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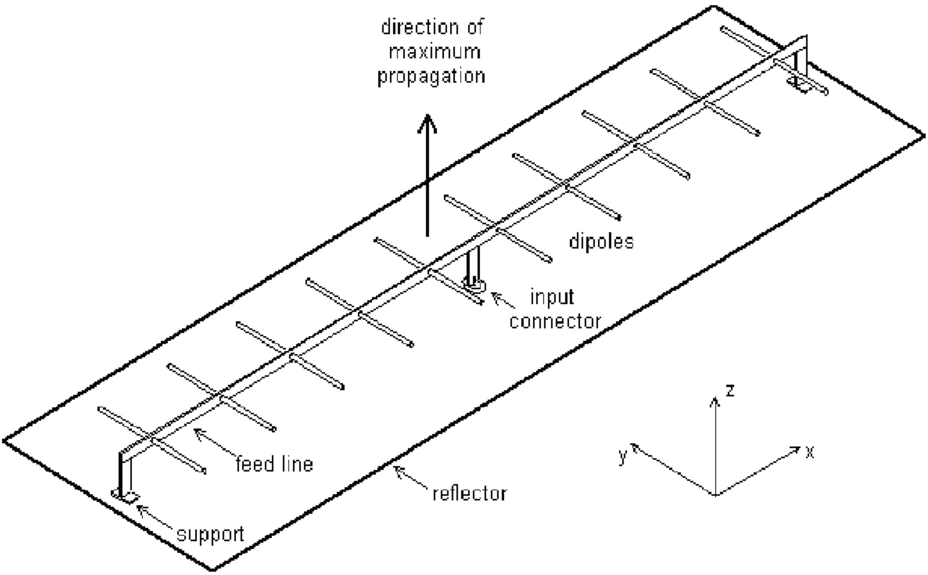

Antenna Array for the 6cm Band

José Geraldo Chiquito

Contents

I was very sad to hear that the American agent for the magazine, Gene Harlan WB9MMM, passed away on 26th November 2008. He published ATVQ magazine, it will continue with new publishers who are also taking on the agency for VHF Communications Magazine - see P60 for details.

I am very pleased to welcome two new authors in this magazine, Marty Singer, K7AYP has been a subscriber since 2002 and José Geraldo Chiquito was introduced to me by one of the magazines long standing authors, Carl Lodström. Both articles are very interesting and document projects that the authors have constructed. I am always interested to hear from anyone who has similar projects that they want to publish as articles.

I have been asked by the RSGB to produce a third book in the Microwave Projects series. If anyone can suggest suitable articles I will be pleased to consider them for inclusion in the new book due for publication early in 2010.

73s - Andy

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Marty Singer, K7AYP

Digital T-R control sequencer

1.0 Introduction

For the ham operator, a transmit/receive (T-R) sequencer is an electronic circuit that controls communication circuits going from receive-to-transmit mode, or to return from transmit-to-receive. This control is essential if sensitive circuits are to be protected from RF, transients and to conserve power, heat, etc. Conventional T-R approaches document analogue component designs to generate timing, and control [1] [2] [3] [4]. One article has used a microcontroller [5].

This article presents a reliable system T-R sequencer with flexibility for unique application, and creative modification. Content includes a simple digital state machine controlling coaxial relay circuits, and system control circuits. Non-critical small, and medium scale digital ICs are used together with solid-state components.

2.0 T-R sequencer - block diagram

Two multiplexing integrated circuits, and one synchronous medium scale counter chip form a state machine that sequences through unique event widows where actions move RF system circuits between receive and transmit modes. Eight, individual event windows are stepped through from receive-to-transmit mode and the same number back to receive mode. Six of these event windows control RF system conditioning to/from receive and transmit modes. Two, are used for state machine control. Completed activity in each of the six event windows must provide a positive completion signal back to the state machine before it continues sequencing. Without a signal, the state counter does not change state. Sequencing stops regardless of the mode change in process providing “fail safe” performance for the system under control.

Fig 1 shows the T-R sequencer block diagram. The sequencer control engine represents the three logic devices that generate six unique, clocked, event windows after *Reset occurs at power up. PTT or a manual switch initiates control engine sequencing. Event #0 represents the receive mode idle condition. Event #1 represents one-of-six event window signals, shown to operate a pulse latching coaxial RF relay. A *C1fini signal responds back to the control engine indicating success or a failed operation. The control engine logic promptly initiates

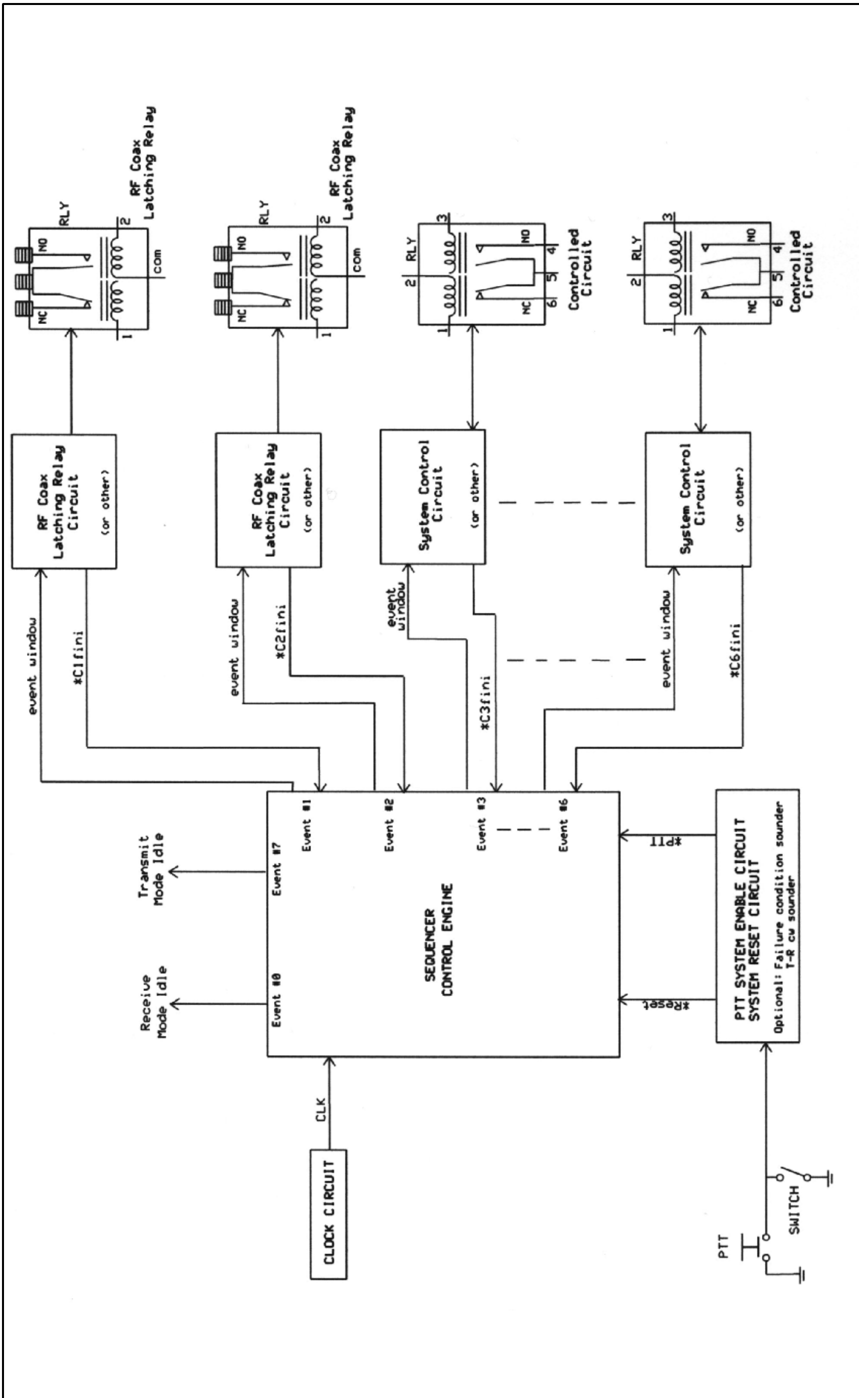

Fig 1: T-R sequencer block diagram.

Event #2 event window, shown to control a latching relay circuit ending with *C2fini back to the control engine. This orderly process continues for six state sequences. At the end of the sixth event window (shown operating a system control circuit) the sequencer engine moves to Event #7 and becomes idle in sequencer transmit mode, completing the actions necessary for an RF system receive-to-transmit transition.

A change in the PTT or manual switch position initiates the control engine to decrement its sequencing events - Event #7 to Event #0. Completion puts the sequencer, and RF system it controls back in receive mode idle.

Several conclusions and design requirements are identified with this architecture:

- There are three primary functional modules: a state machine circuit, a coaxial relay control circuit, and system control circuit (non-coaxial) that execute an RF system transition from receive-to-transmit, or the return back to receive mode.
- The order in which a particular event window activity occurs need not be the same for a receive-to-transmit sequence, as for transmit-to-receive transition.
- More than one activity can be performed in any event window, reducing sequence mode transition time, and possibly parts count.

The reader should not limit thinking to descriptions detailing a particular relay circuit operating during a particular event window. Although workable, this is done to explain circuit designs, and sequencer operation.

3.0

T-R sequencer control engine - circuit description

Fig 2 shows the core control engine parts:

- U2, an up/down counter 74LS169 used as a state machine counter
- U3, a 74HCT138 state select decoder
- U4 a 74HC151, one-of eight state condition multiplexer

The CLK signal on U2, pin 2, together with the Advance signal, pin 7 enables increment or decrement counting in a direction controlled by the RX Mode polarity on pin 1; logic 1, U2 increments, logic 0 it decrements. The counter parallel-load function is not used (pins 3, 4, 5, 6). Pin 9 is the power-reset pin, used at power-up.

Decoder chip U3 input pins 1, 2, 3 connect to the three state machine counter outputs Q0, Q1, Q2 (pins 14, 13, 12,) with binary weights “1”, “2”, “4”. The one-of-eight decoder chip detects counter binary states, and changes a single output – *C0, or *C2, …through to *C7 for each counter state change.

Multiplexer chip U4 input pins 9, 10, 11 connect to the same counter outputs as device U3. With each counter state change, only one logic 0 multiplexer input at pin 4, or 3, or 2, or 1, or 15, or 14, or 13, or 12 (D0, or D1, or ”Dx”) is used to generate a logic 0, Advance signal output. The CLK and Advance again enable the counter to change state in a direction determined by RX Mode.

The D type flip-flop, R-T mode control F/F, U1A, stores the active receive or transmit operating mode. Latched out-

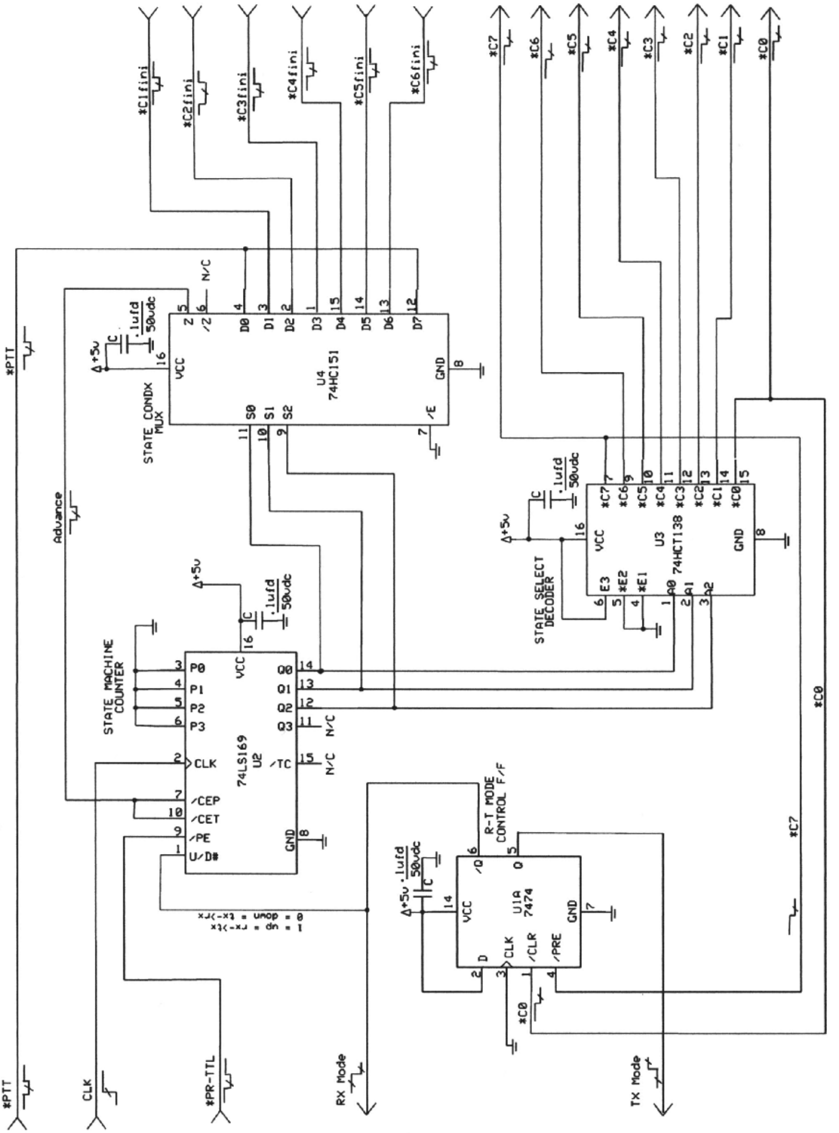

Fig 2: T-R sequencer control state engine.

puts (pins 5,6) steer circuit operations during a receive-to-transmit, or transmit-to-receive transition.

4.0

Putting the three chip state machine to work

Receive-to-Transmit sequence

Power-up signal *PR-TTL at counter U2 pin 9 resets output pins 12, 13, 14 to logic 0. This puts a logic 0 level only at event window *C0, device U3, pin 15. *C0, is the sequencer state of receive mode idle. State condition multiplexer U4 enables only the D0, pin 4, input due to counter logic 0 outputs on S0, S1, S2. Other Dx inputs are disabled. Device U1A is asserted in the clear state at pin 1 by *C0 so RX Mode, pin 6, is logic 1, pin 5 is logic 0. And, other applicable relay, and control circuits are reset at power up.

The state machine is now in receive mode idle, including necessary relay and switching circuits controlled by the sequencer (later circuit descriptions include necessary power up reset). This places the RF system controlled by the sequencer in receive mode.

Sequencing from receive-to-transmit mode begins when *PTT activates a logic 0 at device U4, pins 4, 12. The source for *PTT can be from an IF radio, manual switch or other control device (later descriptions show circuit design and application). The multiplexer passes *PTT through to an Advance signal on pin 5 and to counter U2 pins 7, 10. Advance and a positive-edge of CLK increments the counter one state.

Decoder U3 inputs A0 through A2 “see” the one-count increment, and change output *C0, pin 15, to logic 1, and event window *C1, pin 14, to logic 0. Counter outputs at multiplexer U4 now enable sensing for a logic 0 from signal *C1fini at pin 3.

The sequencer is now in event window *C1. Planned activity for a receive-to-transmit mode transition is done during this time period. The state machine waits for a *C1fini signal at U4, pin 3. Any/all peripheral action(s) taken by sequenced circuits in the *C1 event window must end with an active *C1fini signal at pin 3, or the state machine stops in the *C1 event window until power loss, or a power reset. Later sequence-controlled circuits show implementation of completion signals for *C1fini.

Upon receipt of *C1fini, an Advance signal from U4, pin 5 and CLK again increment the machine counter one state. Similar to previous description, C1, pin 14 goes to logic 1, C2, pin 13 is asserted logic 0 at device U3. This opens a second, event window (*C2) for another, different sequencer state activity. Multiplexer device U4 will sense for *C2fini at pin 2 input before incrementing the state machine counter.

