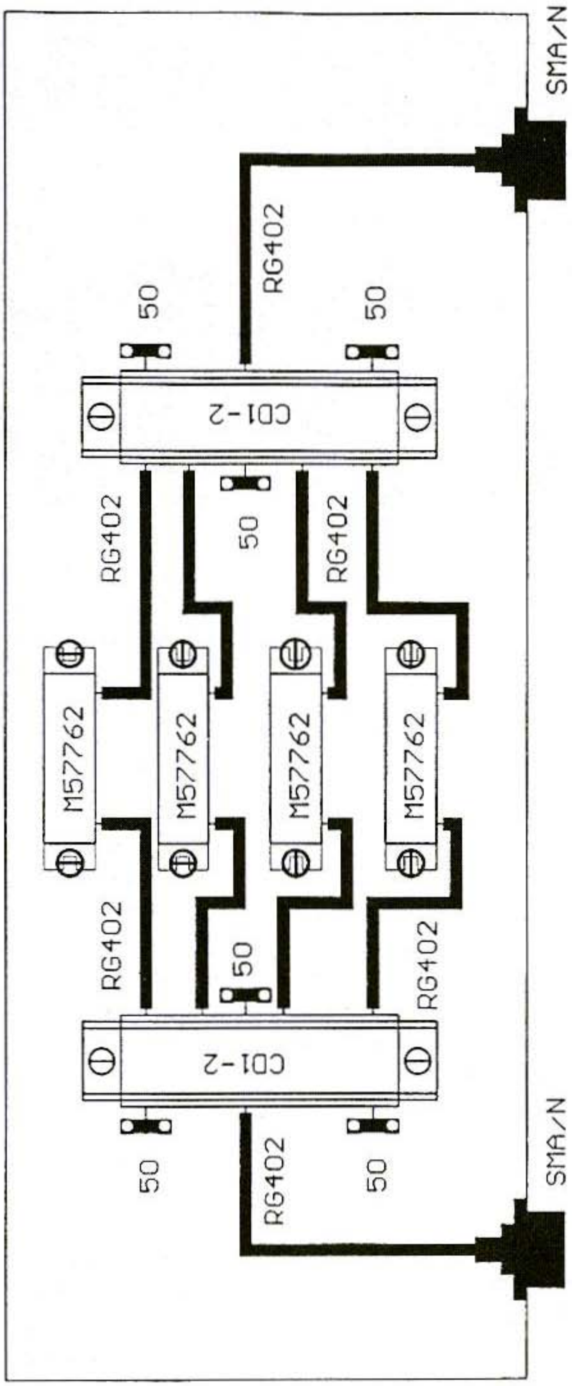
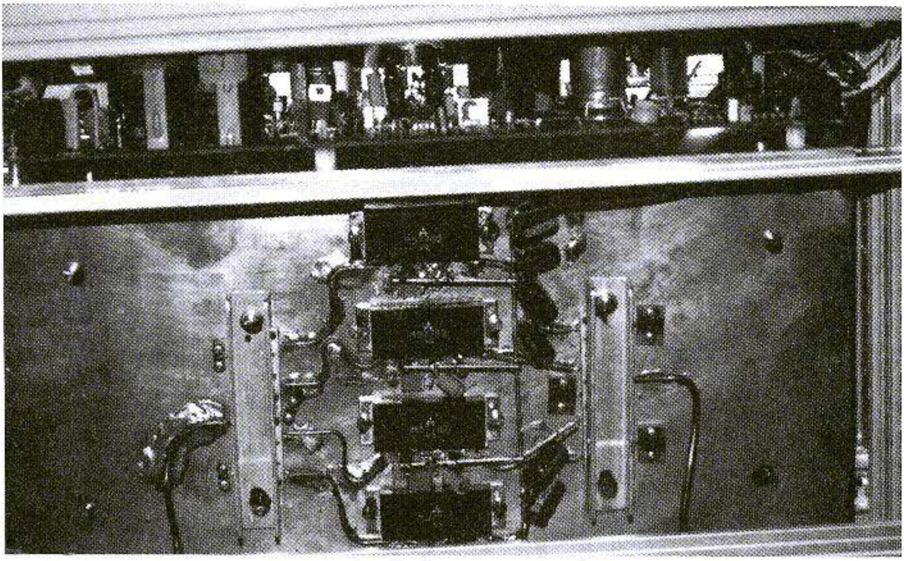


Fig.2: The RF Paths



The circuit connections are shown in figure 1 and the RF paths can be seen in more detail in figure 2 and in photographs 1 and 3.

The input RF power reaches the four-way power splitter through low-loss (semi-rigid) coaxial cable. The power splitter is broadband and needs only 50Ω low-reactance resistors, which can be seen in photograph 2, as external components. The latter are required to absorb any unbalance at the splitters outputs. With 4 Watts at the amplifiers input, allowing for a slight insertion loss in the splitter and the coax stubs (0.3dB global) about 1 Watt comes out of each of the four splitter outputs, and is brought through four identical lengths of semi-rigid coax to the four hybrid amplifiers inputs.

Power supply decoupling is very important and must be carried out using good quality capacitors and short leads at each hybrid amplifiers power input.

No printed circuit board is required for the amplifiers to avoid losses. RF input and output leads are soldered directly to the coax inner conductors. The same technique is used for the power supply leads. This can be seen in figure 3. Photograph 3 shows a close-up view.

The outputs of the four amplifiers are connected with semi-rigid coax to a second hybrid power splitter, which is used the other way around to merge the outputs from the four hybrid amplifiers. Three balancing resistors are also used here, somewhat larger than at the input.

Semi-rigid type RG 402 coax cable is used throughout the RF paths. It has low losses, can be readily soldered and its power handling capability is quite high at these frequencies.

The whole circuit is mounted on a large copper plate, which acts as a ground and is bolted onto an aluminium heatsink. About 200 Watts will have to be dissipated and a fan can be useful here.

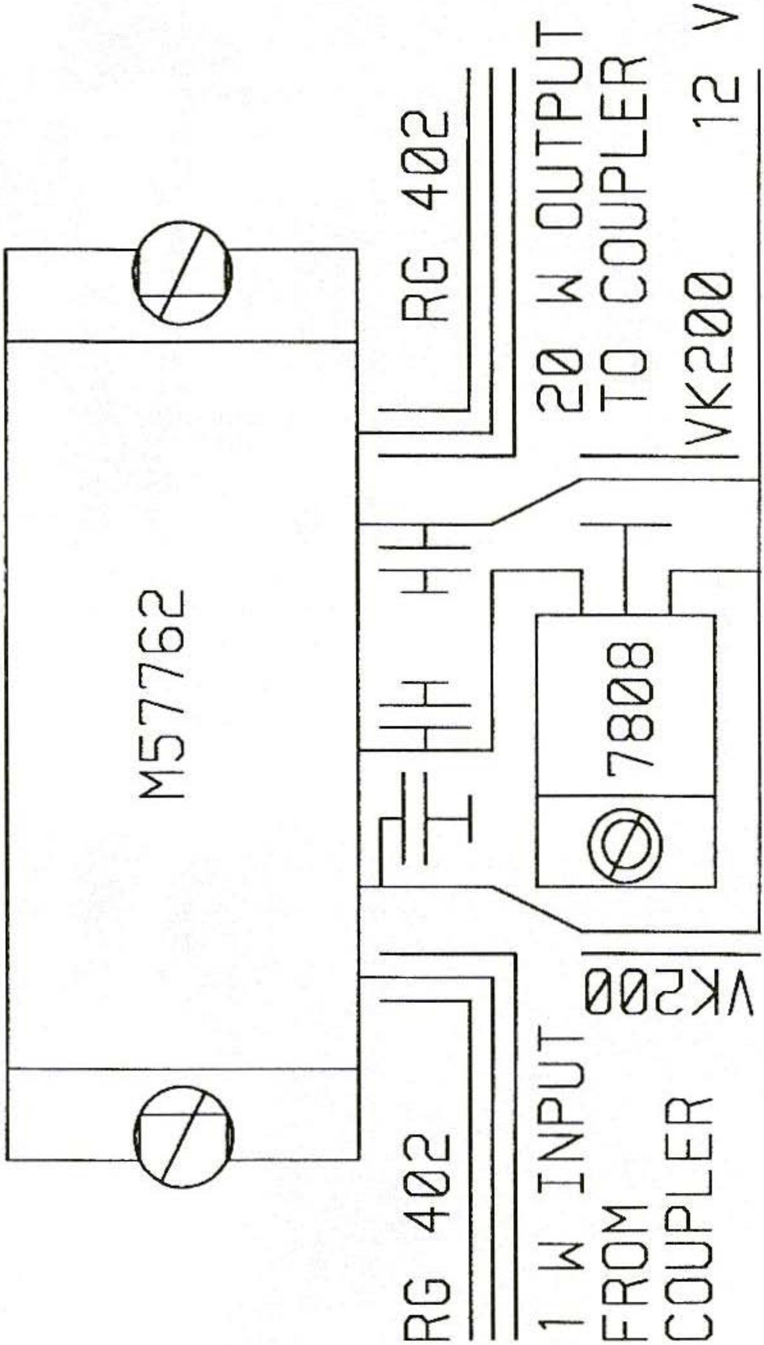
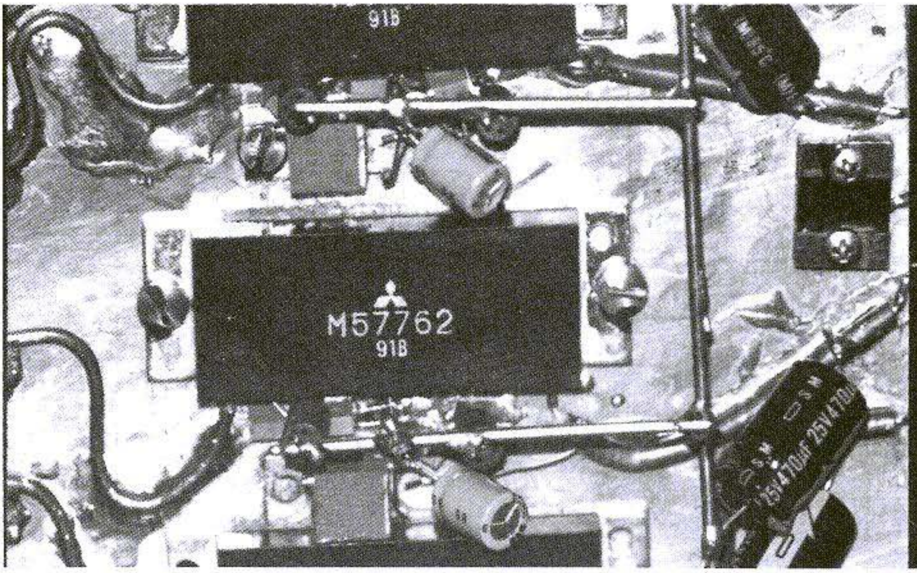


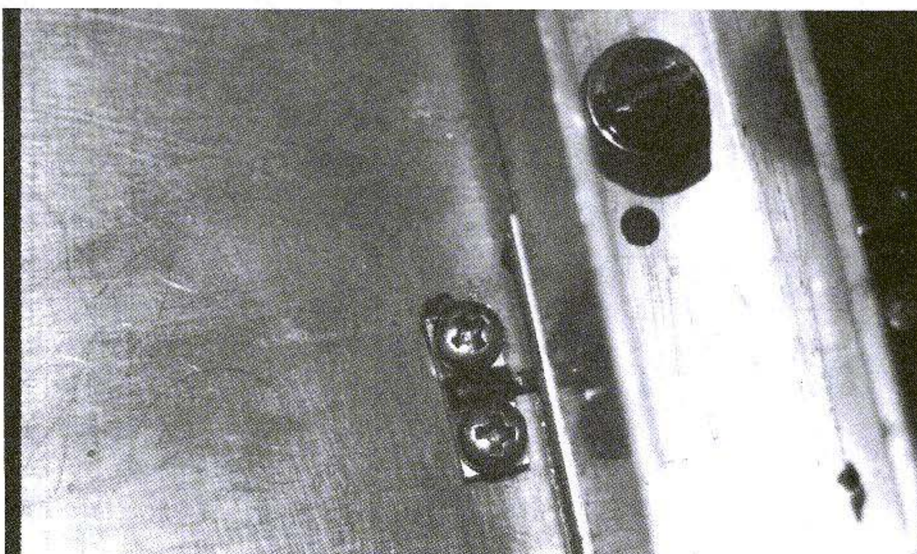
Fig.3: Method of connection to the "Brick" Amplifiers



The coax stubs are soldered directly to the copper plate, which makes a very stable assembly. This is essential in this kind of circuit.

The 50Ω resistors and the hybrid amplifiers are bolted to the heatsink. The copper plate and the heatsink are drilled together. The holes in the copper plate are then slightly enlarged and only the heatsink is tapped.

The 6dB hybrid couplers are pressed against the copper plate by a U-shaped metal profile, visible in photographs 1 and 2.



The power splitters packages must make good contact with the copper plate on their whole length. This requires small holes to be drilled in the copper plate to make space for the rivet heads in the splitters packages.

Thermal compound should be used under the hybrid amplifiers and between the copper plate and the heatsink. It should be stressed that the final quality of this circuit depends mainly upon the mechanical stability of the assembly and the quality of the RF connections.

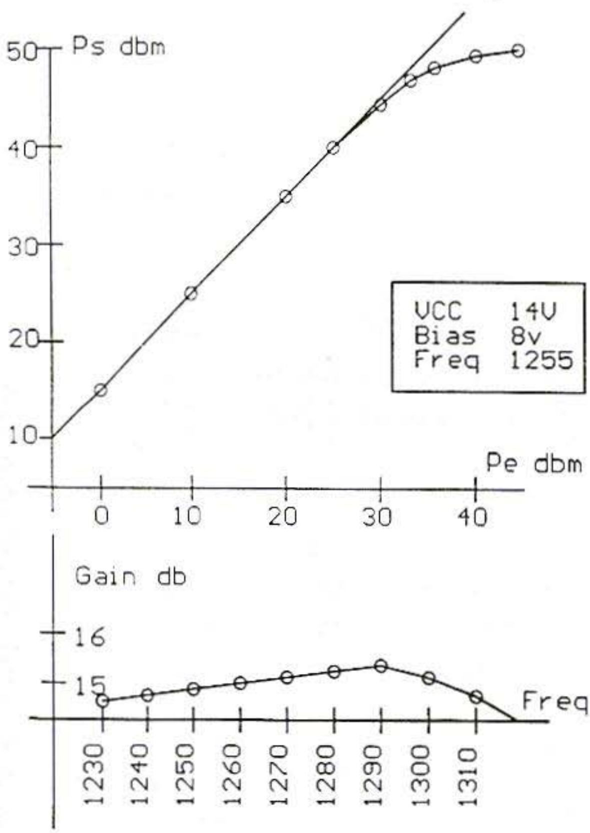


Fig.4: Test Results

2. TESTING

This is the easiest part, because there is nothing to be tweaked ! Connect a 50Ω 100 Watt 1.2 GHz specified dummy load at the output through a power meter and another dummy load at the input. Turn the power supply on and check that the 12 volt and 8 volt rails on the amplifiers. Check that NO RF is present at the output.

Turn the power off and replace the dummy load at the input with an RF generator at the correct frequency. Increase the power progressively at the input and check the output power as well as the balancing resistors temperature, which should not rise at all unless there is an unbalance somewhere ! The gain must be about 15dB.

The test equipment used was:

HP 435A Power Meter
HP 8481A Power Sensor
Narda 769 30dB Attenuator
HP 8558B Spectrum Analyser
HP 8444A Tracking Generator
HP 5386 Frequency counter

On our prototype the impedance match was better than 25dB from 1240 to 1310 MHz. Saturated output RF output power was 80 Watts with a 12 volt power supply, 90 Watts with 13 volts supply and 100 Watts with 14 volts.

Power supply current is of the order of 18 Amps at 13 volts and output power is saturated and 20 Amps at 14 volts.

Figure 4 shows power output versus power input.

Our thanks go to Marc F3XY for confirming our results.

3. PARTS LIST

1m	RG 402 Semi-Rigid Coaxial Cable	RS Components
6	Resistive Loads	ELHYTE SARL type T-250-500-10
2	6dB Couplers	Nucletudes SCD type CD 1-2
4	Hybrid Amplifiers	Mitsubishi type M 57762; RF Parts Co.

4. SUPPLIERS ADDRESSES

Elhyte Sarl, B.P. 34, 91620 La-Ville-Du-Bois, France. Tel: 16 (1) 69 01 68 51

Nucletudes SCD, Av. du Hoggar, Z.A. du Courtabeuf, B.P. 117, 91944

Les-Ulis 2, France. Tel: 16 (1) 69 07 10 20

RF Parts Co., P.B. 700, San Marco, California, USA

Detlef Burchard, Dipl.-Ing., Box 14426, Nairobi, Kenya

Don't be Afraid of High-Frequency Transformers

Transformers for high-frequency applications offer a multitude of interesting applications, and can also be influenced by the developer with regard to their characteristics; on the other hand, it is precisely this many-sided nature of theirs which still frightens circuit developers away.

1. INTRODUCTION

In high-frequency applications, transformers are used as broad-band four-terminal networks for impedance changes, phase inversion, isolating earths, balancing circuits and much more. They consist of a core with magnetic conductivity and coils. Basically, they work just like a mains transformer, which also becomes more and more defective as its size decreases. High-frequency transformers actually have many more inherent deficiencies, which are triggered by slight and complex core permeability, skin and proximity effects, coil capacity, and leakage and feed inductances. Core materials and shapes are scarcely standardised. So it's small wonder that the circuit developer initially shrinks back from high-frequency transformers.

But there is another way of looking at it - as a challenge! Apart from the ordinary high-frequency coil, a high-frequency transformer is one of the few components, the characteristics of which circuit developers can still influence. They can only obtain the others - diodes, transistors, integrated circuits, resistances and capacitors - from catalogues, without being able to

There are a few manufacturers who have specialised in high-frequency transformers, but they too still work with the generally accessible physical bases - they have no special tricks. So initially we shall concern ourselves with the basics, then go through a relatively simple method of measurement, and finally put the knowledge we have gained to use in dimensioning high-frequency transformers.

2. CORE MATERIALS AND SHAPES

Cores made of ferrite, iron powder and iron oxide powder are used for the frequency range above 3 MHz (high-frequency, VHF, UHF). As the frequency rises, the losses in the core from magnetic conversion effects and eddy currents rise too. Other ferrite mixtures or finer powder allow the frequency range which the transformer can use to be extended upwards - to 0.5...1 GHz at present, using materials, the permeability of which is still only 3...8. It is obvious to think of "air" cores too, which are also on sale from some manufacturers. They are compressed together from a low-loss insulating material, and thus essentially form a carrier for the coil. However, on the market these cores cost just as much as ferrite cores. So you can just as easily make them yourself on a lathe, to any dimensions you like, out of Plexiglas, Trolitul or Teflon semi-finished products.

2.1. Core shapes

Examples of core shapes involving an enclosed iron path are shell cores, EE or EI shapes, multi-bore cores and ring cores. The first-named can still be used in the high-frequency range. Because of their relatively high permeability, they still guide the flux through the angular flow line path well. But as frequencies rise only a few materials remain available.

Ring cores and dual-bore cores come into use for non-angular flow line paths. So we shall concern ourselves predominantly with these, although the procedure for dimensioning at low frequencies and with other core shapes is exactly the same, and indeed often brings success earlier, because the inherent deficiencies are much less significant.

2.2. Core sizes

A high-frequency transformer is a compact component. So essentially its dimensions should not exceed $\lambda/100$. This requirement permits the selection of a core with suitable dimensions. At 100

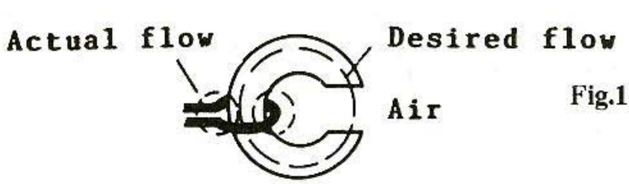


Fig.1: Cores with low μ_r are ineffective if they are unfavourably wound

MHz a ring core may have a diameter of up to 30mm, at 3 GHz only 1mm.

We expect all the current flowing into a coil to emerge again at the other end. But this can't always be guaranteed, due to divided internal capacitances. If the coiled wire is very long, it acts as a circuit, and our expectations can not be fulfilled at all. Experience teaches us that the maximum coiled wire length is about $\lambda/20$. Thus a 100 MHz transformer can have a wire length of 150mm, while for a 3 GHz transformer the figure is only 5mm! This effect can even be erected into a principle and corresponding circuit transformers can be developed. But that is outside the scope of this article.

Once the size has been determined, we have to look for a suitable core material. Manufacturers usually specify the optimal frequency range for broad-band transformers, which can be used for orientation purposes. In the same way, curves showing the complex permeability, $\underrightarrow{\mu_r}$, which is made up of a real element and an imaginary one, are often printed in data sheets.

$$\mu_r = \mu_r^* - j\mu_r^{**} \qquad (1)$$

Thus the core material becomes more unfavourable as μ_r^{**} gets closer to μ_r^*. It can also be seen from the same measuring curves that μ_r^* diminishes considerably at high frequencies - right down to 1. In this case, the material is no better than air, but creates greater losses. Anyone wishing to transfer more power must also check the magnetic induction, B.

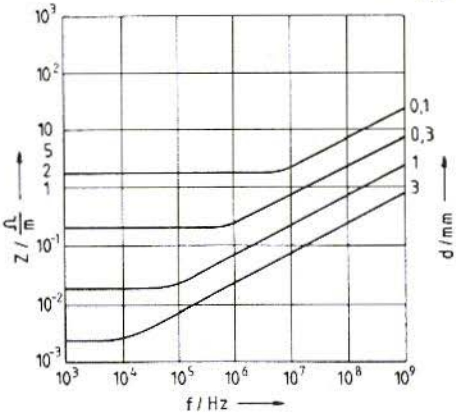


Fig.2: Skin Effect for Round Copper Conductors

$$B = \frac{U}{\omega \cdot w \cdot AF} \qquad (2)$$

where:

U = peak voltage
ω = circuit frequency
w = coil number
AF = magnetic cross-section

For ferrite cores, B should not exceed 100 to 300 mT, whereas iron cores can even carry 1 T. In small signal mode we remain considerably below this, so as to keep non-linearities small. A practical limit is 1 to 10 mT. For example, in a core with a cross-section of $4mm^2$ which is wound in 10 turns of wire, it is reached at a frequency of 10 MHz if the high-frequency voltage applied amounts to 1.8 to 18 V_{eff}. Anyone wishing to know the precise value should measure the non-linearity values under operating conditions as a k-factor or inter-modulation factor.

A ring core with a permeability of 10 can even be envisaged on the lines of

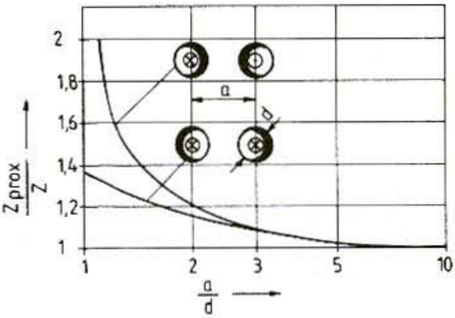


Fig.3: Proximity Effect for Round Conductors

Fig.1, made from 90% magnetically ideal conductive material and a 10% air gap. The flow lines of an individual turn of wire in the unfavourable position (in relation to the air gap) will find a shorter path if they run directly around the wire. The core is thus largely ineffective. This is not the case with double-bore cores, especially if the wire completely fills the bore.

The selection of a suitable core is thus dependent on the number of turns. A double-bore core is used for a few turns, while a ring core is preferable for many. It is even possible now to estimate the wire diameter. For double-bore cores the bore should be filled as completely as possible, and for ring cores one turn should lie on top of another in the bore. We shall see later

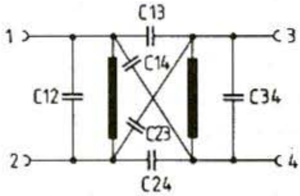


Fig.4: Partial Capacitances of a Transformer with two turns

that this kind of winding can also have disadvantages. So the advantages and disadvantages must be balanced against each other in each individual case. In any event, uniform distribution of the coil around the perimeter is very important for ring cores, much more important than reducing the capacitance between the beginning and the end of a turn by leaving a sector spare, as is often recommended. All turns must be threaded through double-bore cores twice, whilst naturally they only need be threaded through ring cores once. The physical bases also have a beneficial effect on the amount of work involved in the winding!