Cycling continues if each *Cx event window activates necessary peripheral relay and control circuits which, in turn, respond back to the state machine with a *Cxfini signal.

When *C7 event window is reached the state machine stops sequencing in transmit mode idle. No additional peripheral circuits need attention. Satisfactory transition from receive-to-transmit mode has occurred. Signal *C7 also presets U1A, latching RX Mode to logic 0 at U1A, pin 6. U2 up/down input pin 1 changes state, steering the machine counter to decrement when the next group of Advance signals clock the counter back to receive mode.

Transmit-to Receive Sequence

Transmit mode ends when *PTT changes logic level at U4, D7 input. This can occur when a PTT or manual switch is

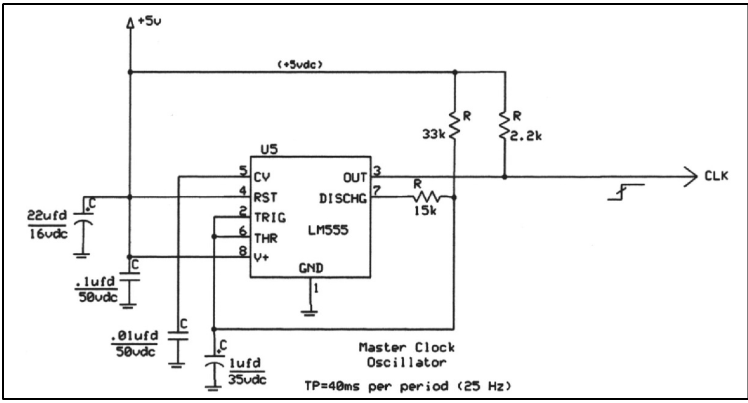

Fig 3: T-R sequencer clock oscillator.

released. An Advance signal with CLK decrements the state machine counter, one state. Following a sequential procedure similar to the receive-to-transmit transition, event window *C6 is activated. During each event window actions are performed to step the RF system back to receive mode. Cycling from *C6 to *C0 occurs if their associated *Cxfini signal is returned to initiate an Advance output at U4, pin 5. Successful event window operations generate *Cxfini signals, decrementing the clocked machine counter with the Advance signal. At *C0, the U1A, pin 1, latches the R-T flip-flop, returning RX Mode to logic 0, and idle receive mode.

System clock

The T-R sequencer clock generator is shown in Fig 3. The circuit consists of a single LM555 device, wired as a free-running oscillator. Circuit description for this configuration is widely published so will not be duplicated here.

The clock frequency is very low, making layout non-critical for all circuit operation. Counter states change on the CLK rising edge (logic 0 to 1). The period is important for toggling pulse-latching relays. Operating parameters for coaxial relays found at a local flea market have the following specification:

- Latching signal pulse width: 20 milliseconds to 1 second.
- Switch time to settle: 15 millisecond, maximum.

To meet a 35 millisecond operating time, CLK was made 40 milliseconds, minimum. This defines the time of each, *Cx event window (except idle states *C0, *C7) made by the state machine counter. The total sequence time to complete a receive-to transmit mode transition (or, visa-versa) using a 40 millisecond clock period, without sequence failure, and beginning from the receive mode idle condition when the *PTT signal is asserted:

- Assume *PTT is asserted just after a rising clock edge (worst case). Sequence waits one, 40 millisecond clock period before counter U2 advances from *C0 to *C1.
- There are (six event windows) x (40 milliseconds per window) = 240 milliseconds.
- Relatively no time is consumed in event window *C7 to preset R-T flip-flop (TX Mode to RX Mode) for transmit mode idle.

Total time is 280 milliseconds, not conducive for QSK operation, but nobody the writer knows requires fast break-in CW on the 23cm ham band.

Clock frequency tolerance, stability, and duty cycle are not critical except for meeting latching relay requirements.

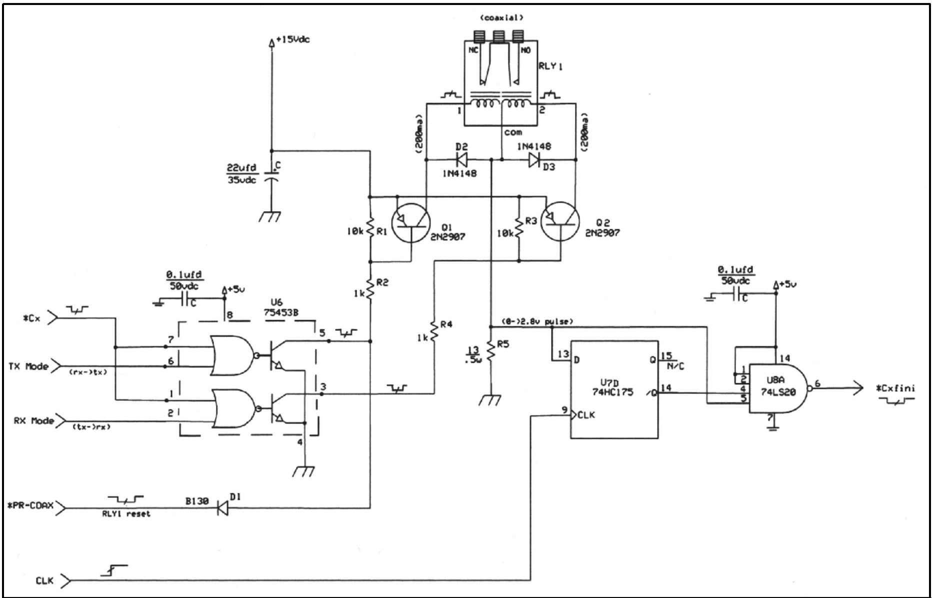

Fig 4: T-R sequencer single pulse latching coaxial relay.

Other devices, circuits can be used with the capability to meet output voltage amplitude, frequency, and source/sink current requirements for all clocked devices.

5.0

Sequencer control circuit descriptions

Single RF Pulse Latching Relay with “Fail Detection” Circuit

Typically, pulse-latching coax relays are sequence-controlled to switch a transverter antenna from a receive down-converter input to the final amplifier output. Another coax relay switches the IF radio to either a down-converter output or input an up-converter/amplifier path. For added protection, an RF coaxial relay could switch the down-converter input into a 50Ω termination in transmit mode. All relays are controlled during selected *Cx event window intervals. Each relay drive circuit provides a *Cxfini signal back to the state machine to advance the sequencing process.

In Fig 4, latching relay RLY1 has important circuit design specifications (besides RF parameters):

- Latching coil voltage, pulse, typical: 12V
- Current, each coil: 200mA

The coaxial relay has no built-in “fail safe” feature. Each latching coil has outputs “1”, “2”, together with a common (”com”) tie between coils. Shunt diodes D2, D3 prevent back-emf issues. Each independent coil is wired to a PNP transistor collector Q1, Q2 respectively. Resistor pairs R1-R2, and R3-R4 bias the transistors “off” until their base is switched to ground by an NPN transistor in device U6.

Device U6 has two, NOR logic gates integrated with open-collector NPN tran-

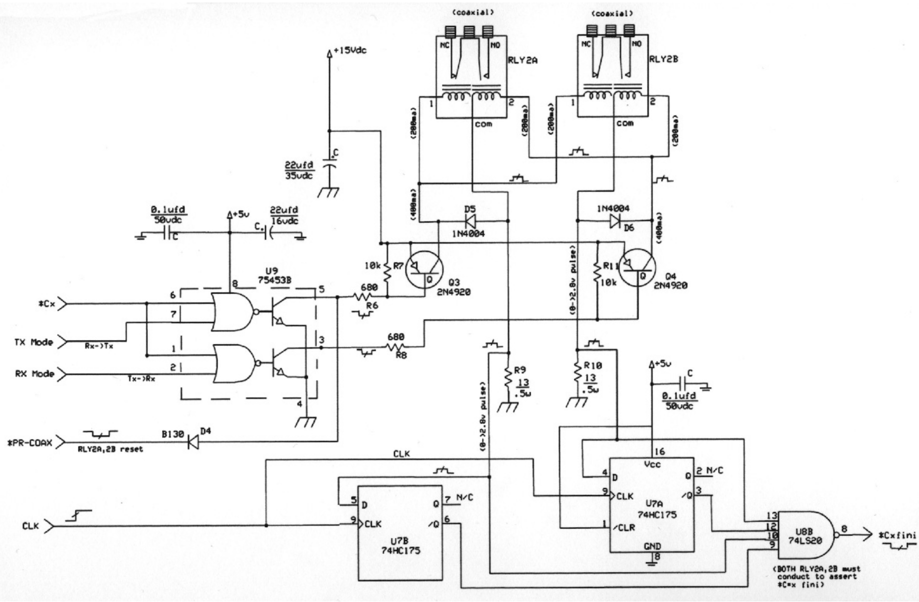

Fig 5: T-R sequencer dual pulse latching coaxial relay.

sistors having a common emitter ground. The logic gates are powered with +5V making them compatible with other logic devices. NOR gate inputs directly interface to the state machine decoder outputs, and R-T flip-flop RX Mode, TX Mode signals. An event window controls U6, and a PNP transistor to switch a 200mA, 40-millisecond current pulse from +15V through a latching coil, out "com" to ground through resistor R5. Measuring a V_{ce-sat} of about 0.2 volts, with a +15V supply voltage, and 12V relay coil, a +2.8V pulse signal is present across R5. This is enough for digital logic compatibility.

Device U6 inputs steer RF relay operation. A selected *Cx event window with RX Mode or TX Mode control which side of the RF relay will operate during a receive-to-transmit, or transmit-to-receive mode change. Either coil activations put a voltage level change across resistor R5 during the event window.

At T-R sequencer power-up, relay RLY1 is initialised in receive mode by diode D1, wire-ored to U6, pin 5. Signal *PR-COAX connects the diode cathode to the sequencer power reset circuit (described later).

A voltage level change across resistor R5, is used to determine circuit fault for these RF relays that do not feature fail-safe capability. Consider possible circuit faults that would not cause a change across resistor R5 during an event window period:

- Open or shorted relay coil; either side
- Open or shorted collector-emitter; Q1, Q2
- Shorted or open collector-emitter junction; U6
- Bad solder joint, +15V supply path through R5 to ground; either side
- Faulty R5 resistor

- Loss of +15V supply
- Shorted back-emf diodes; D2, D3
- Non-operating circuits inside device U6

Devices U7D, U8A detect a voltage change across resistor R5. If change occurs, regardless of which latching coil activates, a *Cxfini signal goes to state machine multiplexer U4 to continue sequencing. U7D, is one-of-four D flip-flops in a 74HC175 package. Gate U8A is shown as ½ of a 74LS20 4-input NAND gate (other gate used in Fig 5). Without an active event window at U6, there are zero volts across R5 so U7D, pin 13, and U8A, pin 5 are at logic 0. The CLK signal sets the /Q output, pin 14, to logic 1. At this time the two logic level inputs at U8A cause a logic 1 at output pin 6.

For a voltage change across R5, U8A pin 6 changes to logic 0, sending a *Cxfini signal back to the state machine. The state machine will move to the next event window, changing the R5 voltage back to zero, in turn changing U7D, and U8A back to their original states, ready for the next event window operation.

Without a voltage change across resistor R5, U8A output pin 6 does not change to logic 0. No *Cxfini signal is generated, and state machine sequencing is halted at the specific event window that failed.

Dual RF pulse latching relay with “fail detection” circuit

Fig 5 shows a dual coaxial relay set up. Driver transistors Q3, Q4 with U9 duplicate the single relay circuit in Fig 4, but with higher collector current PNP transistors since twice the single relay current is switched (2N4920 do not need heat sinks). A detection circuit is constructed for each relay. RLY2A is sensed at resistor R7 with U7B and one-half of a 4-input NAND gate U8B, pins 9, 10. Relay RLY2B is sensed at R8 with U7A and U8B, pins 12, 13. Both relays and circuits must operate satisfactorily to initiate a *Cxfini signal at U8B, pin 8. The failure detection criteria for both are the same as single relay RLY1 described earlier. Schottky diode D4, at U9 pin 5 presets both relays after power-up with *PR-COAX.

A 75453 is used to drive the relay driver transistors. RX Mode and TX mode control lines together with a *Cx signal steer both relays like the single relay case. U9, (pins 1, 6) could receive different *Cx signals from the 74HCT138, device U3 Decoder, to switch the relay pair at different sequential times during the receive-to-transmit, and transmit-to-receive sequences. This also applies to the pulse latching coaxial relay circuit.

6.0

System control circuit description

To compliment pulse-latching coaxial relays, a small, pulse-latching relay is used to control non-coaxial control operations. A +5V latching pulse makes it compatible with +5V logic operating voltage. This control circuit is used to enable/disable a system module operating voltage or on/off operation circuit. Where applicable, switching +15V through RLY3, pins 4, 5 guarantees individual module voltage regulators have enough input voltage, including their minimum input dropout voltage. When RF system modules are sequenced they typically remain on, or off until sequenced back by the state machine. These relays provide this “memory element” requirement. RLY3 contacts have 30V DC, 2A specifications.

The +5V T-R sequencer supply connects to RLY3, pin 2. External diodes D8, D9 protect against back-emf conditions. For this application, relays with built-in back-

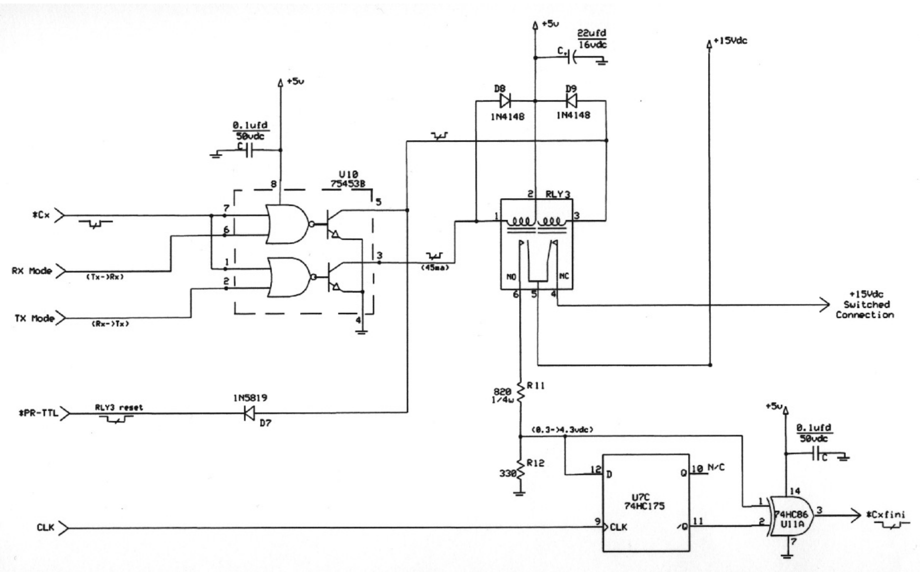

Fig 6: T-R sequencer sequence control circuit.

emf protection diodes must not have their anodes connected to pin 2, cathodes connected to pins 1, 3. Individual NPN transistors in U10 connect to each relay coil at pins 1, 2, and conduct a 40mA latching pulse when logically controlled by RX Mode or TX Mode together with a selected *Cx event window. The *PR-TTL signal at diode D7 presets relay contacts during power up.

Fault detection is also used with this control circuit design. One-quarter of a 75HC175 (U7C) with a 74HC86, Ex-Or gate (U11A) detect the voltage level change similar to that explained for co-axial relays. Resistors R11, R12 form a voltage divider, reducing +15V at relay pin 6 down to logic level compatibility. Active voltage level changes appear on pin 12, U7C, and U11A, pin 1. The CLK signal at pin 9 toggles U7C pin 11 to generate a logic 0 *Cxfini pulse, one clock period wide at U11A, pin 3 that continues state machine sequencing. Resistor string R11, R12 may need to be exchanged with RLY3, pin 4 to accommodate the external module being controlled.

7.0

Presetting relays

After system power up in receive mode, power is typically connected to a down-converter (and other functions) supporting receive mode. RF latching relays are preset so the antenna connects to the down-converter, and IF radio input. Transmit related modules are disabled at power up, but require +15V during transmit mode (e.g. up-converter, driver, final amplifier control). After determining how connected modules wire to relay contacts, a *PR-TTL, or *PR-COAX reset signal must be connected to place a

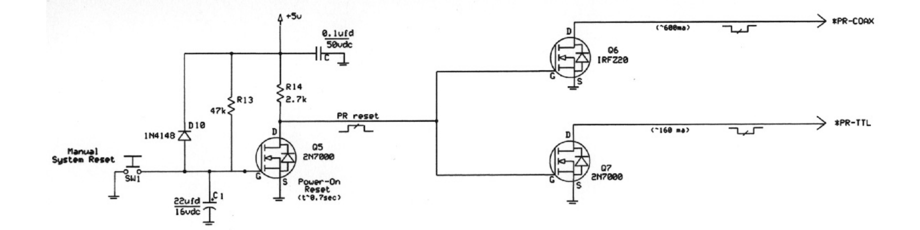

Fig 7: T-R sequencer reset - power-on, manual.

relay in its correct initialised condition.

For receive mode control circuits RLY3 pin 5 could connect +15V through pin 4 to the down-converter DC power input (Fig 6 shows relay contacts in receive mode). Diode D7 is wire-ored to U10, pin 5, and *PR-TTL correctly initiates the relay at power up. A system transmit module could require exchanging relay pin 4 and pin 6

Fig 4 shows a typical case with diode D1 wired to U6, pin 5, setting up RF coaxial relay RLY1 for receive mode when *PR-COAX reset signal is asserted.

8.0

Sequencer support circuits

Power Up Reset

Fig 7 shows a power-up circuit. FET Q5 is an active switch, controlled by R13, and capacitor C1. Diode D10, rapidly discharges C1 when 5V power is removed. Manual reset switch SW1 also discharges C1 for as long as the switch is asserted. Switch contact de-bounce circuitry was not deemed necessary. For R13, C1 component values shown an approximate 700 millisecond logic 1 level occurs after 5V power is initiated. The active PR reset pulse must be long enough for +5V, and +15V power supplies to turn on, and settle. Also consider the maximum pulse width allowed for the relays.

Two, FET transistors are driven by Q5. The writer's sequencer uses one coaxial relay circuit shown in Fig 4, and one dual relay design in Fig 5. Device Q6 output connects to all *PR-COAX system reset signals so approximately 600mA is shorted to ground. The IRFZ20 is an overkill, but was available in the writer's part stock. An N-Channel FET with at least a 20V V_{ds}, and 2A I_{ds} is satisfactory (safe "ceiling"). A reasonably low R_{ds}

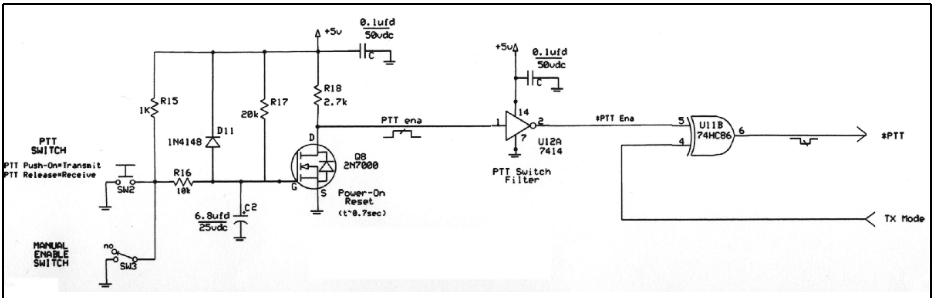

Fig 8: T-R sequencer initiation - PTT, manual.

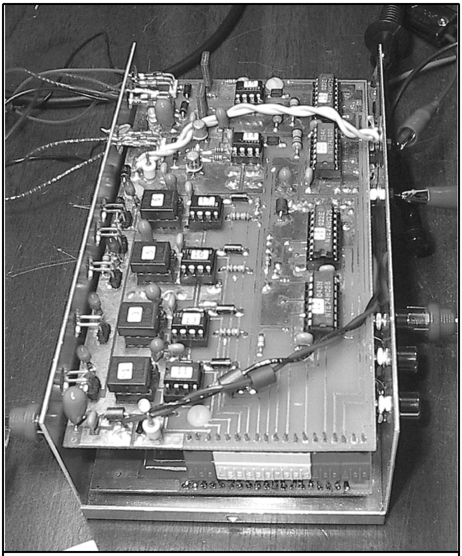

Fig 9: The author's T-R sequencer.

specification is desirable for adequate voltage drop across the relay coils. With all *PR-TTL lines wired to Q7-drain initialises four control relay circuits. Approximately 160mA conducts to ground.

PTT or switch control

The circuit in Fig 8 initiates receive-to-transmit, and transmit-to-receive state machine sequencing. Switches SW2, or SW3 enable a PTT-ena signal from Q7 drain. PTT SWITCH SW2 is assumed to be located remotely at an IF radio, but is shown to explain circuit operation. Manual enable switch SW3 toggles and holds the circuit until switched back to the normally "open" state. Use of this switch can be for system testing or when operating CW mode.

The Q7 circuit is similar to the power-up reset circuit in Fig 7, except for the R17, C2 time constants. Resistor R15, R16 and C2 form a simple filter to help eliminate false/noisy switch activation. The voltage level of PTT-ena is equal to the time duration SW2, or SW3 is asserted. The signal's rising edge is exponential at Q8 drain, so a 74HCT14 gate with Schmitt trigger input is used for signal inversion and to shape *PTT-ena.

Control of state machine activation (at multiplexer pins 4, 12) is done by *PTT, a U11B Ex-Or output signal developed by *PTT-ena together RX Mode. The state machine senses the *PTT active level at multiplexer chip U4, pins 4, 12. These inputs associate with receive mode idle or transmit mode idle. So PTT SW2, or manual switch SW3 conditions are not detected during sequenced transition to/from these two operating modes, only after the state machine completes a full transition cycle (without fault). Sequencing stops if either switch remains asserted after the state machine reaches the activated operating mode idle.

9.0

Construction tips

Implementation

There are many design, and component type modifications possible. Many logic families are used because of device availability in the writer's parts stock. In that regard, it does show the flexible implementation capability for this project. If other logic functions, or families are substituted, take care to satisfy device input/output logic level compatibility, and output current source/sink specifications (i.e. "unit loading"). This is especially true at logic and discrete component interfaces.

Construction

Circuit and component layout is not critical, helped by the 25Hz clock frequency. The author's unit is shown in Fig 9. Besides etched circuit board(s), a plated-through hole, 0.1inch hole centre, strip board will do nicely. It is suggested that logic device ground be separated from relay circuit grounds, until they connect at one board, and chassis enclosure location (i.e. "star" grounding). This reduces relay switching ground currents from possibly effecting normal logic device operation. All component packages, and circuits need good capacitor bypassing. Entering power supply voltages need to be properly bypassed. A ferrite bead on power wiring might be a good precaution. The enclosure should be metal, not plastic, since the sequencer will most likely be near RF fields from the system modules it controls.

Interconnection between coaxial relays and sequencer circuits will likely involve connectors and cabling to remote system locations. For circuits not needing remote location control, wiring to RF system modules can loop-through connect at the sequencer's physical location.

10.0

Appendix

10.1 Test and debug

Testing the T-R sequencer is not difficult because the ordered steps of the state machine help identify faults or proper operation at particular circuit locations. Most often, only a simple VOM meter, or low frequency scope is required.

After power is applied verify that the power up reset circuit output is correct. The drain pins of Q6 and Q7 should not be at ground after about a 700 millisecond power-up. The manual reset switch SW1 should generate 700 millisecond reset pulses whenever activated. Measure all relay contacts for their proper preset condition. Check: state machine counter Q0, Q1, Q2 outputs for logic 0 levels, decoder U3, pin 15 (*C0), and multiplexer U4, pins, 4 and 12 for logic 0 levels. Also verify U1A is correctly initialised with logic 1 at pin 6. Basic start up conditions is achieved at this point.

Press and hold the PTT switch (or toggle SW3) to begin event window cycling from receive idle to transmit mode idle. Then measure logic 0 at U3, pin 7 to verify the decoder (and sequencer) reached transmit mode idle. If not, look at each *Cxfini input pin on the multiplexer, U4 (D1 to D6). If one is static at logic 0, this indicates the Advance signal connection to Counter U2 was not made. If all are at logic 1, this might indicate that a *Cx event window did not occur, or did not operate its associated sequenced (relay) circuit. Trace to the circuit connected to the specific *Cx window fixed at logic 0 to determine if it is controlling the relay, and generating a fault detection response. After any receive-to-transmit issues are solved, release the PTT switch and verify complete/correct transition back to receive mode idle.

"Lock-up" conditions on U4 inputs, and event window outputs at U3 are very useful in identifying where to begin solving problems. During debug, using the PTT switch with the manual reset switch helps narrow fault location. Note that some faults can cause continuous current flowing through a relay coil not intended for continuous operation. Locating LED diodes throughout might be helpful for debugging, and monitoring circuit operation.

10.2 Circuit modifications and enhancements

Functional modifications

1) Fig 4 shows a *Cx signal operating the relay circuit for receive-to-transmit, and transmit-to-receive operations (connections to pins 7, 1). This need not be the case since U6 pins 1, and pin 7 allow independent control for different event windows coming from the decoder chip U3. This applies to all relay circuits.

2) Implementing more than one operation in any event window could re-

duce mode cycling time, and possibly parts count (refer below to reducing number of event windows).

3) If fault detection is not used, remove pulse sense resistors, and sense logic. Ground inputs on the multiplexer chip connected to the fault circuits being removed. The state machine will cycle through such states, without stopping.

4) Another modification reduces the number of event windows, and sequencer transition cycle time. Refer to Fig 2:

a) At U4: remove unused connections Dx inputs, starting with D7, toward input D1. Keep *PTT connected to D0 (pin 4). Remove the *PTT signal line from D7 (pin 12), connecting it to the next higher Dx pin just above the highest input used, and connect unused "higher order" Dx inputs to +5V.

b) At U3: remove wires connected to *C7, pin 7. Remove connects to all *Cx outputs that will not be used, and that correspond to unused U4, Dx input pins done in part a). Connect wires that were at U3, *C7 pin 7, to the last *Cx output to be used by the state machine. Leave unused *Cx outputs not connected.

c) At U2: no changes required. The counter will terminate when incremented or decremented to the newly wired U4 *PTT inputs. When complete, there should be a U4 *Cxfini input associated with a *Cx event window output at U3, only fewer of them. Fewer event windows are used to reach transmit mode idle, or back to receive mode idle, accounting for shorter sequencer cycling time.

10.3 Enhancements

Fault condition buzzer

The original sequencer design had no indication when the state machine stops due to a detected fault, or possible component failure. A "dead-man" timer circuit was added with a fault indicator. The schematic is in Fig 10. Ex-Or gate U11C, has *C0, and *C7 inputs, creating the *cycle-stop signal to timer counter U14, pins 1, 2. The timer counter is held reset during receive idle, and transmit idle. When *C0 changes to logic 1, state machine activity begins, so *cycle-stop releases the timer counter reset. Count increments using CLK at U14, pin 14. Terminal count is wired to occur if *C7 does not occur in a time longer than one complete, six-event window sequence (approximately 320 milliseconds). If *C7 is reached before terminal count, reset is applied to the timer counter. The process repeats itself; initiating on *C7, terminating on *C0. Beyond terminal count, QD output, pin 12 continues to cycle. Output QD connects to a noisy buzzer, announcing a fault, and sequencer shutdown.

Resistor R19 sets the buzzer sound intensity. The timer counter (and noise) stops when the manual reset switch, SW1 is activated.

Further improvement might be to automatically reset the sequencer at terminal count, stop further sequence operation, but let the fault buzzer continue to sound until SW1 is activated.

T-R CW sounder

Another circuit was added to sound familiar characters in Morse code (cw) when a successful transition to receive or transmit mode occurs.

The circuit schematic is shown in Fig 10. It consists of U13, a single 74HC165, parallel-data-in, serial-data-out shift register. A bit map was made to determine common, and unique logic levels required to generate Morse character "T", and character "R". Each shift register cell is used to represent a dash, dot or space for the characters. Parallel inputs P0 – P7 (pins 3,4,5,6,11,12,13,14) con-

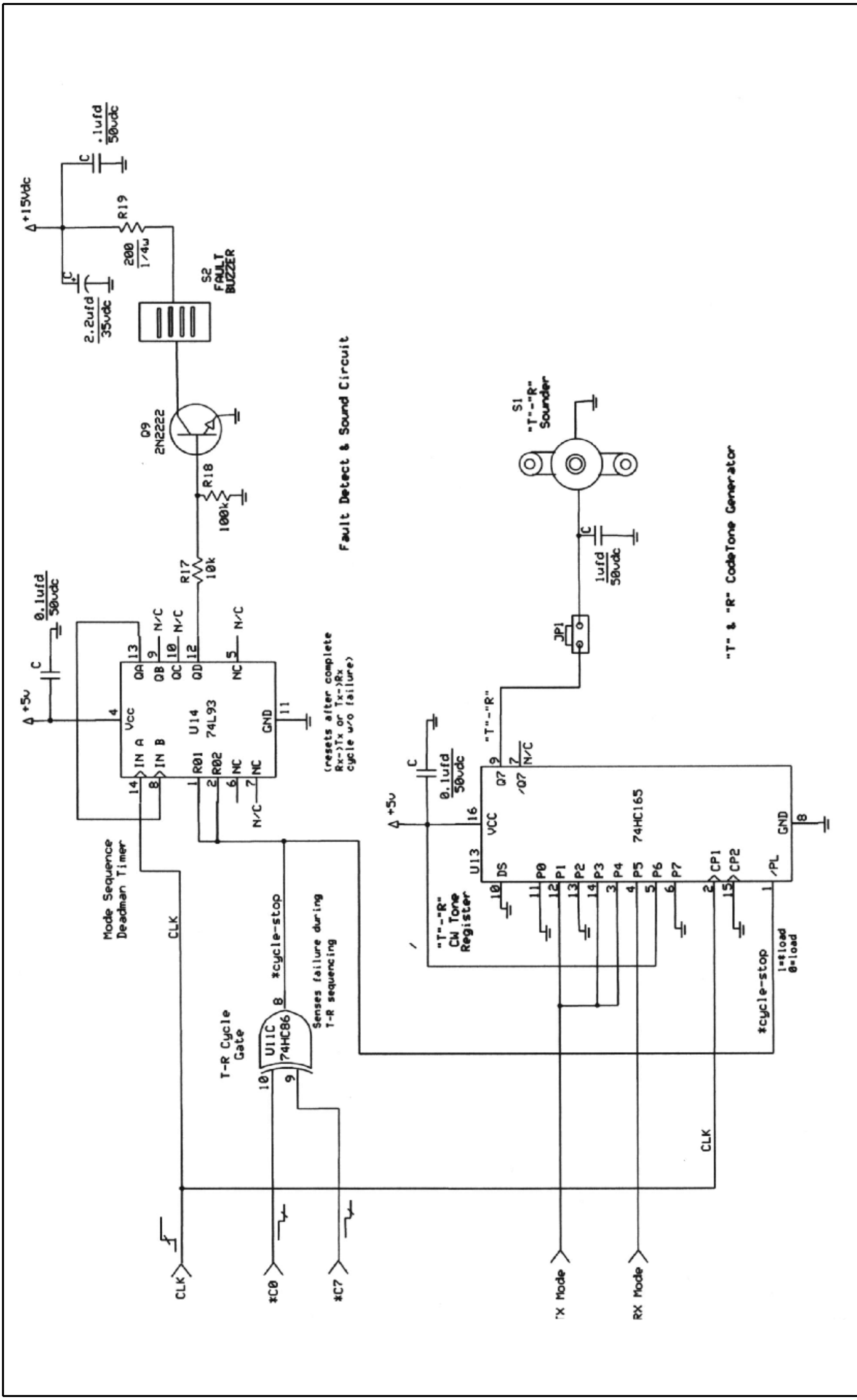

Fig 10: T-R sequencer enhancements.