3. SKIN AND PROXIMITY EFFECTS

As is generally known, at high frequency the current flows only in a thin layer on the surface of a conductor. From a specific cut-off frequency onwards, the impedance continuously increases in accordance with the law of $\sqrt{f}$ (see Fig.2). Wires with a diameter of 0.1mm have a cut-off frequency amounting to 10 MHz, and for those with a diameter of 1mm the value is 100 kHz. Nevertheless, no matter how high the frequency, the thicker wire has a lower impedance than the thinner. The law of $\sqrt{f}$ can be understood only as stating that the ohmic resistance and the inductance increase simultaneously and by the same amount. So, in the frequency range which interests us here, not only does the coil resistance increase with the frequency, but the inductance does as well. The rise can be so considerable that the DC resistance of a coil essentially plays no further role.

The less well-known proximity effect increases the resistance and inductance further if other conductors are carrying the same current in the vicinity. This is naturally always the case with transformers. Now, it still comes down to whether other conductors are carrying current flowing in the same direction or the opposite direction. The former is the case if the turns are in the same coil, and the second circumstance arises if turns from the secondary coil are in the vicinity.

Fig.3 presupposes only a single additional conductor in the vicinity. Here the current is pushed out into the areas of the conductor surface shown in black. The increase in the impedance, as against the conductor, which is anyway already plagued by the skin effect, is again very considerable. Matters become even more complicated if several live wires are running in the vicinity, which means advance calculation is scarcely possible.

From the skin and proximity effects we learn that it can be advantageous to keep a certain distance between conductors. This can be done through double enamelling or through additional silk covering. The distance to the windings of another coil is more critical than that to windings of the same coil. Later this will bring us into conflict with an opposite requirement - to reduce the leakage inductance as much as possible.

4. INTERNAL CAPACITANCE

All windings have a partial capacitance against each turn in their own coil, and against all other turns. This monstrously complicated situation can be easily surveyed only if you make very considerable simplifications. Fig.4, showing the six possible artificial capacitance values between the accessible ends of the primary and secondary windings, constitutes such an incomplete representation of reality. The two windings are often twisted together before coiling, in order to keep the leakage inductance low. C_{13} and C_{24} can then be assumed to be large in relation to the other partial capacitances. This case can also, to some extent, be theoretically comprehended and calculated as a double wire circuit. The equation applying is:

$$C/pF = \frac{0{,}28 \cdot \varepsilon_r \cdot l/cm}{\ln\left(\frac{a}{d} + \sqrt{\frac{a^2}{d^2} - 1}\right)} \quad (3)$$

where:

ε_r = dielectric constant
l = length
a = distance between centres
d = wire diameter

Accordingly, $C \rightarrow a$ if $a \rightarrow d$. Fortunately, the practical situation is more user-friendly. Depending on the amount of twisting, wires which are simply twisted yield values of between 0.8 and 2pF/cm., irrespective of the wire diameter. C_{13} and C_{24} are each allocated half of the value thus arrived at. With the assistance of estimated values for lengths and distances, the values for the other capacitors can be estimated, at least in terms of their order of magnitude.

5. INDUCTANCES

The main inductance of a transformer is calculated in the well-known fashion from equation (4):

$$L = AL \cdot w^2 \qquad (4)$$

where:

AL = inductance factor
w = number of turns

The AL value is either known (from the core manufacturer) or is determined, using a sample coil, at a frequency far below that of the main resonance. Should a ring core be used, with a known permeability, then the inductance can be calculated, even from the dimensions:

$$L_h/nH = 2 \cdot w^2 \cdot \mu_r \cdot h/cm \cdot \ln\frac{D_2}{D_1} \qquad (5)$$

where:

h = ring core height
D_2 = external diameter
D_1 = internal diameter

Equation (5) reveals two important facts. First, the inductance does not depend on the diameter, but only on the ratio of the external diameter to the internal diameter. This allows for miniaturisation without loss of inductance, provided the ring core height is retained. Secondly, the inductance is proportional to a dimension if similar ring cores - i.e. D_2, D_1 in the same ratio - are in use. If a transformer has been optimised for a specific frequency, an optimal transformer can immediately be obtained for another frequency by increasing or decreasing all the dimensions in relation to the frequencies and keeping the same number of turns.

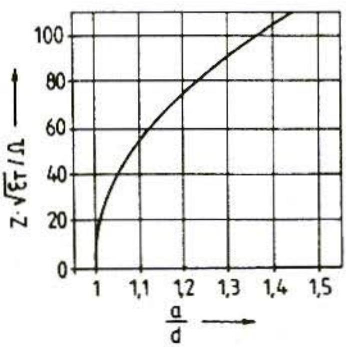


Fig.5: Impedance level of a Double Circuit

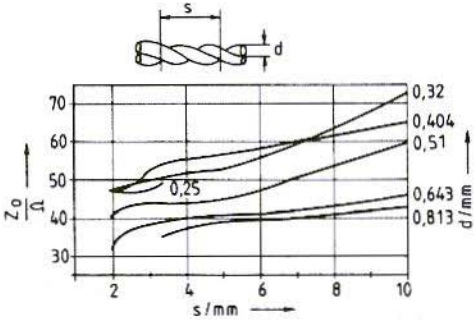


Fig.6: Impedance level of a Double Circuit made from twisted CuL wires (as per ARRL)

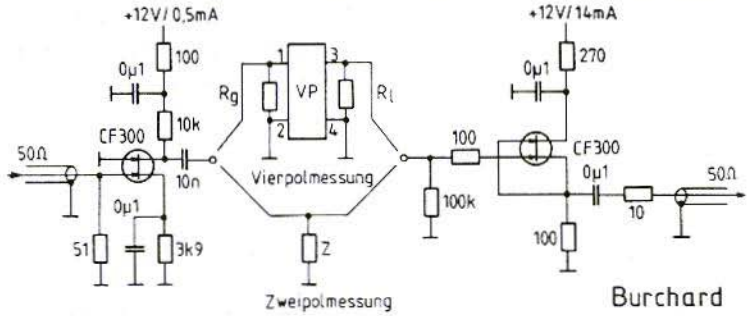


Fig.7: Circuit for 2-pole and 4-pole measurements on Transformers
Note: Vierpolmessung = 2-pole measurement; Zweipolmessung = 4-pole

Leakage inductance arises because fractions of the primary flow avoid the secondary windings. For windings twisted together, it is the fraction which flows between two wires through the coating and not through the core. The core material therefore plays no part, and the leakage inductance can be calculated for the double wire circuit using formula (6):

$$Ls/nH = 4 \cdot l/cm \left(\ln \frac{2a}{d} - \frac{a}{l}\right) \qquad (6)$$

For a ® d and l » a, it changes into Ls = 3.7 nH/cm., which applies for all wire diameters. So if a transformer has a twisted coil with a length of 4cm, then its leakage inductance will be approx. 15 nH.

The impedance level, Z_0, of the double wire circuit is often also of interest if, for example, the feed to the transformer is of a significant length and can be executed to match the impedance level.

$$Z_0/\Omega = \frac{120}{\sqrt{\varepsilon_r}} \ln\left(\frac{a}{d} + \sqrt{\frac{a^2}{d^2} - 1}\right) \qquad (7)$$

This function is also plotted in Fig.5. Z_0 changes very abruptly for low a/d ratios. Small interval changes are expressed by large changes in Z_0. Here too, the real technical situation is much more user-friendly. After ARRL measurements (1) on twisted enamelled wires, a desired impedance level can be repeatably set, with the length of twist and the wire diameter within specific limits. For Fig.6, the values given in AWG and inches in (1) were converted into the metric system. No general regularity can be discerned in the curves alone.

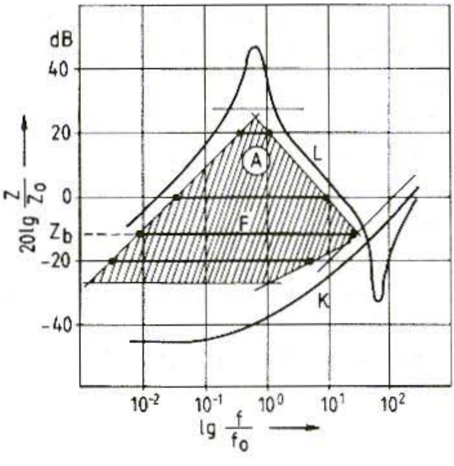
Those working with thinner wires must carry out Z_0 measurements themselves. A further application for a transformer development matching an impedance level can be seen later in Fig.11c.

The feed inductances, which are themselves not zero if the coil start and end are twisted together, are of particular importance. This case can be calculated in accordance with equation (6). Separated feed wires can be comprehended as a circuit with a diameter, D. We then obtain:

$$L/nH = 2 \cdot \pi \cdot D/cm \cdot \ln\frac{D}{d} \quad (8)$$

In order to get some idea of the size of a feed inductance, we set D = 0.5cm and D/d = 20, and we obtain exactly 10 nH. The feed has a similar inductance to the leakage inductance of a transformer with a wire length of 3cm. The feed wire is thus generally not negligible, and in practise must be kept as short as possible.

Some curves deviate from the formula laid down (3). They serve more to explain the principles than for calculation purposes here. Owing to the fact that the complex permeability, the skin effect and the proximity effect depend on the frequency, the case arises here that the derivation of Murphy's Law - "all constants are variable" - not only applies but is physically demonstrable. Thus it is still necessary to proceed empirically to obtain the practical dimensions of high-frequency transformers. But the principles can provide very valuable hints on the direction in which changes are to be made if the result is to satisfy your desires.



6. EMPIRICAL DIMENSIONING

Having selected the core size, shape and material, calculated or estimated the coil and determined the inductances and capacitances, we now want to find out as quickly as possible how well our transformer works. We can't do this entirely without measuring equipment. A circuit will be needed with which two- and four-pole measurements can be carried out over a wide frequency range. Such a circuit was taken from (2) and is reproduced in Fig.7. At the front is a signal generator, which can be manually adjusted, and the tracking generator of a spectrum analyser. At the rear are a sensitive high-frequency voltmeter and the receiver section of a spectrum analyser. A network analyser can naturally be used as well.

No long mathematical calculations are required to substantiate the theory that a good transformer running idle at the

Fig.8: Dimensioning of a Transformer by neans of Measured Curves for Idle-Running (L) and Short Circuit (K) Impedance values. Useful region (A), Frequency range (F) when Operating Impedance value (Z_b) is selected

output will have a very high impedance on the input side, and vice versa. If there is a short-circuit at the output, its input impedance must be small. Measuring the blocked impedance and short-circuit impedance levels is the fastest way of determining the frequency range and the faults of the transformer, and simultaneously obtaining enough information to carry out any improvements which may prove necessary. The picture obtained could be something like Fig.8. Here all the "variable constants", together with the only approximately calculated partial capacitances and inductances, have the same effects as they will in operation later.

The L-curve (idle running) initially rises from low to high frequencies, conditioned by the main inductance. There then follows a resonance with the internal capacitance, which is sharp if the quality of the material is sufficiently high, but which can also be of little significance, or can be absent altogether if the core material displays big losses. Subsequently, because of the internal capacitance, a decrease in the impedance occurs, which frequently also indicates a resonance hole, should the internal capacitance and the leakage inductance form a series resonance of a sufficiently high level. At very low frequencies, the short-circuit curve, K, is horizontal, and corresponds to the DC resistance there. This is followed by a section with a gradient of 22.5° if a skin effect arises, or with a gradient of 45° if the leakage inductance predominates. A resonance also often forms at high frequencies in which capacitances of which we have not yet spoken play a role, e.g. the switching capacitance of the measurement circuit in Fig.7 (about 3pF) or a capacitance lying parallel to the leakage inductance in the transformer. Below the L-curve and above the K-curve lies region A - it keeps its distance from both curves. It is reasonable to use the transformer here. The following considerations apply.