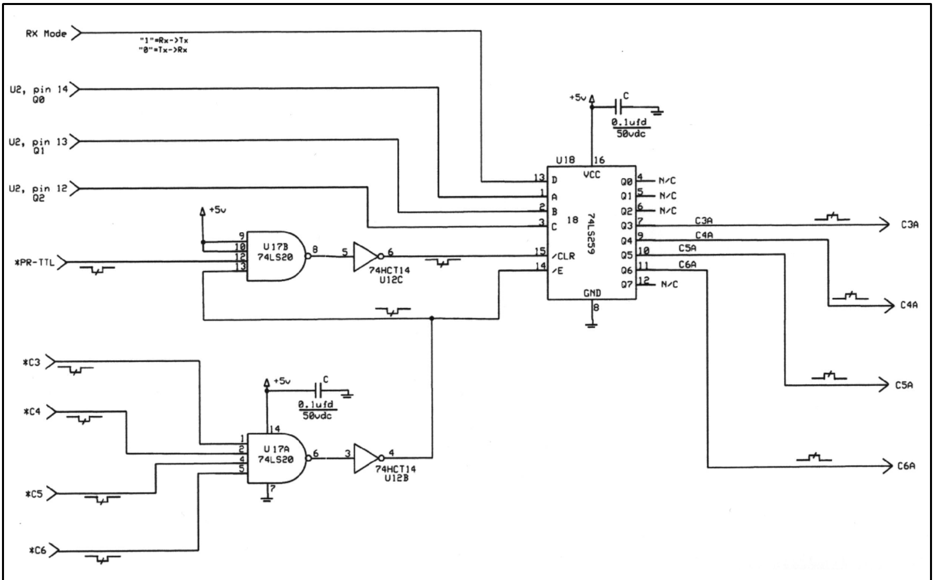

Fig 11: T-R sequencer solid state sequencer control.

nect to either ground, or +5V for dash and dot elements common to both characters. Where mapping shows different character elements needed in the same register cell, TX Mode, or RX mode control lines are used depending on sequence ending in receive or transmit mode. The *cycle-stop signal determines if the shift register is parallel loading character elements, or serial data shifting the Morse character out at Q7, pin 9. Note *cycle-stop is a function of *C0, *C7 signals defining receive and transmit idle states. Output Q7 (pin 9) goes through a 2-pin jumper JP1 before connection to the sounder, S1.

If a successful transition is made to transmit mode idle, the single character "T" is sounded by S1. Reaching receive mode idle sounds an "R" Morse character. Jumper JP1 is available for those desiring peace/quiet.

Solid state sequencer control and circuit

A design is presented in Fig 11, and 12 for a more solid state approach to this T-R sequencer design. This could at least replace non-coaxial relay circuits shown in Fig 4, but this circuit design is intended to replaces only system control circuits (Fig 6).

The design includes U18, an eight bit addressable latch device, 74LS259. It is capable of storing a logic signal present on input D, pin 13 into a unique storage location selected by three address input lines A, B, C wired directly to state the machine counter pins 14, 13, 12 respectively. Control signals on the /CLR, pin 15, and /E input, pin 14, enable a storage operation.

U17B and U12C form an Or logic function. Input *PR-TTL resets U18 outputs to logic 0 at power up. The other Or input, pin 13, comes from another Or gate made from U17A, and U12B. A logic 0 input signal at U17A, pins 1,2,4,5, sends a signal to U17B, pin 13, and the /E input for a *C3, or *C4, or *C5, or *C6 event window generated by the state machine decoder, U3. At this

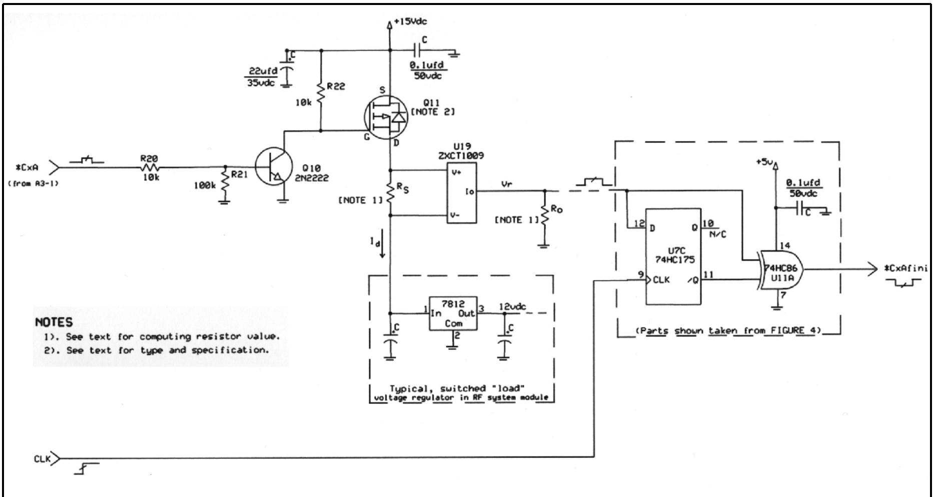

Fig 12: T-R sequencer solid state control power circuit.

time, the logic level at U18, D input, is stored into addressed storage locations.

Logic signal RX Mode at the D input is logic 1 throughout a receive-to-transmit transition. This logic level is stored into U18, Q outputs, and used to activate switches controlling power, enable/disable signals at the RF system modules.

When sequencing back from transmit-to-receive mode, logic 0 level RX Mode is loaded into each addressed memory location.

Like the latching relay, U18 storage locations fill the "memory" requirement, keeping event window tasks enabled/disabled during sequencing. Unlike the original design using 75453 device logic, this implementation only allows the same event window to control a task at the same sequence time in a receive-to-transmit transition, as in the transmit-to-receive mode transition.

As shown, only U18, Q3, Q4, Q5, Q6 output pins 7,9,10,11, are used. These correspond to *Cx decoder outputs, described earlier as operating non-coaxial relay circuits. Any four of the six sequencer states not using coaxial relay switching can be wired into this design. But for continued discussion, event windows *C3 through *C6 are assumed to be the applicable event windows.

To interface with external RF system modules a possible switching circuit is shown in Fig 12. Output from a U18, *CxA connects to a driver made from R20, R21, Q10, and R22. When a logic 1 asserts at R20, device Q10 saturates, pulling Q11 gate to near zero volts, and P-Channel FET Q11 conducts I_d current to a load. Q11 is normally disabled after power up because of the driver's input signal polarity from U18. If I_d current must flow after power up and be turned off during receive-to-transmit cycling, put an inverter gate between the necessary U18 output and R20, or change the driver design.

To show what might be a typical application Q11 drain connects through U19 to a remote +12V, 3-terminal regulator. Because of the +15V supply little voltage drop can be tolerated at the regulator input for its satisfactory output regulation. For lower output voltage 3-terminal regulators this is not an issue, but in all

Table 1: Regulator output for values of R_s and R_o.

R_s Ω	R_o Ω	V_r V	I_c A +12V regulator O/P
0.43	3.3k	0.36 - 3.52	0.24
0.43	3.3k	0.36 - 3.10	0.38

cases the 3-terminal regulator low-drop-out voltage specification must be met.

This application requires a low R_{ds}. Many P-Channel devices have an acceptable I_d current specification. A low V_{gs} off makes device control easy. Switching speed is not important, as is BV_{dss} if above +20V for reliability. The writer's part box furnished the following device specs: V_{gs} = 4.5V off maximum, I_d = 8A, R_{ds}, on = 0.4Ω, max, TO-220 package [7].

Continuing fault detection philosophy used for this T-R sequencer, device U19 (ZXCT1009) was selected to sense I_d current switched into the +12V regulator input. This part is a "High-Side Current Monitor" often used to monitor personal computer, and electronic system power supplies [6]. The device senses I_d current through resistor R_s located between what is considered a +15V "source" (Q11 drain) and "load" (regulator input). A small R_s voltage drop provides a logic level signal V_r across R_o. An example for determining R_s, and R_o values:

Conditions:

a) +15V power supply is the "top side" circuit voltage.

b) A 3-terminal, +12V regulator is the Q11 load, with a +14V drop-out specification.

c) Regulator output load current is assumed 0.1A, and needs about 10mA to operate. Total I_d current required = 0.110A.

d) P-Channel FET has R_{ds} = 0.4Ω maximum plus satisfactory I_d, and BV_{dss} specification

Computations:

1) Voltage drop across Q11 is 44mV therefore +14.95V is available at Q11 drain, close to a volt over its minimum requirement.

2) With 0.10V R_s voltage drop chosen, regulator voltage input is +14.85V.

3) Regulator input requires +14V (includes dropout voltage).

4) R_s = (+14.95V) – (+14.85V) / 0.11A = 0.91Ω. R_s, power dissipation: (0.11A)2 x 0.91Ω = 11mW.

5) From the U19 data sheet, a 0.10V sense voltage will output about 1mA from the I_o output. To get a +3.5V logic 1 level across R_o when I_d current flows: R_o = 3.5V / 1mA = 3.5kΩ.

Using R_s = 1Ω, and R_o = 3.3kΩ, +12V regulation measured good with its 0.1A load. Other measured voltages with this circuit configuration: Q11 source = +15V; U19, V+ = +14.96V; U19, V- = +14.85V; V_r = 3.65V (no load)

To a limited degree, R_s and R_o values may be adjusted to vary V_r while minimising sense voltage drop across R_s. Table 1 shows +12V regulator output load current capability with other R_s resistor values, and measured V_r voltage.

For all figures in Table 1 voltage regulation output was good. One, 74L04, inverter logic gate was connected across R_o during test measurements.

To complete the fault logic a *CxAfini fault signal is sent back to the multiplexer U4 (replacing a *Cxfini signal). A level sense circuit described earlier, redrawn from Fig 6, is connected to resistor R_o to show this signal path.

The load placed across R_o must allow V_r to transition between acceptable logic 0 to logic 1 levels. Resistor R_o is a pull-down connecting logic device input source current, I_{IL}, to ground when switch Q11 is off. This voltage drop can be too large to properly operate some logic devices.

For example, a single 74XX input with 1mA I_{IL} source current develops Vr = 3.3V. A 74LXX with 0.1mA develops 0.33V across R_o, a good logic 0 voltage level.

Four, switch/fault circuits can replace non-coaxial relay circuits (like Fig 6). Remove signals *C3fini to *C6fini from the multiplexer U4, and replace with *C3Afini, *C4Afini, *C5Afini, *C6Afini signals respectively.

If fault detection is not necessary, remove U19 and equivalent devices U7C, U11A. Connect Q11 drain directly to the external module's voltage regulator or control input. This will give a higher, switched +12V regulator input voltage, so more regulator load current capability. Ground the appropriate Dx input on the multiplexer chip U4 so the state machine can fully sequence. For lower output voltage regulators, switch (and U19) voltage drops are not critical. For these cases possibly a NPN – PNP transistor switch pair could be used. [5].

11.0

References

[1] TR Time-Delay Generator, Chip Angle, N6CA, The ARRL Handbook for Radio Amateurs, 1997, pp 22.53 – 22.56.

[2] A Simple T/R Sequencer, Zack Lau, W1VT, QEX Magazine, 10/1996, p 15.

[3] A Fast, Simple Transceiver Sequencer, Tom Cefalo Jr, W1EX, QEX Magazine, 9/2005, p 25.

[4] A "Fool Resistant" Sequencer Controller and IF Switch for Microwave Transverters, Paul Wade, N1BWT, QEX Magazine, 5/1996, p14

[5] Sequencer, Bo Hansen, OZ2M, VHF Communications Magazine, 4/2007, pp 239-241.

[6] Data sheet, ISSUE 10, 7/2007, ZETEX Semiconductors, p/n ZXCT1009, www.digichip.com

[7] Other P-Channel MOSFET devices:

Siliconix, SI3441BDV-T1: V_{dss} = 20V, I_d = 2.45A, V_{gs} = 4.5V, R_{ds} = 90 mΩ, 6-TSOP package.

International Rectifier, IRFR9014TRL: V_{dss} = 60V, I_d = 5.1A, V_{gs} = 10V, R_{ds} = 0.5Ω @ 3.1A, TO-252 package.

NDS356AP (lower I_d current): V_{dss} = 30V, I_d = 1.1A, V_{gs} = 4.5v, R_{ds} = 0.3Ω @ 1.1A, SSOT-3 package.

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RST Code and S-Meter revisited

1.0 RST Code

The RST code reports how we receive our correspondent. Reciprocally, it sends us a report how we are received. The code is a three-digit number, with one digit for conveying an assessment of the signal's readability, strength and tone (Table 1).

2.0 Current situation

It is obvious that the use of the RST code has evolved. As I'm not working in CW, but at 10GHz for very weak signals, I am not able to give precise information about the current practice of CW aficionados! Nevertheless, it seems to me that the current report for tone should be T9, not to say T9X, at least up to 24GHz. Above, owing to the enormous frequency multiplication from a 10MHz or even a VHF basic reference, the phase noise is becoming significant. F5CAU wrote to me: "Above 76GHz, it is very difficult to get a T9 note and I'm not ashamed of my poor note at 145GHz". WA1ZMS made a masterly lecture at MUD 2004 about that subject. We can see and hear the sound produced by several oscillators received at 241GHz[1].

The T quote is not used for telephony and it is quite a pity, as some phone signals deserve a bad valuation!

A false use of R is very common. We can

Table 1: RST Code.

R = Readability	T = Tone
1 Unreadable	1 Sixty (fifty) cycle a.c. or less, very rough and broad
2 Barely readable, occasional words distinguishable	2 Very rough a.c. very harsh and broad
3 Readable with considerable difficulty	3 Rough a.c. tone, rectified but not filtered
4 Readable with practically no difficulty	4 Rough note, some trace of filtering
5 Perfectly readable	5 Filtered rectified a.c. but strongly ripple modulated
S = Strength	6 Filtered tone, definite trace of ripple modulation
1 Faint signal, barely perceptible	7 Near pure tone, trace of modulation
2 Very weak	8 Near perfect tone, slight trace of modulation
3 Weak	9 Perfect tone, no trace of ripple or modulation of any kind
4 Fair	The T digit applies to CW. We can add an extra digit like this:
5 Fairly good	C Chirpy (frequency shift when keying)
6 Good	D Drifty signal
7 Moderately strong	K Key clicks
8 Strong	X Stable frequency (cryatal control)
9 Very strong	

Table 2: Standardisation of S-meter readings.
1. One S-unit corresponds to a signal level difference of 6dB. 2. On the bands below 30MHz a meter deviation of S9 corresponds to an available power of -73dBm from a continuous wave signal generator connected to a receiver input terminal. 3. On the bands above 30MHz this available power shall be -93dBm. 4. The metering system shall be based on quasi-peak detection with an attack time of 10 ±2 milliseconds and a decay time constant of at least 500 milliseconds.

hear, mainly during the contests, a report like 59 and, at the same time, an inquiry to repeat the quotation! Either S9 is over quoted, or, if is right, R5 is false because of some QRM, QRN or any perturbation. Obviously, it is easier to quote 59 for everybody! Moreover, some do not like to receive a 56 report… On the microwave bands, we can give 41 and receive the same report without any complex, as signals are often very weak

In some particular events, for example rain-scatter, ATV, digital modes, we need to add an extra quotation. The IARU C5-9 Recommendation, adopted at Lillehammer extends the RST system to the following:

- **a** for signals distorted by auroral propagation
- **s** for signals distorted by rain-scatter propagation
- **m** for signals distorted by multipath propagation

3.0

A more technical approach

The readability, strength and tone reports are in principle subjective and depend of the operator. Therefore, a more technical approach is possible.