A reasonable limit for the additional damping is 1dB. It arises, for example, if the DC resistance of the transformer is 1/9 of the operating impedance, Z_b. An interval of 19dB is associated with this, which is also to be maintained for

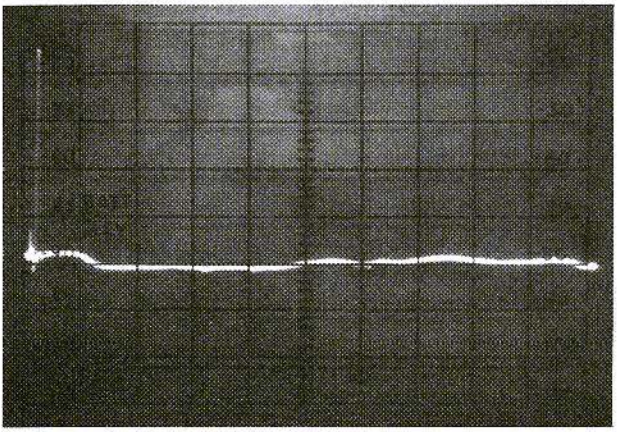


Fig.9: Spectrum of a 51Ω Reference Impedance in a Measurement Circuit as per Fig.7
Y: 10dB/div (right-hand scale)
X: 50 MHz/div, 0...500 MHz
(Zero mark visible on left)

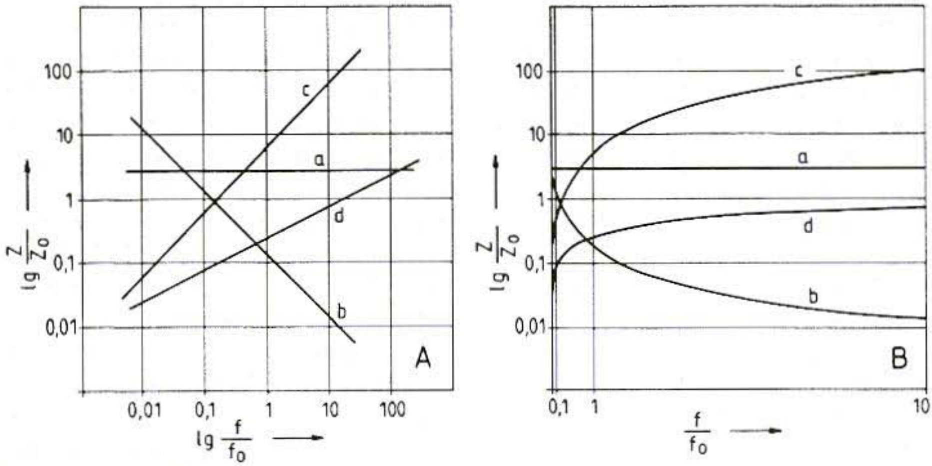


Fig.10: Image Distortion due to Linear Frequency Axis (B) as against usual representation (A) of "High Frequency Wallpaper"

all other horizontal curving sections of the L-curve and the K-curve. Should the curves run below ± 45°, there are thus pure reactive impedances, and a similar consideration applies - that an interval of dB is to be maintained. Finally, for sections of curves below 22.5°, an interval of 12dB must be maintained. It is clear from all this that the transformer in Fig.8 can be operated with various operating impedance levels. In each case, it will have a different frequency range, F, for a rise of 1dB in the additional damping at the frequency range limits. The range F, to which the operating impedance, Z_b, of about 12dB below Z_o belongs, is clearly as broadband as possible. All other ranges are just as useful if they meet the developer's wishes.

Should the transformer have a winding ratio deviating from 1:1, reverse it and measure on the other side, with the first turn either idling or short-circuited. The L-curve and the K-curve will be exactly the same, but displaced up or down. The displacement gives a precise measurement of the resistance-winding ratio.

Empirical dimensioning can go further should the transformer still not meet the developer's wishes. By appropriate use of the relationships known from the principles, the L-curve is displaced upwards as far as possible, into the middle of the desired frequency range. This may mean it is necessary to change the number of turns or to select another core material or another core size. At the same time, the K curve

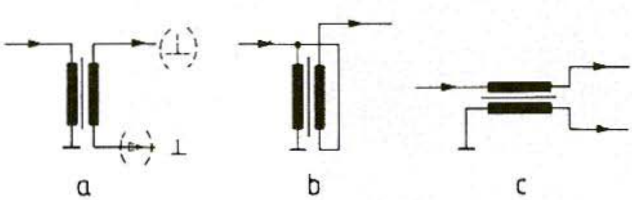


Fig.11: Three possible Connection Diagrams for Spooled Specimen Cores

should be as low as possible. In many cases, no sufficiently large region A is available. A decision then has to be taken as to whether more additional damping is permitted (distances to the L-curve and K-curve shorter), or whether the frequency range, F, can be reduced.

Should the L-curve and K-curve be satisfactory, carry out a four-pole measurement as per Fig.7, using the same measuring rig, with the transformer being balanced on the input and output sides with the desired impedance levels, Rg and Rl.

This measurement should then demonstrate that the transformer has "succeeded".

7. SOME EXAMPLES

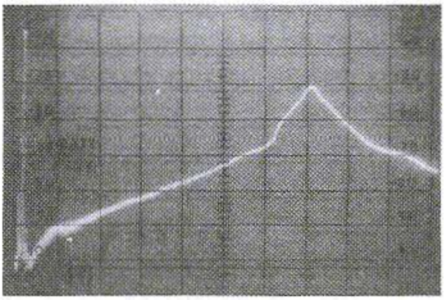
Ring cores made from various materials are available with external diameters of 6mm. Their size makes them suitable for frequencies of up to 500 MHz. But the difficulties increase with the frequency. So the only examples given here are those which deal with frequencies at the upper end of the VHF range.

The ring cores were provided with a twisted coil made from two 0.1 CuL wires with 5 turns. The coiled wire length is 4cm, which should be sufficiently short for up to 375 MHz. Moreover, an additional little double bore core with roughly the same core cross-section was provided with identical winding and also measured.

The measurement circuit from Fig.7 was wired up to a spectrum analyser with a built-in tracking generator. Z-measurement of a 51Ω ohmic resistance gives the reference line of Fig.9. It should actually be completely flat. The oscillations in the 0 to 500 MHz frequency range arise through inadequacies in the measurement circuit, the level control of the tracking generator, the frequency response of the analyser and ripple on the connection cables. No measures were taken to eliminate the oscillations. Most radio amateurs have to live with similar inadequacies in their equipment. Nevertheless, we can find out everything worth knowing about the transformer to be measured.

Fig.8 was plotted on a log-log scale. In this representation, which coincides with the usual "high-frequency wallpaper", the identification of reactive components is particularly simple. Depending on the gradient in Fig.10a, it can be seen at a glance whether you are dealing with an ohmic resistance (a), a capacitance (b), an inductance (c) or a skin effect (d). The analyser used here, like most of them, has a linear frequency axis. The 4 different impedance curves then appear as shown in Fig.10b. This is extremely confusing and takes a lot of getting used to.

The coiled cores should have been used in the measurement circuit in the same way as they are wired up later for operation. With the winding described, three applications are conceivable, the circuits for which are sketched in Fig.11: a 1:1 transformer, with or without phase reversal (a), a 1:2 upwards transformer (b) and a balun (c).


L

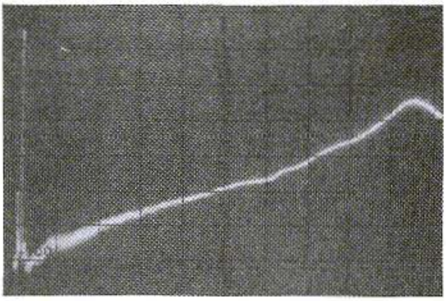
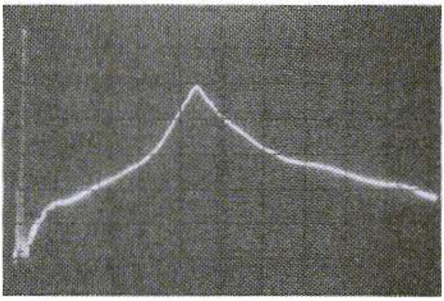

K

Fig.12: Idle-Running (L) and Short Circuit (K) Impedance of Transformer with Teflon Ring Cores (both axes are shown here, and in all subsequent spectrums, on the same scale as for Fig.9)


L

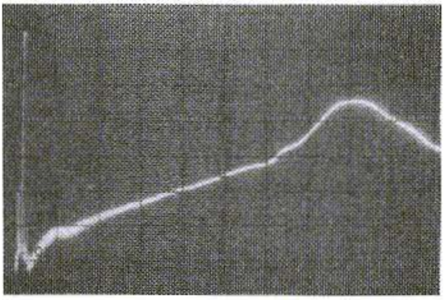
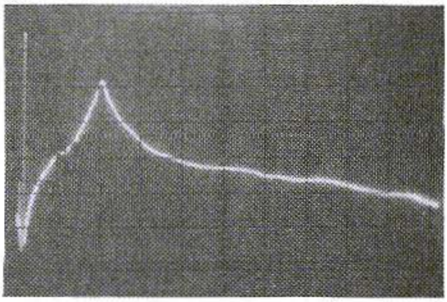

K

Fig.13: Iron Oxide Ring Core made from Vogt Fe803


L

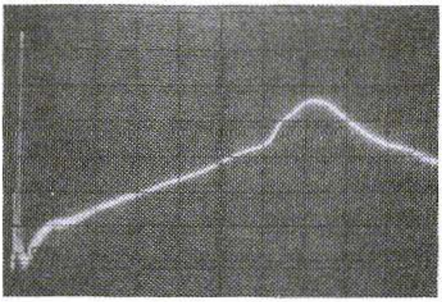

K

Fig.14: Ferrite Ring Core made from Vogt Fi130

The latter is not really a transformer, but rather a choke for the suppression of the asymmetrical current at the output. So we have not pursued this application any further. The measurement curves displayed below relate to application a from Fig.11, with phase reversal. The end of the first strand of the twisted coil is linked to the beginning of the second strand, and this point is earthed. The connection ends of the inputs and outputs of the transformer were deliberately selected to be not very short (7 to 8mm), so the feed inductance is considerable. Apart from the commercially available ring cores with D2 = 6mm, D1 = 2mm and h = 2mm, a Teflon ring was also wound with about the same dimensions and was also measured. Fig's.12 to 16 show the L-curves and K-curves for all the cores. The scale corresponds to that of Fig.9. Readers with image storage will prefer to store the curves for the reference impedance, together with the two other curves, and display them simultaneously on the screen. The rest of us will now have to compare the curves, in order to find the region A in which a feasible transformer can be made into a reality using the core in question and this winding.

From this we can learn that: The transformer with the Teflon ring core can really be used at only one frequency, app. 350 MHz, with an operating impedance of about 250Ω. But it will be explained later how it can be made into a perfectly good transformer for a higher frequency range.

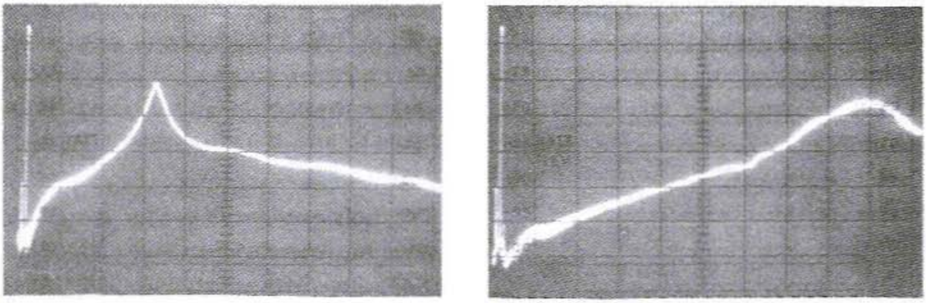
The transformer with the iron oxide ring core is suitable for frequencies between 175 and 260 MHz, with Zb = 120 to 170Ω. With ferrite ring cores, the frequency range is 20 to 160 MHz. At low frequencies Zb = 15Ω should be optimal, and at high frequencies Zb = 110Ω.

The iron powder ring core should be good for frequencies of 60 to 200 MHz, with the best operating impedance rising from 30 to 100Ω.

Finally, the ferrite double bore core comes out pretty well like the ferrite ring core, with a frequency range of 20 to 175 MHz for a Z_b value of 10 to 110Ω.