3.1 Readability

It can be measured by logatoms. A logatom is a phonetic element, chosen without inherent meaning, for use in telephonometry; consisting of a vowel sound preceded and followed by a consonant sound or a consonant combination sound [2]. Logatoms are used for dyslexia detection and to measure the vocal transmission system efficiency. Using those, no meanings «words» avoid any subjectivity.

That is not used in our field but some audiometric experimentation allowed finding the optimal bandwidth. The 300 to 3000 hertz bandwidth is the best compromise between readability and bandwidth for most languages.

3.2 Strength

The strength report can be objective as a simple measurement of the received signal in voltage or power is possible. Such reports are, in particular at the VHF and higher frequencies, useful for more precise evaluation of propagation, antenna properties and receiver sensitivities. At the 1990 Torremolinos Conference, a standardization of S-units was confirmed [3]. See Table 2, from this recommendation, we can draw the Table 3.

3.3 The S-Meter

A RRSI (Received Signal Strength Indicator) is used in professional receivers [4]. It measures the signal input at the first intermediary stage. After processing, the analogue or digital output is used for various applications, for example to display the actual signal received level or to indicate a too low level to be readable.

3.4 It is our S-Meter!

To get a linear scale in S points, that

Table 3: Values of S-units.

	<30MHz		≥30MHz	
S1	-121dBm	200nV$_{eff}$	-141dBm	20nV$_{eff}$
S2	-115	400	-135	40
S3	-109	792	-129	79.2
S4	-103	1.58µV$_{eff}$	-123	158
S5	-97	3.15	-117	315
S6	-91	6.29	-111	629
S7	-85	12.56	-105	1.256µV$_{eff}$
S8	-79	25.06	-99	2.506
S9	-73	50	-93	5

means dBm, we only need a logarithmical processing of the received signal level. It seems simple to design an S-Meter in accordance to the recommendation. However, even the most costly transceivers are very far from the recommendation. We can read in QST and RadCom tests on transceivers with some S-Meter measurements. Transceivers for HF + 50MHz up to date, with a cost from 1000 to 10,000 euros, have an S-Meter scale that is completely unreliable. The level reference between two S-units is not 6dB and S9 is not 50µV, even at 7MHz! For 144 and 144 + 432MHz equipment, only a few tests has been made and none for the S-Meter.

What is the problem to design an S-Meter in accordance to the recommendation? From S1 to S9, the dynamic range is only 48dB. Up to S9+40dB, we have 88dB. A logarithmic amplifier with a dynamic range up to 90dB is an ideal interface to convert the input signal to a linear in dB scale output (for example AD8307).

We can imagine connecting such an IC at the IF crystal filter output to benefit from the HF stages gain. So, our S-Meter is only measuring the received signal and not the adjacent ones. Unfortunately, those IC have a -80 to -85dBm threshold. Therefore, we need a large preamplifier to increase the S1 level up to that threshold. It is not the case in common receivers because a 60dB gain is not required for its behaviour.

Moreover, in transceivers, the amplification stages are subject to the action of AGC. That means a “flat” amplification and the output signal is not proportional to the input. Therefore, it is not correct to put the logarithmic amplifier after that kind of amplification.

In practice, the universally solution is to use the AGC signal to feed the S-Meter. Actually, that signal is very close to the ideal one, but for level signals lower than the AGC threshold. That is why the S-Meter is not in conformity in dB for S-Units and the lack of actual reading under S3. For high-level signals, we can see compression above S9.

Some old receivers were called “panoramic receivers”, and some modern ones have a spectrum display. They can display all or a part of the received band. We can see the strength of signal, any ripple, drift or splatters. Nevertheless, as for a S-Meter, a precise measurement cannot be obtained from the AGC.

Software Defined Receivers can have a very precise S-Meter and spectrum display. They need an amplification chain linear when it is needed, and logarithmical when it is required, directly or with a software correction.

Professional receivers allow precise measurement of signal levels. They have a linear amplification chain. To avoid any saturation for loud signals, a calibrated attenuator in front or in the chain provides a 10, 20 or 30dB attenuation. To obtain the actual strength, we only have to add this attenuation to the current reading. Processing is used to get a dB scale.

As radio amateurs we can calibrate of our receiver with a signal generator to be able to give a precise measurement. As the sensitivity is not the same on each band a calibration must be done for each band. Moreover, the sensitivity is not flat within a band!

On the HF bands, calibration down to -121dBm is quite easy, at and above 144MHz; it is illusory to apply -141dBm at the receiver input. Generator and cables leakage is strong enough to pass the signal. I cannot imagine such measurement at 10GHz!

Receiver noise factor and S1

We know that S1 corresponds to -141dBm above 30MHz. That figure doesn't mean much so we can try to place it within the theoretical limits of our receivers;

We now that:

P_{noise} = kTB
K = 1.28×10^{-23} J/K
T = 273 + 27 = 300 K at +27 °C
B = (3000 – 300) = 2700Hz for SSB

Therefore, we have in dBm:

P_{noise} = $10 \log 1.38 \times 10^{-23} + 10 \log 2700$

If the receiver noise factor is 1dB, it's sensitivity is:

-174 + 34 +1 = -138.7dBm for a signal/noise ratio = 1

Even with a noise factor nil, the sensitivity is:

-174 + 34 = -139.7dBm

and an S1 signal which is -141dBm remains again in the noise!

Although its not formally mentioned, the above shows that for VHF and above the S-Meter deviation starts at S2 with only the thermal noise for 3kHz bandwidth. That confirms the above calculation.

For CW, with a 500Hz bandwidth, we have:

-174 + 10 log 500 +1 = -146dBm

for an S1 signal, we have a signal/radio of 5dB, which is very comfortable!

4.0 Conclusion

After our revisiting, we have more precise ideas about RST and S-meter readings. In addition, we are comforted in the CW interest for weak signals reception on the microwave bands and for the low bandwidth digital modes for lower bands (and why not above?)

5.0 References:

[1] Millimeter-Wave LO References & Phase Considerations, WA1ZMS, Microwave Update 2004

[2] VHF Managers Handbook, IARU 2006, IARU Region 1, CH-6330 CHAM ZG, Suisse

[3] Vocabulaire Electrotechnique International, IEV 722-01-15

[4] IEEE Glossary Term : RSSI

Ralph Berres, DF6WU

Improving harmonic frequency measurements with the HP8555A

1. Introduction

Sometimes there are small things that make cause more trouble than they should.

Approximately two years ago I wanted to examine the output spectrum of my VHF transceivers and used in ageing HP 141T spectrum analyser with the HP8552B IF module and HP8555A RF module. I was not absolutely content with the measurement results because I thought that my transceiver had poor harmonic suppression.

I searched for the cause of the allegedly poor harmonic suppression of my transceivers for many weeks. According to the display of the spectrum analyser the second harmonic was barley suppressed by 50dB.

Only after I had attached a 500kHz bandwidth cavity resonator filter to the output of the transceivers and the harmonic suppression was no better did I begin to doubt the result of the measurement and looked for the cause in the spectrum analyser.

A study of the HP8555A module data sheet showed that a maximum of –40dBm at the mixer input results in a –65dB internally generated harmonic.

The self-noise of the HP8555A on the 1kHz range is -115dBm. For a reasonable execution speed the 300kHz range must be used on the 2GHz frequency range otherwise the reading is incorrect. The self-noise is then -92dBm. Thus the usable dynamic range of the HP8555A is -92dBm - -40dBm = 52dBm. The dynamic range can be improved by reducing the IF range, however theoretically this requires scan times of 100 seconds or more. Meaningful working (e.g. alignment of a lowpass filter) is therefore no longer possible. The HP8555A data shows this to be correct but I was not very happy with this result.

A detailed examination of the circuits showed that the HP8555A only uses a single diode mixer. Depending on the input signal amplitude the first mixer of the HP8555A can produce many mix products with the first and fourth harmonic of the local oscillator. There is no balanced diode mixer as used on other spectrum analysers.

The local oscillator is coupled to the mixer with a 10db attenuator so the actual mixer diode is only fed with approximately +3dBm oscillator levels. With balanced diode mixers this is usually +13dBm.

The dynamic range of the HP8555A cannot be better because of the technique

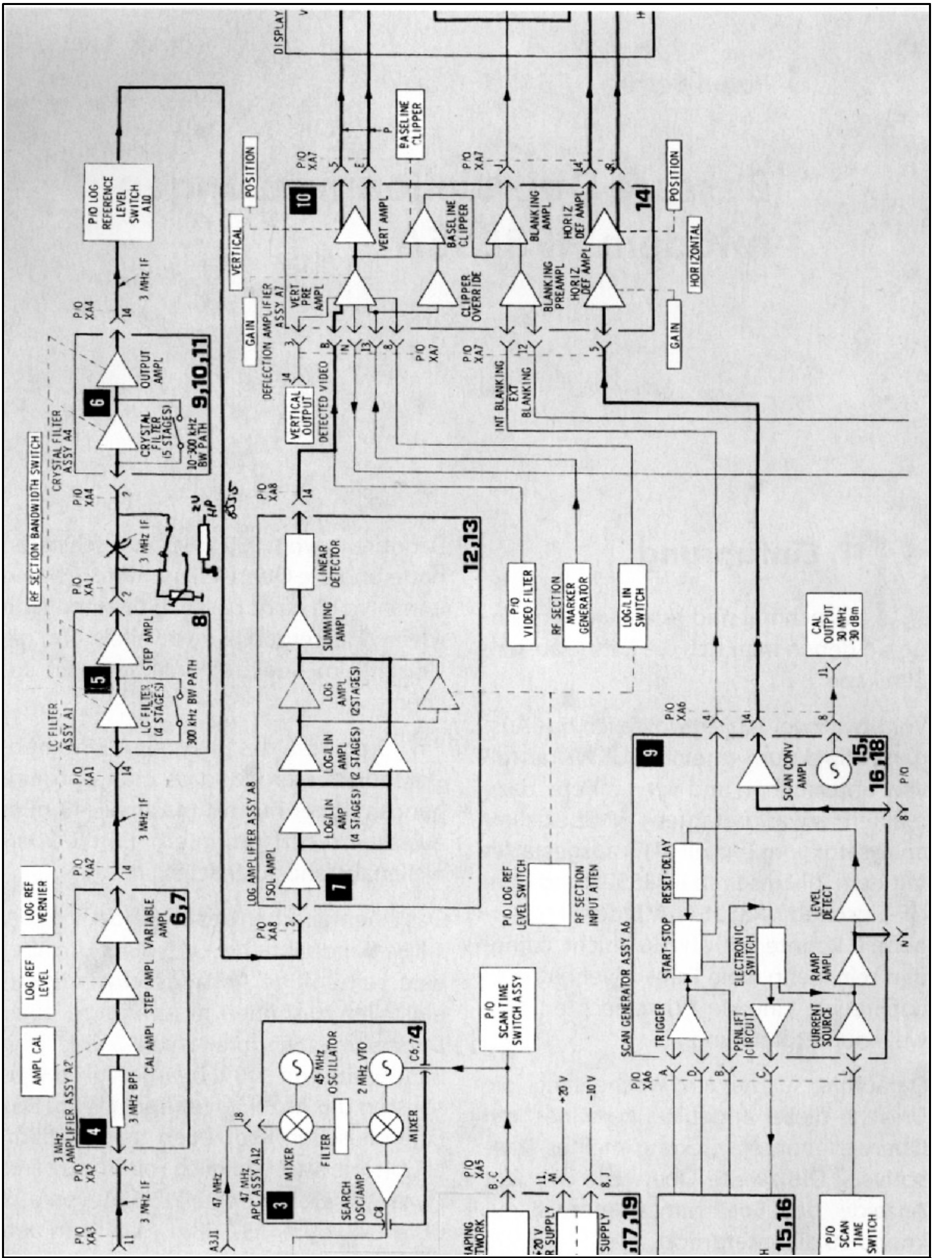

Fig 1: Extract from the block diagram of the HP8552B with handwritten changes.

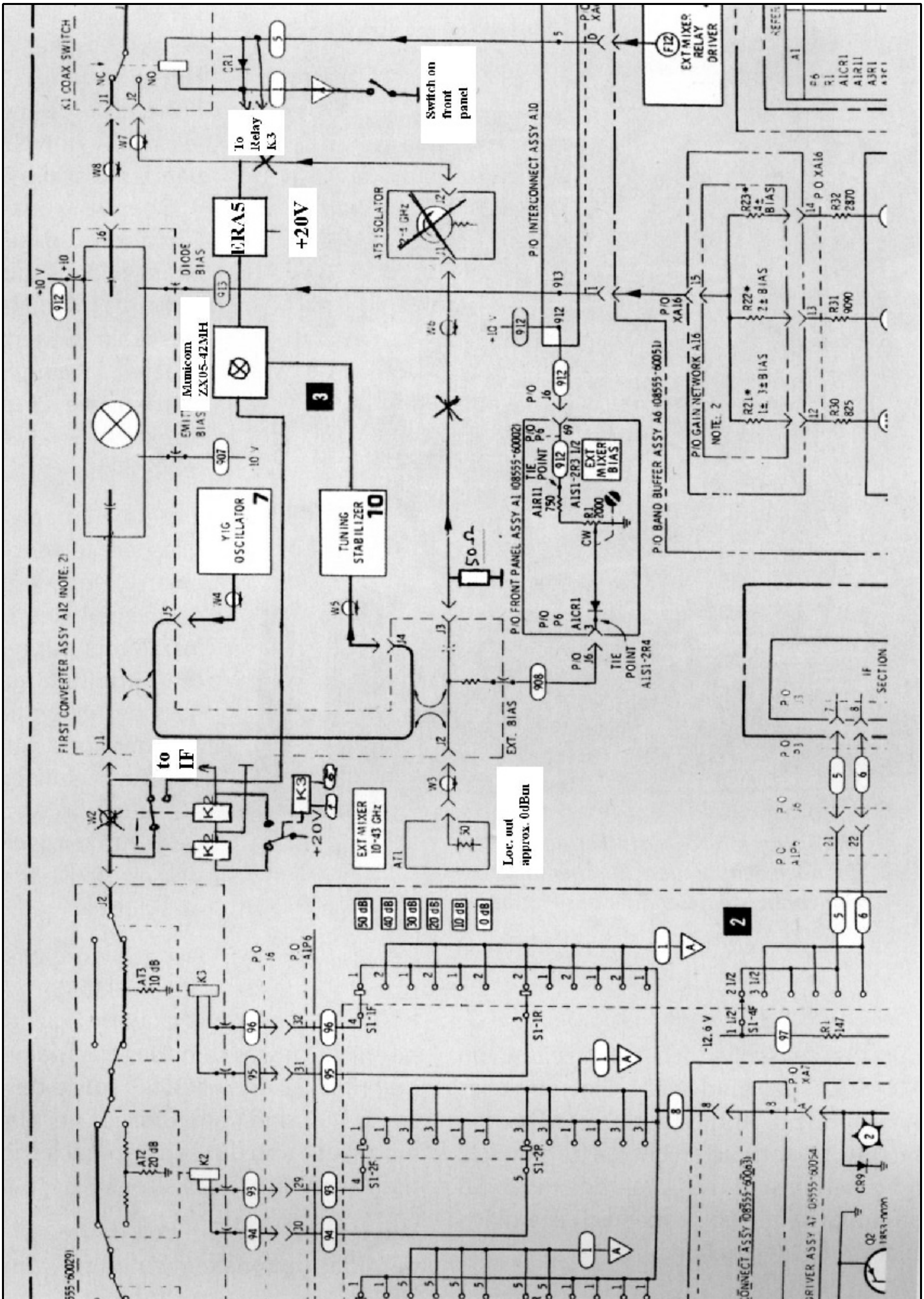

Fig 2: Extract from the block diagram of the HP855A. The changes and additions are shown in bold text.