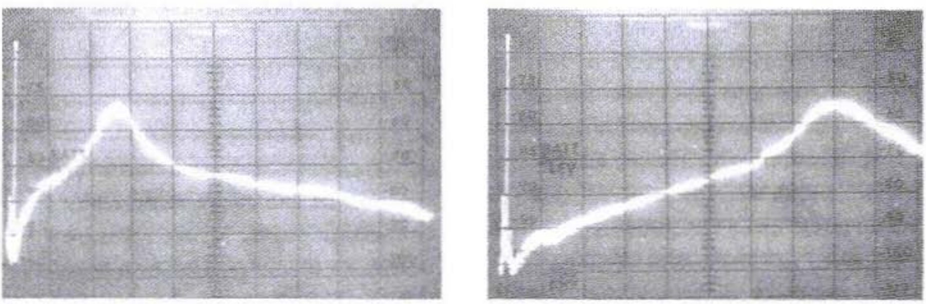
The reasons for the relatively unfavourable results are: in the case of the Teflon ring, the low permeability of the core; for all the ring cores, the loose winding, which is far from the optimal winding density; and for all transformers the feed inductance. 3 to 6dB can be gained by using really short connection wires, laying them near the frame, or twisting. The winding of the double bore core can not be improved as this coil completely fills the bore. Thicker wires could be used for the ring cores. Smaller cores can also bring the K-curve down.

Finally, all transformers, within a limited frequency range, can be considerably improved if the leakage inductance is compensated for. This can often be done by the suitable selection of a coupling capacitor which is already available. The K-curve then has a resonance hole, and can drop to the sum of the DC resistance and the skin effect resistance.



L K

Fig.15: Iron Powder Ring Core made from Amidon 10-Mix



L K

Fig.16: Double Bore Core made from Siemens U17

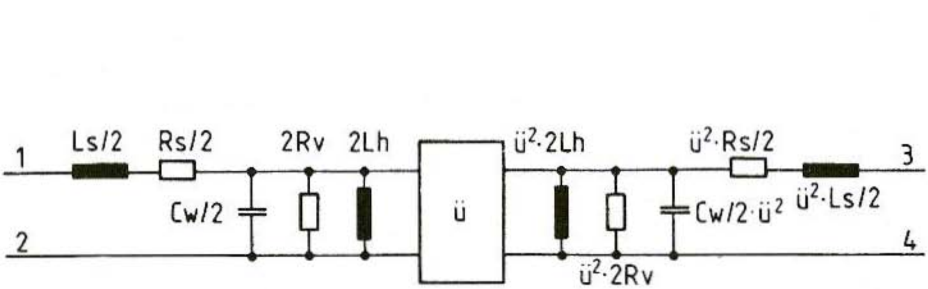


Fig.17: Equivalent Circuit Diagram for explaining Compensation

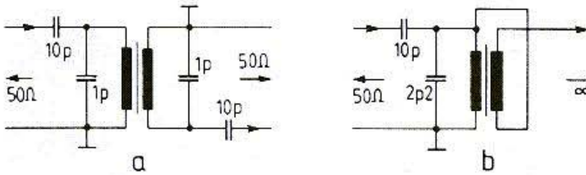


Fig.18: Series and Parallel Compensation
a. 1:1 Phase Transformer
b. 1:2 Upwards Transformer

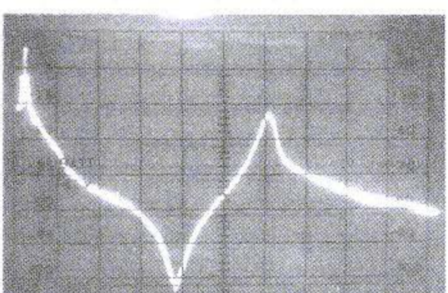
8. COMPENSATION

Fig.17 shows an equivalent circuit diagram which can be used to clarify compensation measures. The main inductance, L_h, the internal capacity, C_w, the loss resistance, R_v, the series resistance, R_s, and the leakage inductance, L_s, are uniformly distributed on both sides around an ideal transformer, Ü.

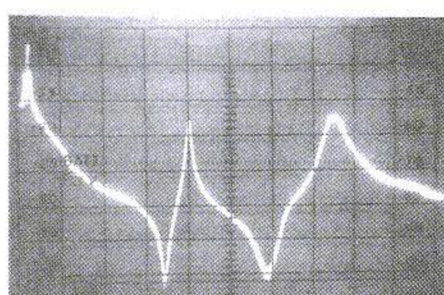
On the assumption that the series impedance values are low, as against the shunt impedances, we mainly measure L_h, C_w and R_v at the terminal pair 1 - 2 and at the other terminal pair 3 - 4 in idle running; for a short circuit, by contrast, we mainly measure L_s and R_s. The series compensation consists of capacitors in series, with an input and an output which are precisely in resonance with L_s at the desired frequency. It is possible, even without parallel compensation, that the resonance peak of the L curve will be displaced to another desired frequency. Both capacities and inductances which can do this are conceivable.

The parallel compensation measures will be used only in cases when, due to variation in the number of turns, the main resonance can not be displaced to the desired frequency.

Both types of compensation can be quickly and conveniently dimensioned with the aid of the measurement circuit. Let's try it on the transformer with the Teflon ring cores, which is really unsuitable.

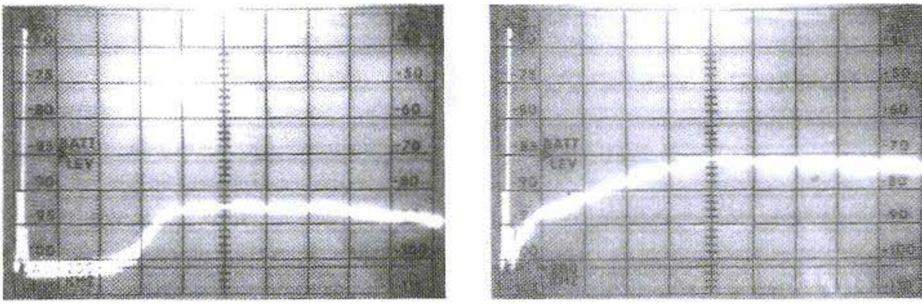


L



K

Fig.19: Idle-Running (L) and Short Circuit Impedance (K) of Compensated 1:1 Transformer with Teflon Ring Core



a b

Fig.20: Propagation Factor of Transformer with Teflon Ring Core
a. as per circuit, Fig.18a
b. as per circuit, Fig.18b

If a 10pF capacitor is wired up in series to both terminal pairs, this generates a resonance hole in the K curve at exactly 300 MHz. Parallel switching of 1pF to each terminal pair displaces the resonance in the L curve from 360 to 300 MHz. The circuit obtained can be seen in Fig.18a. The L curve and K curve are now as shown in Fig.19. There is a big enough region A for a frequency range of 260 to 320 MHz with operating impedance values of 50 to 120Ω. Four-pole measurement (Fig.20a) at 50Ω, with both sides sealed off, certainly gives us a flat line over a wider frequency range, but the resolution is not adequate for locating the 1dB band width. Point-by-point measurement using a signal generator and a high-frequency voltmeter make a more accurate analysis possible. The transformer created can be used between 180 and 310 MHz (Fig.21a).

The same winding can also be used, as per Fig.18b, as a 1:2 upwards transformer, in order, for example, to carry out noise matching on an FET. For this application, the secondary runs idle. There is thus no point in carrying out a series compensation there. So all compensation measures are carried out on the primary, as Fig.18b shows.

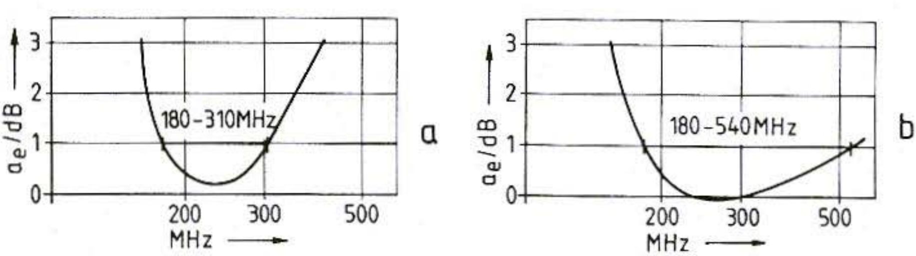


Fig.21: Very Precise measurement of Additional Damping of Transformer with Teflon Ring Core
a. as per circuit, Fig.18a
b. as per circuit, Fig.18b

Fig.20b shows four-pole measurement over the wide frequency range. More precise measurement, as per Fig.21b, reveals that the equipment is usable between 180 and 540 MHz. The reduction in the additional damping above 310 MHz arises through an additional resonance of the secondary leakage inductance with the measurement circuit capacitance. Should the secondary be sealed off at 200Ω, we obtain a frequency response which is very similar to Fig.21a.

9. SUMMARY

Measurement of the idle-running and short-circuit impedance values over a wide frequency range is proposed, in order to develop high-frequency transformers. The measurement circuit required is not expensive. Examples are given showing what curves can be expected, how they should be interpreted, and what measures are required to match their course to the problem definition in each individual case. Defective transformers can be considerably improved by compensation, but the frequency range is reduced. Very small cores and very short connection lines allow equipment to be used at much higher frequencies than could be demonstrated here, owing to a lack of appropriate measuring equipment. At any rate, very low-loss transformers can still be manufactured without sub-miniature technology up to 500 MHz. The use of cores with $\mu_r = 1$ makes it possible to manufacture transformers for which there is no longer any suitable core material, even for such frequency ranges.

10. LITERATURE

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Josef Fehrenbach, DJ 7 FJ

Dual-Band Exciter for 10 GHz and 24 GHz

A reflector, two GHz bands and only one dual-band exciter - this arrangement offers many advantages, whether in terms of cutting down on weight or of the possibility of positioning the aerial at 10 GHz and then working at 24 GHz with the aerial already aligned.

1. INTRODUCTION

Anyone who is already QRV on 10 GHz and would now like to extend his or her activities to the next highest amateur band at 24 GHz has to struggle with a few unusual features in the process. Narrow-band technology is already established, with adequate frequency stability. However, the receiver sensitivity and the transmission power available (if the price is kept reasonable) are still markedly below those now available at the 10 GHz level. Whilst the noise factors of the input stages tumble year by year with the help of HEMT's, transmission power levels of more than 100mW unfortunately continue to remain unattainable for most OM's. Moreover, the 24 GHz technology surplus market provides very little which is of use here.

Apart from the already higher path attenuation, active radio operation is also made more difficult by additional attenuation, e.g. due to water vapour (moisture, mist, rain, snow, hail).

If your receiver noise figure has already been optimised and the transmission power has been matched to your money bag, the only way left open to improve the technical potential is through the aerial. Only by this means is it still possible to make up for a few of the losses generated, for example, by too low a transmission power and by the high path attenuation. For any further gain to be obtained, the aerial has to be as large as possible. The easiest way to obtain high gain levels is using satellite dishes. The gain from satellite dishes is calculated as:

$$G[dB] = 10 \cdot \log\{(\frac{\pi \cdot d}{\lambda})^2 \cdot \eta\}$$

where:

d = diameter of dish
λ = wavelength
η = aperture efficiency using amateur reflectors - usually 0.5 to 0.6

Unfortunately, the apex angle of an aerial is inversely proportional to its gain. The apex angles of satellite dishes can be easily estimated using the equation below:

$$\varphi \cong 70\lambda/d$$

where:

φ = horizontal and vertical apex angle
λ = wavelength
d = diameter of dish

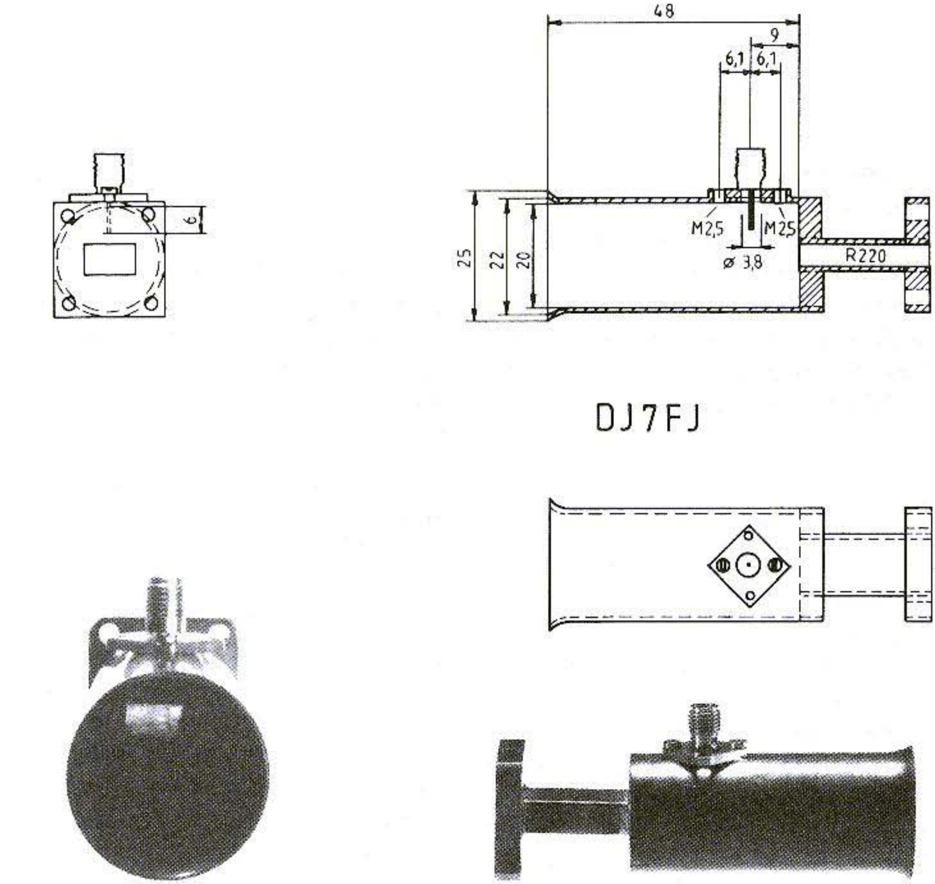


Fig.1: Dimensions of the 10/24 GHz Dual-Band Exciter
Eqpt: Copper Pipe, R220 Waveguide, 2 Flanges and an SMA Socket

Tables relating to the above relationships between aerial size, gain, apex angle and wavelength can be found in the literature, e.g. in (1).