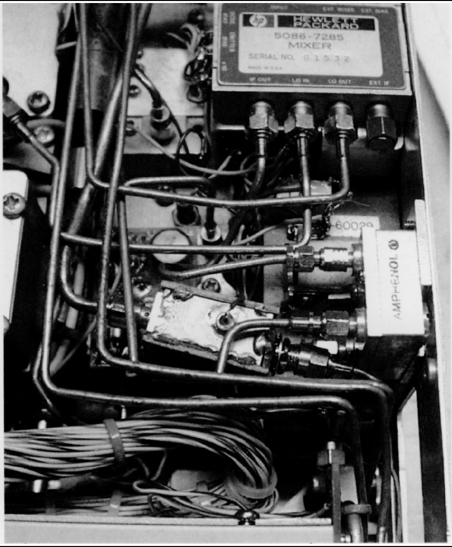

Fig 3: The MMIC amplifier module that follows after the new mixer; built into a tinplate housing.

used. The successors HP8559 and HP8569, as well as the Tektronix 492 were developed using a similar concept and theoretically might be no better.

If the optional HP8444 tracking filters are used this problem does not arise because the tracking filter suppresses the fundamental by about 10dB if the harmonic appears in the IF filter. A tracking filter should always be used because ambiguities in the display are reliably avoided.

2.

The solution

The only solution is using another input mixer, the following text described this in detail.

A separate mixer with a downstream amplifier is used for the primary frequency range up to 1.8GHz. This is operated with a level of +13dBm at the local oscillator input. The following amplifier (Fig 3) raises the signal by approximately 13dB and is a wideband output for the mixer. Without this amplifier the new mixer would only work on the 2.05GHz IF range with a bandwidth of approximately 50MHz due to the following 3 stage cavity resonator filter. It would be just as bad as the original mixer.

The new mixer is a three-way balanced ring mixer from Municom, type ZX05 42MH in a housing with 3 SMA sockets (Fig 4).

The amplifier is an ERA5 MMIC with

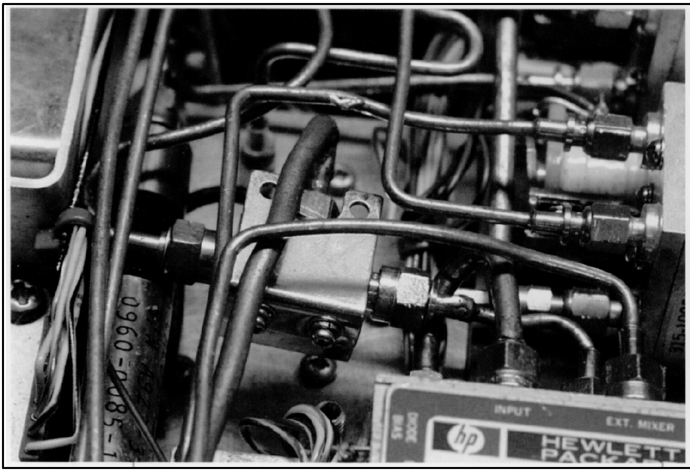

Fig 4: The additional mixer is fitted under the metal angle. The connections: left to the LO; middle IF output; right of RF input.

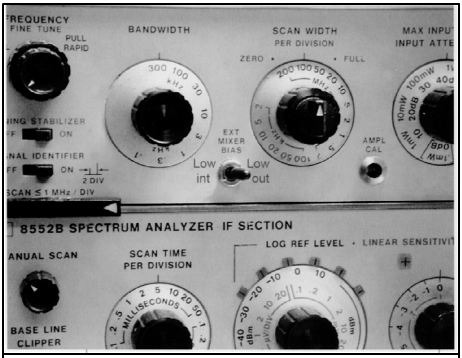

Fig. 5: Arrangement of the new switch on the front panel.

approximately 18dB gain. It is operated with +20V and has a bias resistor of 270Ω. The redundant gain of approximately 10dB could be reduced at this point so that the overall gain is correct. It is better to attenuate the redundant 10dB in the IF amplifier before the crystal filter. This gives better control and 10dB more sensitivity.

This comes with the disadvantage of worse third order intermodulation in the first filter of the IF. Therefore two switchable outputs of the mixer for the primary frequency range are used.

It is to be noted that with the additional mixer the second mixer is fed too highly therefore no modulation and intermodulation measurements are possible.

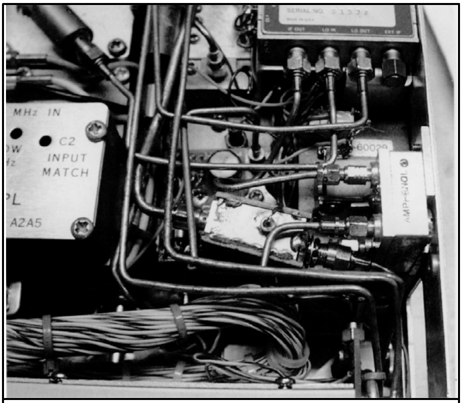

Fig 7: All RF connections are made using semi rigid coax with SMA plugs.

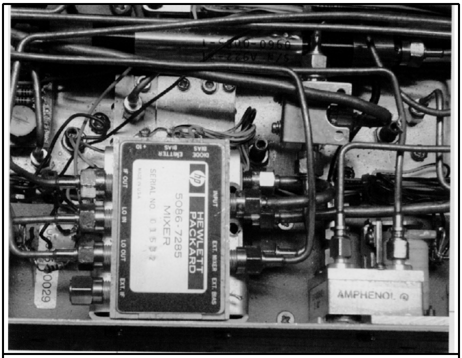

Fig 6: 18GHz relays sit under the original mixer. The SMA terminating resistor can be seen on the left.

However the noise figure is better by approximately 10dB. The original mixer should be used for narrow band measurements. Because of the small IF range, the dynamic range is better, because the noise decreases. The additional mixer should really only be used for harmonic frequency measurements because the intermodulation in the following mixer does not play a role. The change over switch is fitted on the front panel using the hole normally used for the external mixer bias potentiometer (Fig 5). The switch is used when high inputs would be applied to the original mixer. The input for the external mixer is not available following this modification; this socket now carries a 0dBm signal from the local oscillator.

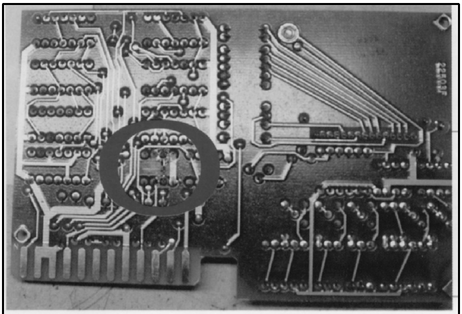

Fig 8: A connection is isolated on the back of the relay PCB.

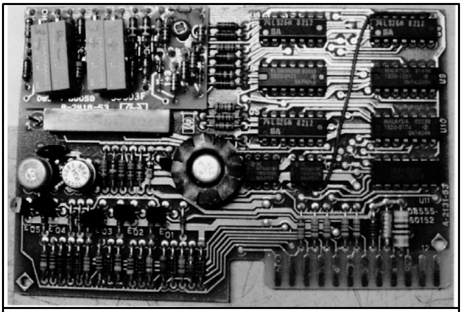

Fig 9: A new connection is made on the front of the relay PCB.

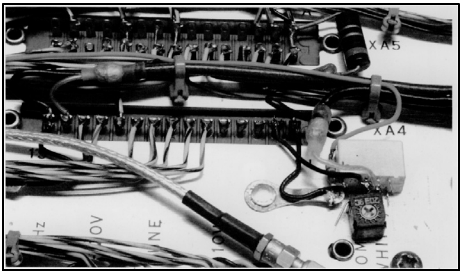

Fig 10: The arrangement of a relay and a potentiometer for adjustment of the level in the IF port is in the right lower half. Also see the block diagram in Fig 1.

3.

The modifications

For the installation of the external mixer the following parts are required:

- 1 Municom ZX05-42MH mixer
- 1 Bistable coax relay for up to 18GHz
- 2 Miniature DPDT DIL 12Volt relay
- 1 Miniature SPDT switch
- 1 trimmer potentiometer 1kΩ.
- 1 SMA terminating resistor

The following aids are required:

- The original service documents for the HP8555A and HP8552
- An extension cable for operating the modules outside the chassis.
- Cable connection between RF and IF module.
- Special key for the SMA plugs this should be very slim and strong.

There are also the plug links for the modules to the basic equipment e.g. RS Electronics. There are Conec plug links to the other modules.

The AT5 isolator under the panel that carries the input mixer is redundant and can be removed. The 18GHz relay is fitted in its place. This connects the input attenuator with the two input mixers (Fig 6).

Relay K1 that was used to switch the intermediate frequency port to the external mixer input now selects the outputs of the two mixers.

The additional relay K3 has its coil in parallel with relay K1. +Pol on relay K1 is no longer connected to the -12.6V return but connected via the switch on the front panel to ground.

The change over switch in relay K3 supplies +20V to the 18GHz coax relay K2 and the relay in the IF port. The 18Ghz relay is a bistable version and has 2 coils that swap themselves over with internal switch contacts after the relay has been operated. This can be seen from the handwritten modifications in Fig 12.

A 50Ω SMA terminating resistor is fitted where the isolator connected to the original mixer. The original input for an external mixer now serves as output for a tracking generator, however it has approximately 13dB lower level (now 0dBm). The original local oscillator output is stabilised by a module on the lower surface of the chassis, the signal from the local oscillator for the additional mixer is taken from here.

The existing semirigid cables can be used for connections between individual RF modules. If these are bent too much they no longer look beautiful and the screen

Continued on page 35

Index of Volume 40 (2008)

Miscellaneous

Oscillators

GPS-LCD, and add on to the GPS disiplined oscillator	Primoz Lemut, S53KS	2008/3	160 - 165
The harmful effects of local oscillator noise	André Jamet, F9HX	2008/3	130 - 139
DDS RF signal generator its frequency range	Matjaz Vidmar. S53MV	2008/4	196 - 208

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2003	Y	Y	Y	Y
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2001	Y	Y	Y	Y
2000	Y	Y	Y	Y
1999	Y	Y	Y	P
1998	Y	Y	Y	Y
1997	Y	Y	P	Y
1996	Y	Y	Y	Y
1995	P	Y	Y	Y
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1978	P	P	P	P
1977	P	P	P	P
1976	P	P	P	P
1975	P	P	P	P
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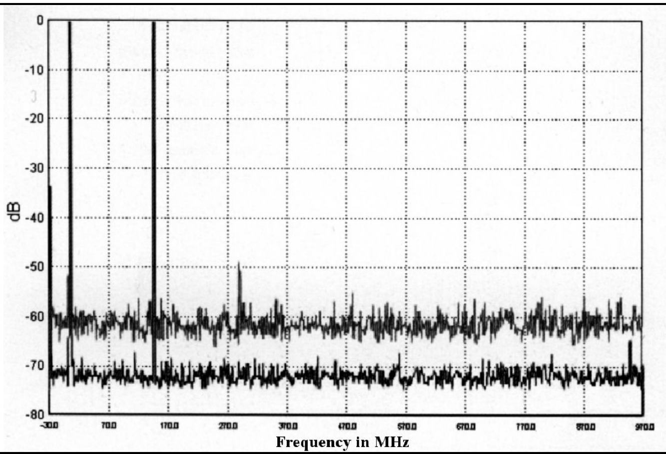

Fig 11: These curves document the difference between the original mixer and the new mixer; the conditions are identical for both curves.

can break open. In addition it can be difficult to fit the parts such as mixers and amplifiers in position (Fig 7). There are semirigid cables that can be used more easily. The relay PCB originally activates relay K1 if the spectrum analyser is switched to an external mixer. It is necessary for the relay driver to supply the relay only if the 10MHz to 1.8GHz range is selected. The tracks on the lower surface of the relay PCB are cut at IC U7 between pin 2 and pin 7 (Fig 8). A new connection is made on the topside of the PCB between pin 2, U7 and pin 8, U8; see Fig 9. In the original diagram the pin allocation of the IC U8 is drawn incorrectly.

The adjuster added in the IF (Fig 10) reduces the additional gain from the new mixer. Feeding a signal into the spectrum analyser and switching between the two mixers while making adjustments until the two give the same display adjust this.

Thus the modification is finished.

4.

Comparative measurement

The conclusion was a comparative measurement with an SMS2 signal source. The harmonics were suppressed with a

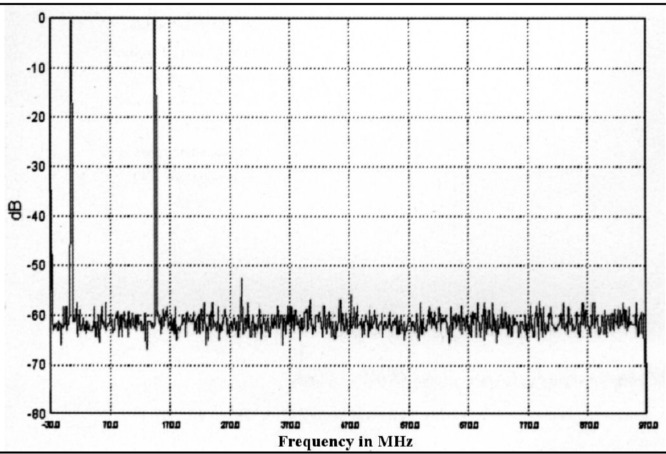

Fig 12: Detailed view of the measurements with the original mixer.

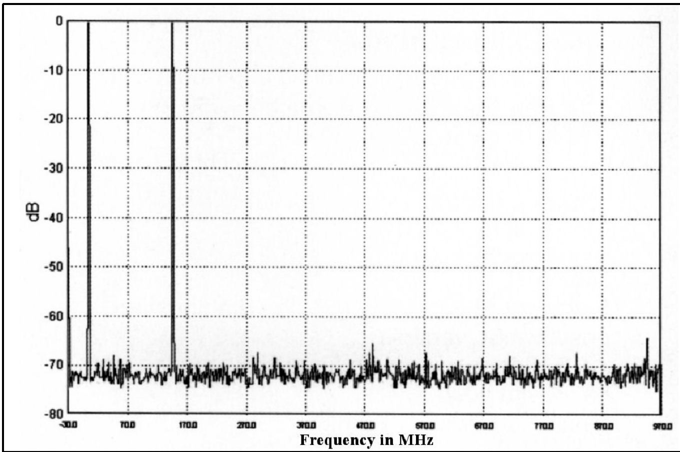

Fig 13: Detailed view of the measurements with the new mixer.

lowpass filter and a cavity resonator filter connected between the signal source and the spectrum analyser. For the measurement accuracy the same settings were used on the spectrum analyser and the same input signal was used, only the mixer was switched. The measured curves are shown in Fig 11 – 13.

5. Literature

The following documents were used:

[1] Service manual HP8555A

[2] Service manual HP8552B

[3] Data sheet Municom ZX05 42-MH

[4] Data sheet ERA5

José Geraldo Chiquito

Antenna Array for the 6cm Band

1. Introduction

When one thinks about antennas for frequencies above a few GHz, the first types that come to mind are probably the parabolic reflector and the horn antennas. Both can be seen as "aperture" antennas, for which gain and directivity are related fundamentally to the ratio between the aperture area and λ^2.

The parabolic reflector antenna is conceptually simple, and can have large gain and high directivity. However, the construction of a high gain parabolic reflector can pose problems to an amateur because of the mechanical precision involved. Additionally, a good feed for a parabolic reflector isn't too easy to implement, especially if it has to be well matched to the dish. Due to losses in the feed, mismatch of the illumination and losses along the edge, the actual efficiency may be no more than 0.6 or so. Last, even if a shallow dish can be considered as an almost 2-D structure, the presence of the feed and its support transforms the parabolic reflector into a 3-D antenna.

The horn antenna, on the other hand, has the advantage that in most cases the construction is based only on plane surfaces. Also, a horn has wide bandwidth and its construction is tolerant to dimensional errors. However, the horn suffers from the same basic drawback as the dish antenna: both are fundamentally 3-D antennas, as the length of a high-gain horn is comparable to its lateral dimensions.