At 24 GHz, satellite dishes between 30cm and 1m obtain from 35 to 45dB with apex angles of 3° to 0.8°. Unfortunately, these apex angles must be respected, not only in the horizontal plane, but in the vertical plane as well.

Consequently, with these preconditions, it is obviously very desirable to operate with an aerial which is precisely aligned even before the QSO begins. With a dual-band aerial which is first optimised in the low-frequency 10 GHz band (with significant signal reserves), this can be achieved considerably more easily than by working directly at 24 GHz. Moreover, such a dual-band aerial is extremely practical for portable mode itself (BBT, among other things).

This article describes a dual-band exciter for 10 and 24 GHz for use in satellite dishes, with a "focus/diameter ratio", f/d, of app. 0.35 to 0.4.

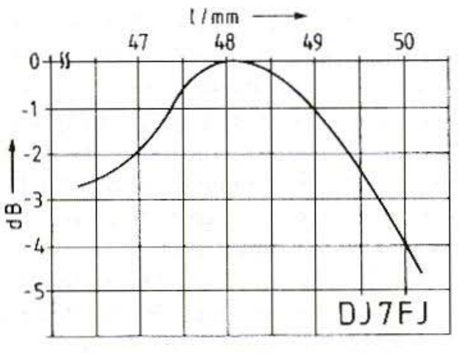


Fig.2: Radiator Length plotted against Gain at 24 GHz

2. OPERATING PRINCIPLE

A pipe radiator for the 13cm band is described in (2). The relationships between f/d and the measurement of the associated feed-horn are also comprehensively explained there. On the basis of a reflector with a "focus/diameter ratio" (f/d) of, for example, 0.38, we obtain a focus angle of about 130°. For a rectangular horn exciter, this gives us an $a \cdot \lambda$ value of 0.79 and a $b \cdot \lambda$ value of 0.66 for 10 GHz. If we start with a round horn, from the average of the two values, we arrive at a horn diameter of app. $0.72 \cdot \lambda$. At 10 GHz (λ = 3cm), that corresponds to a horn diameter of 21.6mm

Copper pipes with external diameters of 20mm and internal diameters of 18mm, obtained on the construction market, come very close to this dimension. Provided you don't bend them, these pipes make splendid wave guides for 10 GHz. If you mount such a round wave guide into the focus of a reflector, without an additional horn, this is

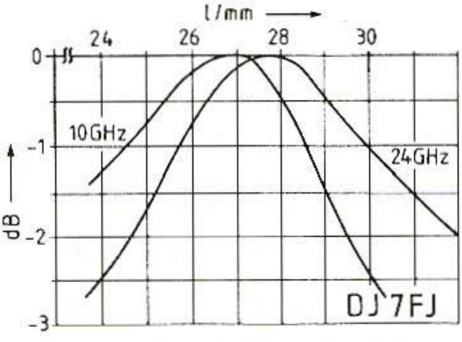


Fig.3: Optimising Gain by changing Radiator Position

already a very good radiator for reflectors with an f/d of app. 0.35. Such a radiator forms the basis of the combination radiator described here.

The 10 GHz section consists of a round wave guide with an internal diameter of 18mm The 10 GHz signal is coupled in through a coaxial wave guide transition by means of a pin.

The length of the wave guide unit is not critical for 10 GHz. The open wave guide end can be used directly as a radiator. You are recommended to press on a small horn (expand the tube end) at f/d ratios approaching 0.4. Experiments have shown that this can increase the aerial gain by up to 1.3dB, as against a wave guide end which has simply been turned. A grooved horn available as basic equipment for 11 GHz satellite television which matched the wave guide diameter gave the same readings.

Fig.1 shows the radiator set-up for 10 GHz and the expansion for 24 GHz. A 24 GHz E220 rectangular wave guide was mounted into the closing plate of the 10 GHz radiator. The lower limiting frequency of the R220 led us to expect total locking for 10 GHz, which measurements confirmed. Measurements of the adaptation at the 10 GHz coupling jack showed that it made no difference whether the pipe was closed off with a continuous plate or with an R220 wave guide inside such a plate.

Measurements of the gain from the 10 GHz section also showed no discernible changes. What we still had to determine was what influences or losses would be generated in the 24 GHz signal by the 10 GHz pin. It also had to

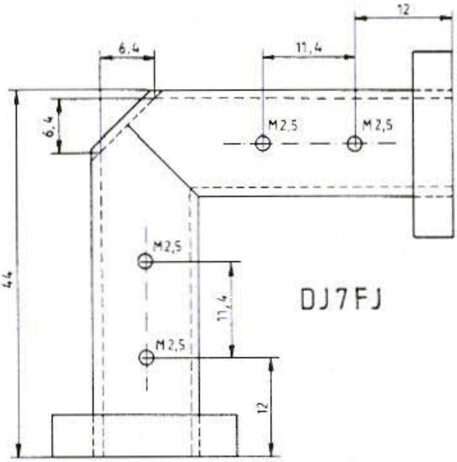


Fig.4: Construction Dimensions for a "Compensated Bend" made from Waveguide Material

be determined how the gain at 24 GHz behaves in relation to the length of the round wave guide.

2.1. Experiments with Prototype

A radio path with a length of 300m and a test reflector with a diameter of 70cm and an f/d of 0.38 were used for the experiments described below.

Fig.2 plots the relationship of the radiator length to the gain obtainable at 24 GHz. The radiator position, i.e. its distance from the reflector, was optimised for maximum gain in each case. The optimum readings were clearly obtained with a radiator length of 48mm - and you don't have to be accurate to a tenth of a millimetre. As expected, the variation in the length tested had no influence at 10 GHz.

Further experiments were carried out with dimensions in accordance with

Fig.1, i.e. with a radiator length of 48mm The relative gain at 10 GHz and 24 GHz was determined in relation to the optimal radiator position in each case. The maximum readings were about 1cm apart, but the drop in each case is insignificant (see Fig.3). When we copied this model, we had to determine the optimum radiator position using a beacon. Anyone who can't receive beacons for the two bands should optimise using the band available. As the experiments showed, the deviation from the optimum radiator position is not evident.

At 10 GHz and 24 GHz, the matching values gave a return loss of better than 15dB. The coupling from 10 GHz to 24 GHz could not be measured. The coupling from 24 GHz to the 10 GHz coaxial output amounted to app. 10dB. Anyone who wants to generate output using this system at 24 GHz should either build a stop filter into the 10 GHz port or switch the 10 GHz pre-amplifier off when transmitting at 24 GHz.

3. SET-UP

The copper radiator pipe is cut to the appropriate length. Before the holes and threads for the SMA 2-bore flanged bush are bored and cut (4-bore units can also be used), a flat surface approximately 4.5mm wide should be prepared at the appropriate location by milling or filing. The coupling pin consists of the connection pin of the SMA bush and a 1mm silver wire, soldered on and nipped off at the correct length. So that the R220 wave guide can be incorporated more easily, it is recommended that you manufacture an adapter unit about 20-30mm long. To do this, solder two wave guide flanges onto a suitable wave guide unit. If one bush is clamped into a four-jaw chuck and held, the other one can be carefully turned to an external diameter of 18mm This adapter is then pushed into the copper pipe, brought into the correct position and soldered in. The soldering should be done using as little solder as possible, preferably on a hotplate or using hot air. The wave guide flanges can be made to bulge on the side facing away from the connection in question before the first soldering by light blows from a centre punch, so that they sit tightly on the wave guides. Otherwise, you run the risk that the adapter will fall apart again during the final soldering process.

Anyone who doesn't have turning equipment available can, for example, mount a cornered closing plate instead of the bush adapted to the pipe, with a suitable opening for the R220 wave guide.

3.1. Installation in Reflector

The radiator can be fastened using a 3 or 4 strut, and the 10 GHz signal can be fed in by means of a high-quality cable. For 24 GHz, in any case, you are recommended to use a kind of "walking stick" made of wave guide material. Those wishing to produce the corners required for this themselves will find some tips in Fig.4 on the construction of a compensated bend using R220 for

24.192 GHz. The only critical dimensions are the chamfers on to the corner cover, which have an internal length of 6.4mm The M2.5 threaded bores need to be manufactured only once as a pair for the whole walking stick. If necessary, they can take adjustment screws.

4. FUNCTIONAL TEST AND GAIN COMPARISON

Completed systems in 60cm and 90cm reflectors showed that no losses of any kind arose due to the building-on of the 24 GHz section.

At 24 GHz, in comparison with standard horn equipment and monoband dishes, reflectors with dual-band exciters obtained a gain which was always 2.5dB below what monoband excitement would have produced. This can essentially be traced back to the fact that the exciter's 24 GHz lobe was slightly too narrow.

4.1. Practical experiences

Several 24 GHz enthusiasts have been working on copying this exciter for some time. QSO operation at 24 GHz is considerably simplified by the ability to preset the direction precisely during the preceding 10 GHz QSO. It has been shown time and again that the aerial setting really needed no further optimisation following the switch from 10 GHz to 24 GHz. This more than compensates for the loss of the final 2.5 dB by comparison with the uncertainty of direction which would otherwise prevail and which often leads to failed QSO's.

Readings for a 90cm satellite dish with a dual-band exciter:

	10 GHz	24 GHz
Gain	37 dB	41.5 dB
Apex angle	2.3°	1.4°

The author wishes all success to those wishing to copy him, and is already looking forward to many 24 GHz QSO's.

5. LITERATURE

(1) RSGB VHF-UHF Manual

(2) Tubular Radiator for Parabolic Antennas on the 13cm Band
DJ 1 SL, VHF Communications no. 4, 1976

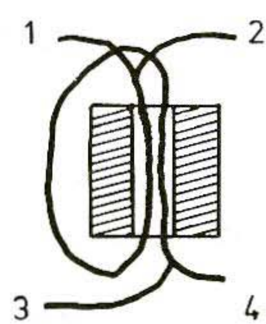
Dr. Ing. Jochen Jirmann, DB 1 NV

Addenda and Comments on the Article: Tracking Generator (Issue 1/92)

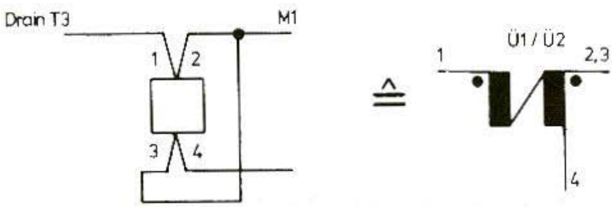
Some comments regarding the tracking generator for the spectrum analyser are set out below:

1. In Fig. 2 (p. 37), the bottom left input is wrongly described: it should read "From 1. LO of the analyser, f = 450 - 950 MHz, P app. -10dBm".
2. The earth connection of the collector for T1 (BF324) and the 100kΩ resistance (right-hand connection) on connection 2 of I3 have been omitted from the equipping drawing (Fig. 4, p. 41).
3. The basic 10Ω/47 pF combinations for T4 for the circuit diagram (Fig. 3, p. 133) and the components diagram (Fig. 4, p. 38) have been transposed; however, this has no influence on the equipment's operation.

Twist individual wires together



Thread through Ferrite Beads



Wire the two-part windings together, observing the correct polarity

4. If the hole pattern of coils L1 to L5 and filter Fi3 does not match the board holes, the coil insert should be carefully pressed out of the filter bowl (using a 3mm drill shank, through the balancing bore), turned through 90° and re-inserted. Remember that the coil unit is made up of two shells, the coil and the winding body, which come apart easily.

5. If the diode you use as the ALC rectifier is not a high-barrier type, then the ALC rectifier generates a considerably higher voltage than in the specimen unit, and the output power level can not be set high enough. Reducing the 33kΩ resistance at the positive end of P1, for example to 10kΩ, is of assistance.

6. It seems there is still some lack of clarity concerning the structure of the broad-band transformers and the significance of the switching polarities: here are the instructions again in an abbreviated form:

   - Twist together two wires of suitable length:
   - Thread the twisted wires through the ferrite beads two or four times:
   - Wire up the two part-windings in series with the right polarity in such a way that a transformer winding is created with centre tapping. In each case, the tap is at the coupling capacitors for the M1 ring mixer.

Reiner Erping (DB9JC) and Wolfgang Schneider (DJ8ES)

13cm FM-ATV Exciter

The following article describes an FM-ATV exciter for the 13cm band. Here the authors are going back to a circuit design which has long been tried and trusted. Following the circuit developed earlier for 23cm, a similar assembly is described here today for the next amateur radio band up (2,320 - 2,450 MHz).

Supplemented by a power amplifier and the necessary baseband preparation for the audio signal at 5.5 MHz, the exciter turns into a complete FM-ATV transmitter.

1. CIRCUIT DESCRIPTION

Fig.1 shows the complete circuit for the FM-ATV exciter. The description below concentrates on the essential functioning elements, such as the voltage-controlled oscillator (VCO), the phase lock loop (PLL) and the broad-band amplifier.

1.1. The Voltage Controlled Oscillator (VCO)

The VCO (Fig.2) was designed using reasonably-priced standard modules. The capacity diodes are normal UHF types of component. The same applies to the oscillator transistor.

The two tuning diodes, D1 and D2 (BB505G), are directly connected up to the base of the VCO transistor, T1 (BFR90a). The incorporation of these structural elements here is determined by the frequency. All the connections are to be as short as possible!

The 10kΩ/100Ω combination of resistances between the collector and the base of the transistor, T1, determines its static current. In addition, the counter-coupling works for stabilisation through the 10Ω resistance in the emitter circuit. As regards high frequencies, the collector is earthed through 220pF (disc capacitor without connecting wires!).

The oscillator signal is tapped off at the emitter for the subsequent amplification

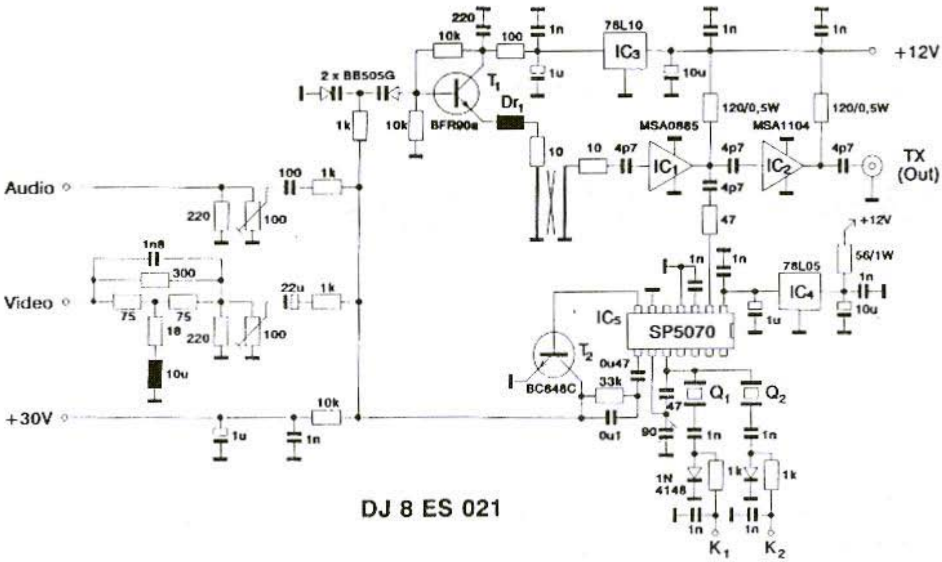


Fig.1: Circuit Diagram of the 13cm FM-ATV Exciter

stages. The connecting wires of the emitter resistance form a coupler. This loose connection avoids, as far as possible, any feedback effects on the VCO from the stages downstream. This oscillator circuit can be adjusted up to max. 2.8 GHz. In the frequency ranges of interest to us, around 2.35 GHz, a frequency variation range of 600 MHz is aimed at using a tuning voltage of 0 to max. 30V. The output power in this case falls from app. 40mW at 1.9 GHz (U_{Tuning} = 0V) to 30mW at 2.5 GHz (U_{Tuning} = 30V). Fig.3 shows the relationships in a graph.

1.2. The Phase Lock Loop (PLL)

Together with a VCO, the SP5070 (IC5) from Plessey forms a complete PLL synthesiser (Fig.4). This system can be adjusted from 300 to 2,500 MHz

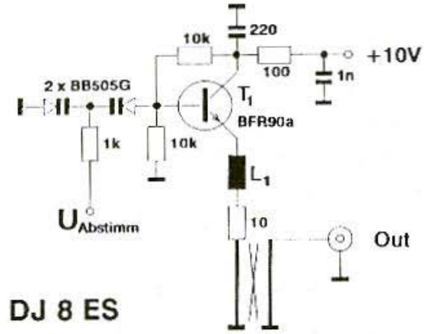


Fig.2: VCO Circuit (abstimm = tuning)

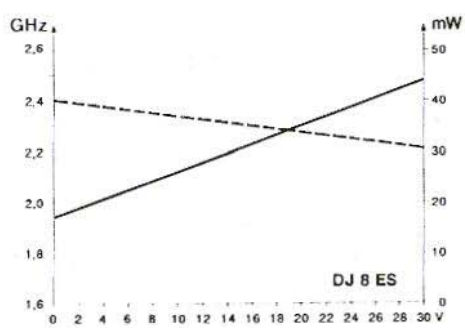


Fig.3: VCO Frequency and Output Power as a function of U_{Tuning}

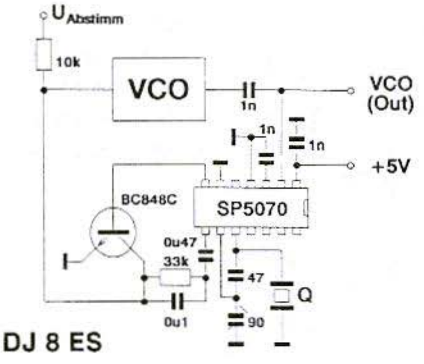


Fig.4: PLL using Plessey IC (abstimm = tuning)

The module comprises a pre-divider with a pre-amplifier and a frequency divider. The latter has a fixed divider factor - the ratio between the synthesiser frequency and the reference frequency is 256:1.

The phase comparator is provided with its reference frequency, derived from an external oscillator or a crystal. The output of the comparator controls the capacity diodes in the VCO through an external transistor (BC848C).

The reference frequency, conditioned by the fixed frequency divider, has to be smaller than the synthesiser frequency by a factor of 256. This means that, for a desired output frequency of, for example, 2,335 MHz, a reference frequency of 2,335 MHz / 256 = 9.1210938 is required.

1.3. The Broadband Amplifier

The amplifier section after the VCO is provided by two MMIC's from Avantek. In all, the circuit yields a maximum of 50mW at app. 20dB amplification. The first stage of the broad-band amplifier is equipped with an MSA0885 (IC1). With its 13dB amplification in the 13cm band, this MMIC greatly increases the only loosely coupled VCO signal. The function of this stage can simply be checked by measuring for a voltage drop of 4.4V across the 120Ω resistor.

The desired output level is obtained only with the second stage, an MSA1104 (IC2). 7V can be measured at the similar point as for the previous stage. The static current of the broadband amplifier at any time is set by means of these pre-resistances. For correct dimensioning, independent of the operating voltage, frequency and required output power, the necessary parameters should be taken from the

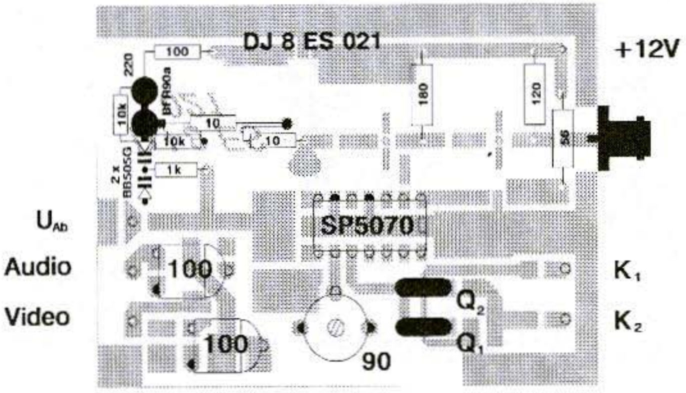


Fig.5: Component Layout on the Ground Plane side

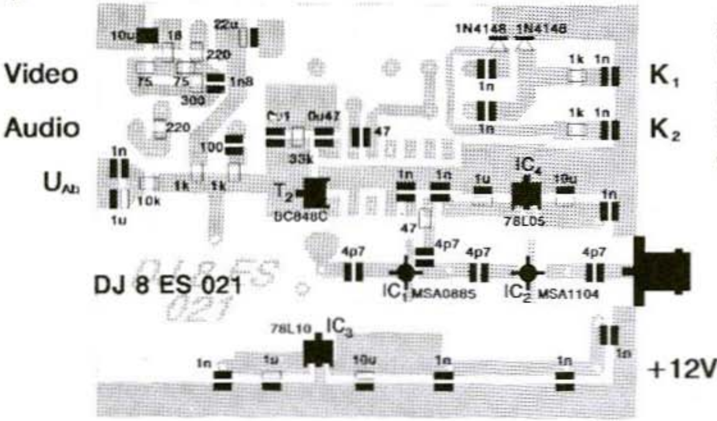


Fig.6: Component Layout of the SMD Components on the print side

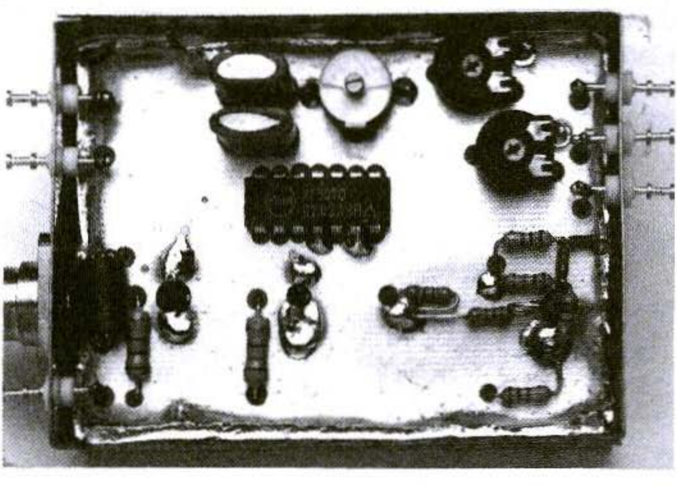
data sheet. And here: $RV = (U_B - U_P)/I_P$ and $PV = (U_B - U_D) \times I_D$ apply. From these calculations, in accordance with their rating, both resistances have a value of 120Ω/0.5 W.

2. ASSEMBLY INSTRUCTIONS

The FM-ATV transmitter for the 13cm band is assembled on double-sided coated epoxy board measuring 72mm x 53mm. The board thus fits into a standard tinplate housing which measures 74mm x 55.5mm x 30mm.

After being cut to size, the board is first cold silver-plated and finally bored. The broad-band amplifiers lie in the board, and suitable holes must be bored for these. A recess measuring app. 1mm x 8mm should be sawn out on the corresponding board edge.

After the boring, the holes for resistances, trimmers, the crystal, etc., can be drilled on the earth side of the board (fully-coated side) using a 2.5mm drill. The copper surface around the drill hole must be removed.



Once this preparatory work has been taken care of, the board can be sprayed with soldering lacquer. After drying, the earth feedthroughs for the broad-band amplifiers are provided.