An array of dipoles can be an interesting alternative to both the parabolic reflector and the horn antennas for the 5.7GHz band, especially if the longitudinal dimensional of the antenna must be minimum. Our intention here is to show that it is relatively easy to construct a broadside antenna array for moderately high gains, say, between 15 and 25dBi. A dipole array is a low profile antenna that can be lightweight, cheap and simple to construct, and yet have good performance. The reflector for an antenna array is plane, not curved, so it is much simpler to construct than, for example, a parabolic one. If more gain is desired, it is easy to scale up the linear array into a bidimensional array with larger gain. Although the described antenna array in this article was designed for 5.2GHz, it is relatively easy to adapt the design for the 5.7GHz amateurs' band. We hope this article will encourage other people to experiment with antenna arrays.

An antenna array is basically a collection of elementary antennas working cooperatively to reinforce the transmitted field (or increase the sensitivity, if the it is a

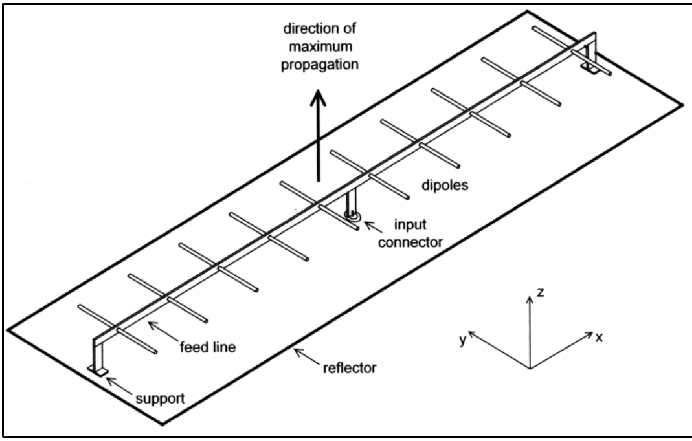

Fig 1: Linear array with 10 dipoles and a reflector.

receiving antenna) at some directions and weaken at others. The elementary antennas are fed by a signal distribution system. The feeding system is very important because it determines the antenna directivity and gain, the direction of maximum radiation or maximum sensitivity, the input impedance, etc. Certain radars, like those used with Patriot missiles, use sophisticated phased arrays antennas. In such antennas, variable phasing circuits in the feed system are used to electronically scan the space under observation. Traditional radars, in contrast, scan the space mechanically, by rotating a parabolic antenna.

In the antenna array design to be described (Fig 1), the goal was maximum simplicity, not sophistication or ultimate performance. The construction was based on materials widely available. The radiators are dipoles with a nominal length of one wavelength. Each radiator is fed with a signal of same amplitude and in phase. That is, the phase is 0º for all dipoles. The array is of “broadside” variety and the direction of maximum propagation is perpendicular to the plane of dipoles. A flat reflector behind the dipoles increases the gain at the same time the backward propagation is virtually eliminated.

The feed line is simply an FR-4 strip. It does the electrical work of sending the signal to the various radiators and is used also as a mechanical support for the dipoles. The reflector is an FR-4 panel with the copper side toward the dipoles. An SMA connector is attached to the center of the FR-4 panel and serves as the input for the antenna. We will suppose it is a transmitting antenna but, because of the reciprocity property, most of the conclusions will also apply when the antenna is receiving.

Although the antenna array appears deceptively simple in principle, several practical and theoretical problems must be solved before we have an operational antenna. Because “God is in the details”, let us briefly examine the theoretical principles of an antenna array.

2.

Theory of antenna array

The Fig 2 shows the linear array, as seen with the axis y normal to the page. The elements of the N-array is represented by the points 0, 1, 2, ... N-2, N-1. It is supposed that the current in each element of the array have same amplitude and phase, so the z is evidently the direction of maximum radiation.

A point P, very far from the array, will

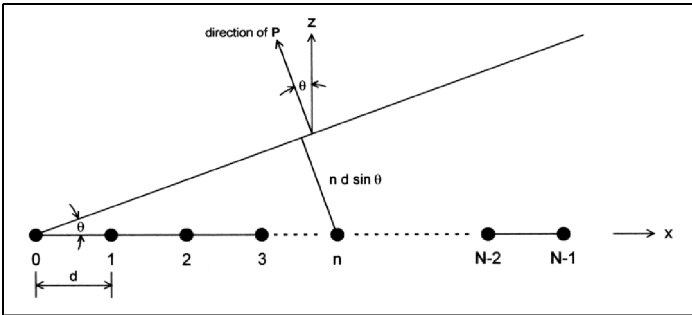

Fig 2: Linear array with N elements.

receive in the general case a sum of waves with different phases (but practically equal amplitudes) since the distance between P and each element of the array will vary. If r is the distance from P to the element 0, the distance from P to the element 1 is (r + dsinθ), the distance from P to the element 2 is (r + 2dsinθ), and so on. The field intensity at P is given by:

$$E = kF(\theta)\sum_{n=0}^{N-1} e^{-j\frac{2\pi nd\sin\theta}{\lambda}} \qquad (1)$$

where:

- k is a complex constant that accounts to the path loss and the transmitting power
- F(θ) is the directivity function of the array elements
- $\sum_{n=0}^{N-1} e^{-j\frac{2\pi nd\sin\theta}{\lambda}}$ is the directivity function of the array. This function is also known as the manifold of the array

Note that θ is given in radians in the formula above. Note also that k is independent of θ if the propagation mean is uniform. he directivity (and the gain) of any array depends on the intrinsic directivity F(θ) of each array element and on the manifold function. The manifold is a purely geometric characteristic of the array. If the array elements are isotropic, that is F(θ) = 1, an array can be very

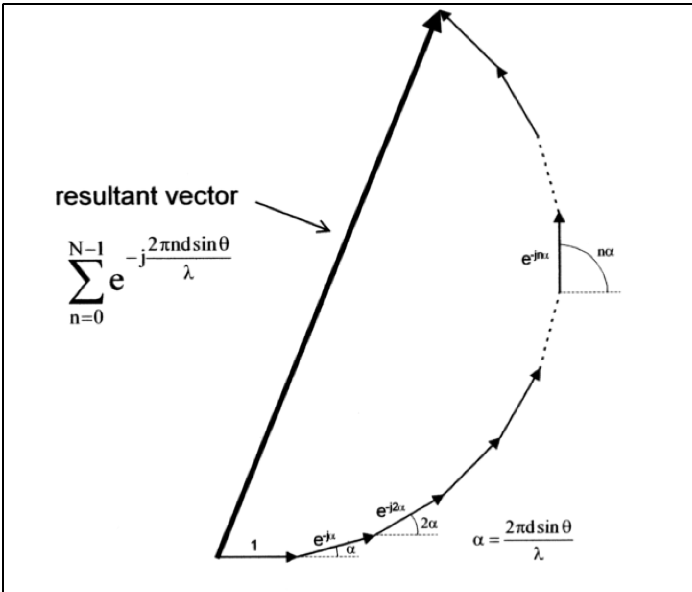

Fig 3: Geometric interpretation of the array manifold.

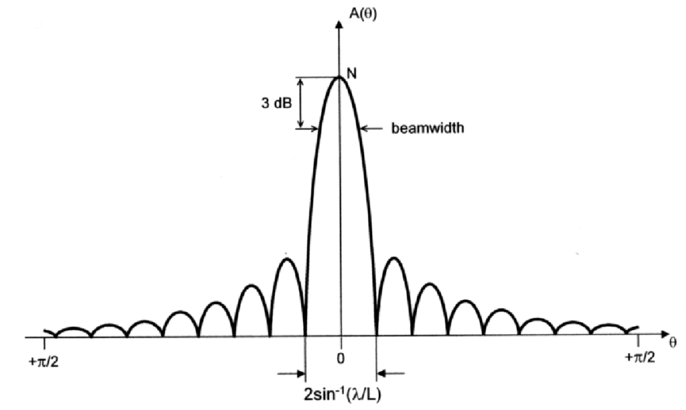

Fig 4: Absolute value of manifold as a function of.

directive because of the array manifold effect.

The behavior of the manifold can be studied from the diagram of Fig 3, which shows that the manifold can be seen as the resultant of a summation of N complex phasors.

The sequence of phasors is a geometric progression so the summation of the N complex phasors can be calculated easily:

$$\sum_{n=0}^{N-1} e^{-j\frac{2\pi nd\sin\theta}{\lambda}} = e^{-j(N-1)\pi d\sin\theta}\,\frac{\sin\left(\frac{N\pi d\sin\theta}{\lambda}\right)}{\sin\left(\frac{\pi d\sin\theta}{\lambda}\right)} \quad (2)$$

Usually we are interested only on the absolute value of the manifold function:

$$A(\theta)=\left|\sum_{n=0}^{N-1} e^{-j\frac{2\pi nd\sin\theta}{\lambda}}\right| = \left|\frac{\sin\left(\frac{N\pi d\sin\theta}{\lambda}\right)}{\sin\left(\frac{\pi d\sin\theta}{\lambda}\right)}\right| \quad (4)$$

2.1 Gain an directivity

A(θ) has a maximum at θ = 0, where A(θ) = N. This isn't a surprise since θ = 0 is the direction for which all the elements of the array send waves with same phase. In the direction of maximum gain, the far field is N times stronger than the field produced by a single element. It is easy to think that the power gain of the array compared to an antenna with a single element would be N^2, since the power is proportional to the square of the field intensity. In fact, the power gain is only N because the input power for the total array is N times the input power for each element.

Fig 4 shows how A(θ) varies with θ. It is enough to represent A(θ) for θ between –π/2 and +π/2 because A(θ) is periodic with a period equal to π. Each time the resultant vector of Fig 3 makes a full turn, A(θ) reaches a null.

The main lobe is centred at θ = 0 and its width, measured as the distance between nulls, is given by:

$$\Delta\theta_{NULLS} = 2\sin^{-1}\left(\frac{\lambda}{Nd}\right) \quad (5)$$

The secondary lobes, also called sidelobes, have half the width of the main lobe. Note that L = Nd is the approximately equal to the physical length of the array, so the main lobe width is given by

$$\Delta\theta_{NULLS} = 2\sin^{-1}\left(\lambda/L\right) \quad (6)$$

The result above shows that the width of the main lobe varies with the inverse of the array length given in wavelengths.

The usual way to evaluate the directivity of an antenna is through the so-called beamwidth, defined as the angular aperture for –3dB. The beamwidth is smaller than the angular aperture given by the first nulls of the radiation pattern. There is no closed formula to calculate the beamwidth, but a good approximation can be found if one notes that in the equation (4) the argument of the sine at denominator is small for θ in the main lobe interval, so the sine can be approximated by its argument:

$$A(\theta) \cong \left| \frac{\sin\left(\frac{N\pi d \sin\theta}{\lambda}\right)}{\frac{\pi d \sin\theta}{\lambda}} \right| = N \left| \frac{\sin\left(\frac{N\pi d \sin\theta}{\lambda}\right)}{\frac{N\pi d \sin\theta}{\lambda}} \right| = N \left| \frac{\sin\left(\frac{\pi L \sin\theta}{\lambda}\right)}{\frac{\pi L \sin\theta}{\lambda}} \right|$$

(7)

The equation (7) shows that A(θ) is a (sin x)/x type function, that has a peak at x = 0 and decays 3dB at x = ±1.391(radians). In terms of beamwidth, this corresponds to:

$$\Delta\theta_{-3dB} = 50.7\left(\lambda/L\right) \quad \text{(degrees)}$$

Table 1 shows how the lobe and the beam width vary for several array lengths:

2.2 Grating lobes

In certain conditions, an antenna array can produce sidelobes with same amplitude of the main lobe. In general, these grating lobes are undesirable in an antenna. A grating lobe occurs if for some θ, dsin(θ) (Fig 2) is equal to a integer multiple of one wavelength. To guarantee that grating lobes will not occur, it is necessary that the distance d between consecutive array elements be less than a wavelength. A separation commonly used between dipoles is λ/2 which is a very convenient value, as we will see when discussing a feed line based on a strip line.

Table 1: Lobe and beamwidth with varying array lengths.

L/λ	lobe width (degrees)	beamwidth (degrees)
2	60	25.4
3	38.9	16.9
5	23.1	10.1
10	11.4	5.1
20	5.6	2.5

2.3 The reflector

The flat reflector placed behind the dipoles has two main effects. One is suppressing the backward radiation; the other is increasing the gain of the array. The gain increases because the energy that is radiated backward is reflected back by the reflector and so it adds to the direct radiation and finally goes to the “right” hemisphere. It is known that an electromagnetic wave suffers a phase reversion after a reflection on a conductor, so it is intuitive that a good position to the reflector is when its distance to the dipoles is equal to λ/4. In this way, the backward wave travels a distance equal to 2 x λ/4 = λ/2 before joining the direct wave, as shown in Fig 5. The λ/2 total path for the reflection introduces a phase variation of 180° but as a metallic reflector introduces a further 180° phase variation, the reflected and the direct radiate in phase. Nonetheless, it can be a little surprise that the gain increases slightly if the distance between the dipole and the reflector is made shorter than λ/4. On the other hand, the antenna bandwidth decreases and the sensitivity for losses increases for shorter distances, so λ/8 is the minimum recommended separation between the dipole and the reflector.

As an observation, the flat reflector is replaced by an electromagnetic image of the dipole in the mathematical analysis called “method of images”. The current in the image dipole is inverted relative to the real dipole, so the tangential electric field and the normal magnetic field are both zero on the surface of the reflector, which is usually considered as a perfect conductor.

2.4 Coupling between the dipoles

The coupling between the dipoles of an array can alter the radiation pattern and the input impedance of each dipole.

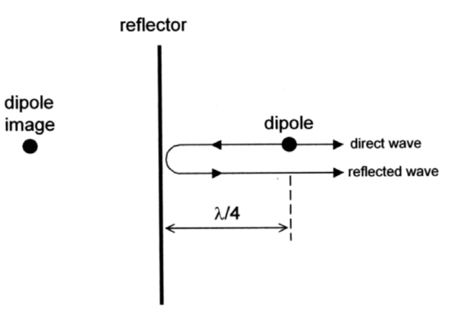

Fig 5: Wave reflection by the reflector.

However, the coupling isn't too strong if the distance between successive dipoles is around λ/2, so it is usual to ignore the coupling except in the case of much sophisticated projects of antenna arrays.

3.

Designing the dipoles

In many types of antennas based on the dipole, the length of the dipole(s) is about a half wavelength. A half-wave dipole is resonant and its input impedance is about 73Ω, a value that is easily matched to a 50Ω transmission line. However, in an antenna array the dipoles are effectively connected in parallel by the feed line, so the input impedance can be inconveniently low, especially if the number of half-wave dipoles in the array is high. So, in large arrays, the full-wave dipoles are preferred over half-wave dipoles because the input impedance of a full-wave dipole is much higher than that of a half-wave dipole. The use of full-wave dipoles eases considerably the matching to the transmission line.

A full-wave dipole doesn't have a well-defined input impedance like the 73Ω of a half-wave dipole. The input impedance of a full-wave dipole can be hundreds or thousands of ohms, depending on the ratio of the diameter d to the length l. Theoretically, as d/l tends to zero, the input impedance tends to infinity. In practical terms, a thin full-wave dipole can have high input impedance (>1000Ω) and a thick dipole has moderately low input impedance (hundreds of Ω). At microwave frequencies, the dipoles are generally thick because l is relatively small. An advantage of the thick dipole is its greater bandwidth, since the input impedance varies less with frequency than that of a thin dipole. It should be noted, too, that a thick full-wave dipole is resonant at a frequency significantly lower than that corresponding to one wavelength. In other words, a "full-wave" dipole for microwave can be considerably shorter than one wavelength.

In the project, the dipoles were made of 2.26mm diameter solid copper wires commonly used in electrical installations. The length that resonates in 5.2GHz was determined by trial-and-error using the EZNEC simulation software. The free-space wavelength for 5.2GHz is 57.7mm but it was found that the 2.26mm diameter full-wave dipole resonate at 5.2GHz when the length is 34mm, just 59% of one wavelength.