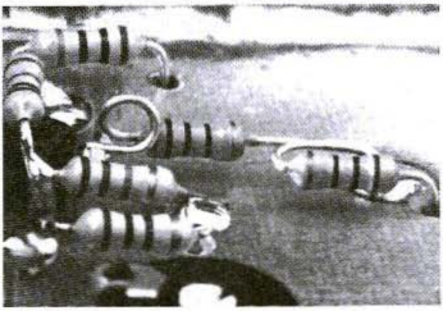


Fig.7: VCO in "Breadboard" format

The actual insertion does not take place until the board has been soldered into the tinplate frame. The BNC flanged bush must lie with its flange on the cover edge. If the board is now inserted in such a way that the bush pin touches the track (cut off projecting Teflon collars with a knife first), the top cover must still go on satisfactorily after the provisional insertion of a crystal.

When the board has been soldered to the lateral surface of the housing, the components can be inserted. SMD components can just as well be used for this as conventional structural elements with connecting wires. Fig's.5 and 6 show the layout of the individual modules.

The VCO is produced in breadboard format (Fig.7). It is important here that all connections are made as short as possible! Only in this way can interference from undesirable inductances be avoided.

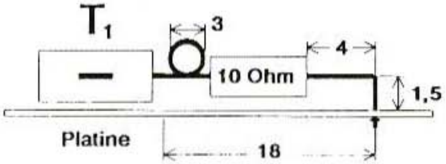


Fig.8: Dimensions of the Coupler (platine = board)

The coupler is provided by the connecting wires of the two 10Ω resistances. The same applies for the choke, Dr1, in the emitter circuit.

The diagram below (Fig.8) shows the most important dimensions for this coupler, and the enlarged photo of the VCO in Fig.7 shows further details.

2.1. Component list

IC1	MSA0885 (Avantek)
IC2	MSA1104 (Avantek)
IC3	Voltage reg TA78L10F (SMD)
IC4	Voltage reg TA78L05F (SMD)
IC5	SP5070 (Plessey)
T1	BFR90a (Siemens)
T2	BC848C (Siemens)
D1, D2	BB505G
Q1, Q2	Crystal, see text
(Dr1	1 winding with 3mm dia. in lead of 10Ω resistor)

2 x	1N4148 (SMD)
2 x	10Ω, 0.1W, 9mm basic grid
1 x	100Ω, 0.1W, 9mm basic grid
1 x	1kΩ, 0.1W, 9mm basic grid
2 x	10kΩ, 0.1W, 9mm basic grid
2 x	120Ω, ½W, 12.5mm basic grid
1 x	56Ω, 1W, 12.5mm basic grid
2 x	Pot, 100Ω, 5/10mm basic grid
1 x	90pF trimm, 10mm basic grid
1 x	220pF, Disc capacitor

6 x	Teflon bush
1 x	BNC socket (UG-290 A/U)
1 x	Tinplate housing 55.5mm x 74mm x 30mm

All other components in SMD format:

3 x	1µF / 20V Tant	1 x 18Ω
2 x	10µF / 20V Tant	1 x 47Ω
1 x	22µF / 20V Tant	2 x 75Ω
1 x	0.1µF Foil	2 x 220Ω
1 x	0.47µF Foil	1 x 300Ω
1 x	10µH Choke	5 x 1kΩ
4 x	4.7pF, ATC chip	1 x 10kΩ
1 x	47pF, ceramic	1 x 33kΩ
1 x	100pF, ceramic	
11 x	1nF, ceramic	
1 x	1.8nF, ceramic	

3. COMMISSIONING

The following measuring equipment should be available: Multi-range meter, Frequency meter up to 2.5 GHz, Power meter up to 2.5 GHz and an Oscilloscope.

Initially only the +12V operating voltage is applied. The power consumption should be app. 170mA. The assembly should work immediately and supply a signal of something under 2 GHz at a level of 40mW. The VCO PLL is not locked onto a frequency determined by the two crystals, but instead oscillates at its lowest frequency. If the tuning voltage (max. 30V) is now applied and one of the two crystals switched on (+12V at corresponding connection), then the oscillator locks onto the desired frequency. Should this not be the case, then the VCO is oscillating outside the capture range of the PLL. To check the tuning range, the control circuit should be separated at the transistor T2 (BC848C).

The reference oscillator oscillates at about 9 MHz. This can easily be checked using an oscilloscope at pin-9 of the synthesiser SP5070 (IC5). The maximum oscillator amplitude is set using the 90pF trimmer.

When the synthesiser is brought into operation, a video signal and a frequency-modulated 5.5 MHz sound sub-carrier can now be connected up and the first practical test can be carried out. The FM deviation is set using the two trim potentiometers.

4. CONCLUSION

The output power level of app. 40mW is naturally too low for an FM-ATV station. With regard to increasing the amplification, various publications are available in the relevant amateur radio literature. An output power of up to 10W is relatively attainable, particularly with modern GaAsFET's.

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Readers' Forum

The Book on the Topic:
GPS, Theory and Practice

By B.Hofmann-Wellenhof / H.Lichtenegger / J.Collins

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Crystal Oscillators with Internal DCF-77 Synchronisation

In order to increase the long-term of OCXO's (oven-controlled crystal oscillators), Tele Quarz GmbH are pursuing new paths.

Since a large part of the OCXO's manufactured industrially today are used as base stations for radio networks, re-balancing is very costly, for many reasons, most of which have been pointed out. It is more advantageous to synchronise the reference frequency to the normal frequency transmission of DCF-77 and thus obtain a long-term stability of 10-9. The complete OCXO, with a DCF-77 receiver and frequency range, has been fitted into a very compact housing by TQ with dimensions of 67 x 60 x 49 (mm.). For clear reception, an external aerial is connected up using coaxial cable.

For further details, contact:

Tele Quarz GmbH,
74924 Neckarbischofsheim
Germany

New Magnetic Field Measuring Heads

General Microwave are introducing two new magnetic field measuring heads as an addition to the existing range of power density measuring systems. Models H103 and H105 operate in a frequency range of from 10 MHz to 1.5 GHz, with a measuring ran of from 0.02 mW/cm.2 to 200 mW/cm.2. They thus widen the range of systems with electrical field sensors for the area from 200 kHz to 40 GHz, which are also known as RAHAM's (RAdiation HAzard Meters).

For further details, contact:

EMCO Elektronik GmbH,
82152 Martinsried
Germany

CERA-TRIM

This is the name selected for the new SMD miniature trimmer capacitors from Tekelec Airtronic. They are available in a range of six different types for the capacity range between 0.6 pF and 25 pF (100 pF). The SMD trimmer capacitors have a high Q factor, Q > 3000, can be used up into the GHz range, and are suitable for use at voltages > 250 V Dc and working temperatures from -55 °C to + 125 °C.

For further details, contact:

Tekelec Airtronic GmbH,
Munich, Gemany

Highly Stable Oven-Controlled Oscillators

When the quality criterion for oscillators is "highly stable", then the so-called "oven-controlled oscillators" are always the right choice. KVG (Neckarbischofsheim) are at present offering oven-controlled oscillators of this type for special conditions with:

Stability: -10 to +55 °C. <- ± 2.5 x 10^{-8}
Ageing: 1.5 x 10^{-7}
2 frequency outputs: 5 MHz & 10 MHz
Operating voltage: 12 Volts
Housing dimensions: 63 x 53 x 30mm
Unit price: DM 300

The Book on the Topic of EMC: Metrology, Simulation and Development in the Field of Electro-Magnetic Compatibility (EMC)

Publisher: Harald Fischer,
1993, 308 pages,
Format 16.5 x 23 (cm.), Boxed
ISBN 3-8007-1890-1
Price DM 68

This book attempts to cover the present state of technology in the EMC field, and to demonstrate the software tools which are available today for simulation, by showing how some problems were solved.

The book can be obtained from book dealers or directly from the VDE Verlag.

Software Catalogue

PC based CAD Software

'PUFF' with handbook (CalTech)	£ 18.50
Motorola/SM6MOM supplement	£ 11.00
Siemens S-parameters	£ 9.50
Philips S-Parameters	£ 14.50
HP AppCAD	£ 16.50

Post & packing: UK +£1.50; Overseas +£3.00; Airmail +£7.50

Easy PC PCB CAD	£121.00
Easy PC Professional advanced PCB CAD	£235.00

Post & packing: UK +£5.00; Overseas +£7.50; Airmail +£10.00

PC based Software Collections

CAE No.1 collection (VHF Communications 1/93)	£ 70.00
CAE No.2 collection (VHF Communications 1/93)	£ 70.00

Post & packing: UK +£5.00; Overseas +£7.50; Airmail +£10.00

Disc of the Quarter issue 1/94 - RFD.exe £ 10.00

PC based Circuit Simulation Packages

PULSAR digital circuit simulator	£235.00
ANALYSER III analogue circuit simulator	£235.00

Post & packing: UK +£5.00; Overseas +£7.50; Airmail +£10.00

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KM Publications, 5 Ware Orchard, Barby, Nr.Rugby, CV23 8UF

Tel: 0788 890365 Int: +44 788 890 365
Fax: 0788 891883 Int: +44 788 891 883

PRINTED CIRCUIT BOARDS - KITS - COMPONENTS
for projects featured in VHF Communications

DB1NV	**Spectrum Analyser**			**Art No.**	
PCB	DB1NV 006	IF Amplifier	Ed.2/89	06997	DM 35.00
PCB	DB1NV 007	LO-PLL	Ed.2/89	06995	DM 35.00
PCB	DB1NV 008	Crystal Filter	Ed.3/89	06998	DM 35.00
PCB	DB1NV 009	Sweep Generator	Ed.3/89	06996	DM 35.00
PCB	DB1NV 010	Digital Store	Ed.3&4/91	06477	DM 44.00
PCB	DB1NV 011	Tracking Generator	Ed.2/92	06479	DM 31.00
PCB	DB1NV 012	Broadband VCO 1450 MHz	Ed.4/92	06480	DM 33.00
PCB	DB1NV 013	Broadband VCO 1900 MHz	Ed.4/92	06481	DM 33.00

DB6NT	**Measuring Aids for UHF Amateurs**		**Art No. ED. 4/1993**	
PCB	DB6NT 001	Measuring Amp to 2.5 GHz	06379	DM 36.00
PCB	DB6NT 002	Frequency Divider to 5.5 GHz	06381	DM 36.00

DJ8ES	**23cm FM-ATV Receiver**		**Art No. ED. 4/1993**	
PCB	DJ8ES 001	Converter	06347	DM 22.00
PCB	DJ8ES 002	Digital Frequency Indicator	06350	DM 19.80
PCB	DJ8ES 003	IF Amplifier	06353	DM 15.00
PCB	DJ8ES 004	Demodulator	06356	DM 19.00

DJ8ES	**28/144 MHz Transverter**		**Art No. ED. 4/1993**	
PCB	DJ8ES 019	Transverter 144/28 MHz	06384	DM 36.00
PCB	DJ8ES 020	Hybrid Amplifier 144 MHz	06386	DM 35.00

DF9PL	**High Stability Low Noise PSU**		**Art No. ED. 1/1993**	
PCB	DF9PL 001	30 Volt PSU	06378	DM 19.00
PCB	DF9PL 002	Pre-Stabiliser	06376	DM 20.00
PCB	DF9PL 003	Precision Stabiliser	06377	DM 21.00

To obtain supplies of the above or any PCB's, kits, components, etc., previously advertised in VHF Communications, please contact KM Publications for Sterling prices, etc., at the address shown on the inside front cover of this magazine.

Publications:

UHF Compendium Pts.1 & 2; K.L.Weiner DJ9HO	£27.50
UHF Compendium Pts.3 & 4; K.L.Weiner DJ9HO	£27.50

Post & packing: UK +£2.50; Overseas +£5.00; Airmail +£10.00

An Introduction to Amateur Television; Mike Wooding G6IQM	£ 5.00
Slow Scan Television Explained; Mike Wooding G6IQM	£ 5.00
The ATV Compendium; Mike Wooding G6IQM	£ 3.50
TV Secrets Volume I; ATVQ Magazine	£ 7.50
TV Secrets Volume II; ATVQ Magazine	£17.95
A Guide to 23cm Television; Severnside TV Group	£ 3.50
Just a Few Lines (a history of the birth of the BBC)	£ 5.00

Post & packing per book: UK +£0.80; Overseas +£1.25; Airmail +£2.75

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