The distance from the dipole to the reflector plane has a significant influence on the input impedance, the gain and the directivity of the dipole. Besides, the input impedance of the dipole gets higher as it is moved toward the reflector. By the way, a half-wave dipole has an opposite behavior, that is, the input impedance decreases as the dipole comes close to the reflector. As a practical rule, the distance between the dipole and the reflector should be about a quarter wavelength. In the project, it was used separation of 15mm, which results in an input impedance nearly equal to 300Ω.

Fig 6 shows the radiation patterns for a single dipole at a distance of λ/4 from an infinite conductor plane. The beamwidth for the H-plane (plane xz of Fig 1) is a wide 122°. The radiation pattern is very

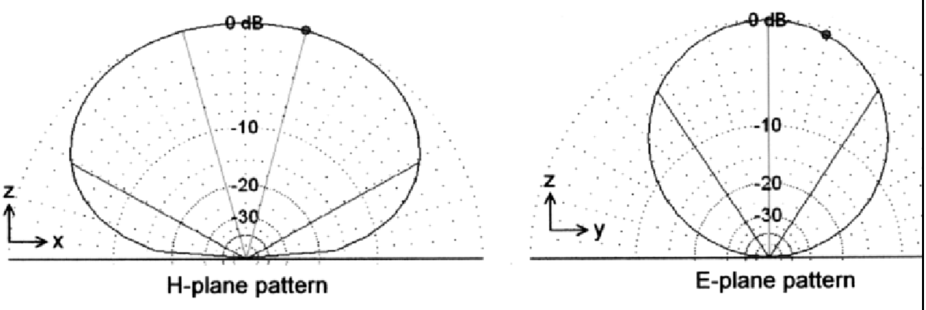

Fig 6: Radiation patterns for a full-wave dipole over a conductor plane.

broad, almost omni-directional. The radiation pattern would be a perfect circle if the reflector didn't exist. Nevertheless, an array with 10 dipoles is very directional in the H-plane because of the array manifold factor, as shown by equation (1).

The beamwidth in E-plane (yz plane of Fig 1) is 67°, which is narrower than the beamwidth in H-plane. However, it should be noted that the E-directivity of the linear array doesn't benefit from the manifold factor since the dipoles are lined up along the x axis.

The 2.26mm diameter 34mm long dipole has a gain of 7.65dBi and an input impedance of 307Ω at 5.2GHz, as predicted by EZNEC. Fig 7 shows how the SWR varies with frequency for a reference impedance equal to 300Ω. The SWR varies slowly and is less than 2 for frequencies between 4.5 and 6GHz. This broadband behaviour is basically the result of the thickness of the dipole, which has d/l equal to 6.6%. In contrast, a dipole for HF bands can have d/l as low as 0.002%.

4.

Designing the feed line

The feed network is probably the most critical component of an antenna array.

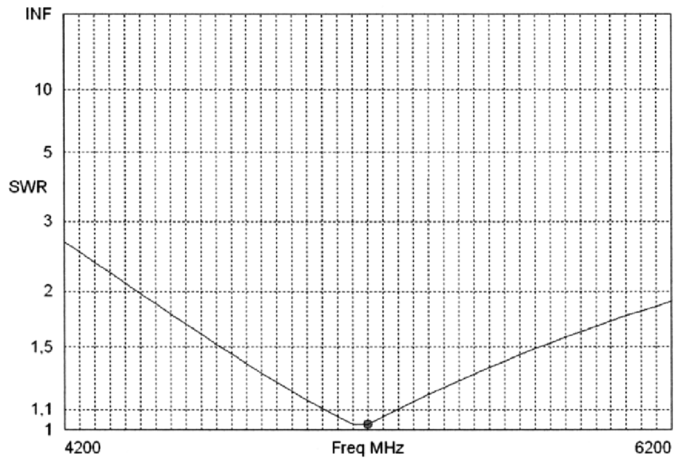

Fig 7: SWR of a single dipole for a reference impedance of 300Ω.

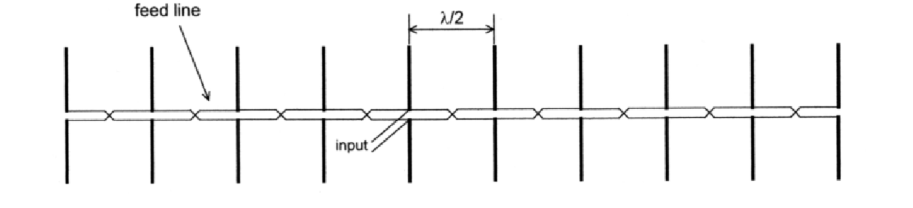

Fig 8: Feed line for an array of dipoles separated by λ/2.

In a broadside array, the feed network must feed all dipoles in phase. The feed network is made of multiple transmission line sections. Each section must have the correct length because in a transmission line the phase is a function of the line length.

In many arrays, the dipoles are separated by half wavelength, as shown in Fig 8. Note the transpositions in the feed line are used to compensate the 180º phase introduced by the propagation in a λ/2 long transmission line sections. Of course, this method of feeding works only if the phase velocity in the feed line is the same as in the free space, for example, when the feed line uses air as dielectric, like the cases where bare wires in "X" are used as feed line.

It is important to note a property of a transmission line of length λ/2 and characteristic impedance Z_0. If the termination load is Z_L, then the input impedance of the line is equal to Z_L, no matter the values of Z_0 or Z_L. This property holds for any length multiple of λ/2, like λ, 3λ/2, 2λ, etc. Even if the feed line SWR isn't equal to one, the phasing of each section will be always 0° or 180°. The input of the array can be the terminals of any dipole but in general it is better to place the input at the center of the array. The input impedance of the array is Z_d/N, where Z_d is the input impedance of each dipole, and N is the number of dipoles.

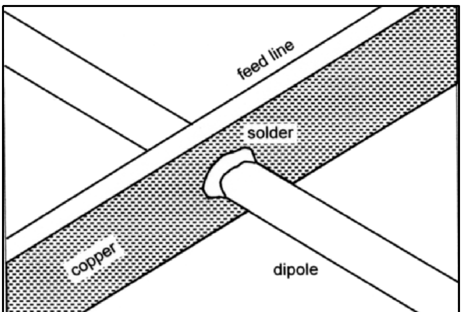

Fig 9: Detail of the dipole soldered to the feed line.

In the project, a strip of double-sided FR-4 board with a thickness of 1.5mm was used as a feed line. The strip also serves as a mechanical support for the dipoles, which are soldered directly to the strip line {see footnote 1} as shown in Fig 9. The strip can be easily cut with a saw from a virgin FR-4 board. The exact width isn't critical and something like 3mm is adequate.

To successfully implement a broadside array, it is necessary that:

- The dipoles work in phase
- The physical distance between them don't depart too much from λ/2

Fortunately, the FR-4 material has the "right" dielectric constant to satisfy both requirements. The (relative) dielectric constant, ε_r, of the FR-4 dielectric is 4.65 but as in the strip line, part of electric field is in the air, ε_{eff}, the effective dielectric constant, is about 3.7. The velocity of propagation in the strip feed line is given by $c/\sqrt{\varepsilon_{eff}}$, where c is the free-space velocity. Using a value of ε_{eff} = 3.7, the calculated velocity in the strip feed line is about 52% of the free-space velocity that means the guided wavelength, λ_g, is equal to 0.52λ where λ is

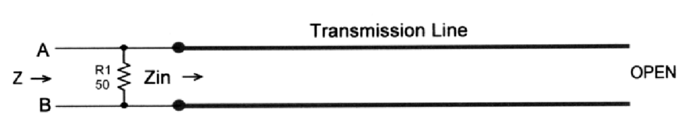

Fig 10: Circuit for measuring the velocity of propagation in a transmission line

the wavelength in free space. Note that λ_g is the length that changes the phase of the guided wave by 360°. So if the physical separation between dipoles is made equal to λ_g, that is almost equal to $\lambda/2$, the requirements for correct feeding are of satisfied, even without using transposition.

4.1 Measuring the velocity of propagation in the FR-4 strip line

The distance between dipoles is probably the most critical dimension of the array. As described before, for a strip line feed based on FR-4 substrate, the distance between dipoles must be exactly one λ_g, the guided wavelength. The method used for determination of λ_g was based on measurement of the return load of an open-loaded transmission line with its input in parallel with a resistance of 50Ω, as shown in Fig 10.

Suppose first that the transmission line is loss-free. The input impedance Zin of an open-loaded transmission line varies from zero to infinity, depending on the frequency. If the transmission line length L is a multiple of $\lambda_g/2$, Zin is infinite, but if L is an odd multiple of $\lambda_g/4$, then Zin is zero. It is easy to see that Z, which is the combination of Zin in parallel with R1, varies from zero to 50Ω, depending on the frequency. The return loss (S11) measured in dB at the point A and B will vary, as a function of frequency, between zero and minus infinity. Theoretically, the RL is minus infinity when the input impedance of the transmission line is infinite and will be zero when the input impedance of the line is zero.

On the other hand, if the transmission

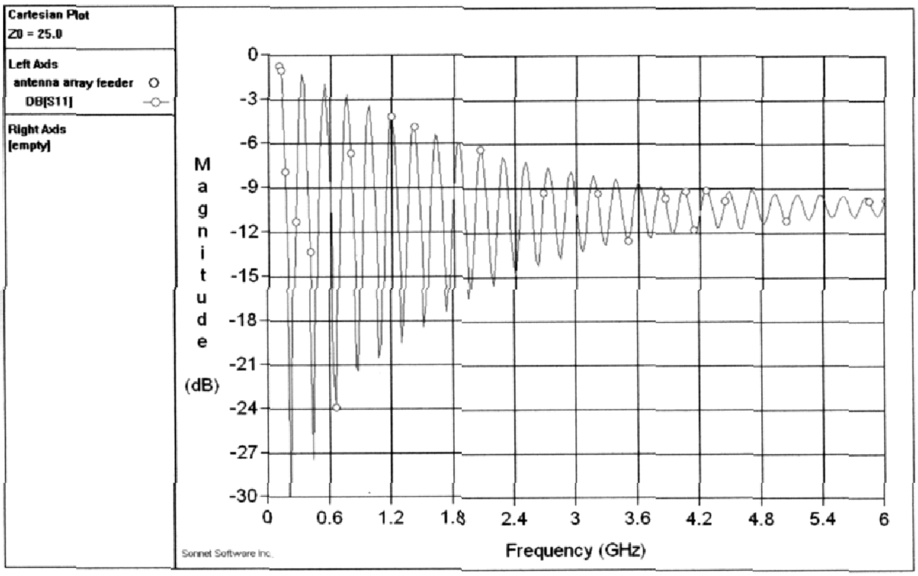

Fig 11: Return loss of a 3mm wide and 355mm long FR-4 strip line.

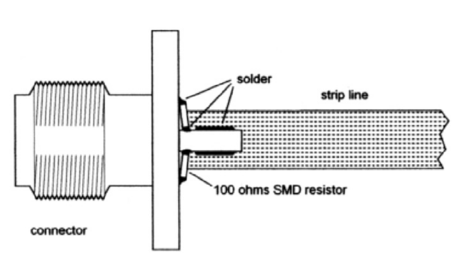

Fig 12: Jig for return loss measurement.

line is lossy, RL will oscillate from something below zero and something above minus infinity, as show in Fig 11 for a FR-4 strip line with width of 3mm and length of 355mm. The graph is from a Sonnet simulation; high-frequency electromagnetic software which can be downloaded free from http://www.sonnetsoftware.com/.

As frequency increases, the line losses also increase but the amplitude of the oscillation of RL decreases. Furthermore, as $f \rightarrow \infty$, $Zin \rightarrow Z_0$ (Z_0 is the characteristic impedance of the transmission line) and RL $\rightarrow$ -9.5dB ($Z \rightarrow Z_0/2$). Note that the positive peaks correspond to odd multiples of $\lambda_g/4$ and the negative peaks to multiples of $\lambda_g/2$.

For measurement of the return loss, two 100Ω SMD resistors were soldered to the back of a female connector (it can be BNC, TNC or SMA). The resistors were soldered directly from the central pin to the outer conductor. This is important to keep the parasitic inductances to a minimum. The end of the strip line was soldered to the central pin and to the ground as shown in Fig 12.

Fig 13 shows the result of the measurement of the return loss. A spectrum analyser with a tracking generator and an RF bridge was used. Note that the marker at 5.2GHz is on the 22nd positive peak. This means that the length of the strip line corresponds to 43 times $\lambda_g/4$ at 5.2GHz. The length of the strip line is 355mm, so λ_g is equal to 355 x 4/43 = 33.0mm, that is also the distance to be used between dipoles. The velocity of propagation calculated from λ_g is 171,700km/s, that is 57.2% of the velocity of propagation in free space.

From the curve of RL simulated by Sonnet (see again Fig 11), it can be calculated that λ_g is equal to 30.2mm and that the velocity of propagation in the FR-4 strip line is equal to 52.4% of the free-space velocity. Comparing the velocity propagation simulated with the

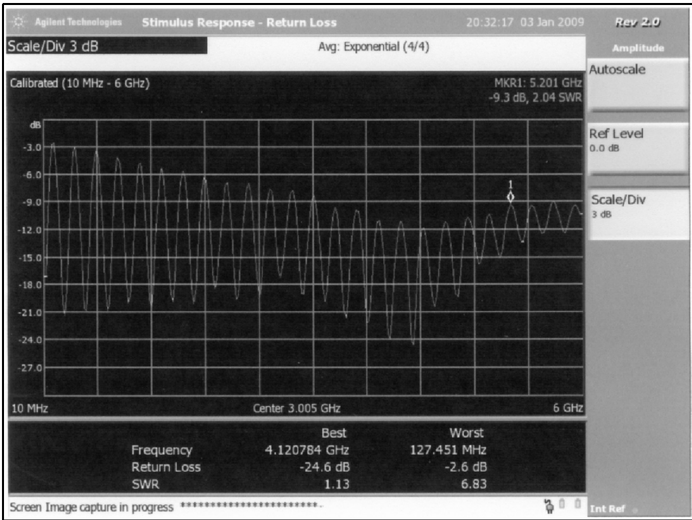

Fig 13: Measured Return Loss of a 3mm wide and 355mm long FR-4 strip line.

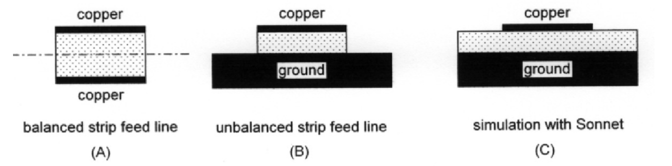

Fig 14: Real and simulation strip line cross views.

measured value there is discrepancy of nearly 10%. The velocity in the real strip line is greater than the simulated value. This discrepancy can be explained for the dielectric layer in the real line is truncated, whereas it is not in the simulation. To understand this point better, see Fig 14 that shows the section views of the strip line, real and simulated.

The real strip line (A) is a balanced structure with a width of 3mm and a thickness of 1.5mm. The characteristic impedance Z_0 is about 50Ω. This balanced line can be analysed from an equivalent unbalanced line (B) with half the dielectric thickness and half the characteristic impedance.

Sadly, Sonnet Lite doesn't allow truncated dielectric layer, so the simulation was done for the continuous dielectric layer (C). This simulated microstrip has a greater fringe capacitance than the real strip line, because in the simulated microstrip, the most part of fringe field lines are in the dielectric, which has greater dielectric constant than the air.

4.2 Feed line losses

Losses in FR-4 increase rapidly with frequencies, so one could be sceptical if FR-4 is appropriate as a substrate for a feed line operating in 5.2GHz. We will show that the losses aren't excessive.

In the prototype a feed line with uniform characteristic impedance was used (Fig 15). The central points A and B are the input for the array. Each line section between consecutive dipoles has the same characteristic impedance Z_0. However, the load impedance for each section varies. For example, the last right line section is loaded by just the dipole 5, but the second last right line section is loaded by the dipole 4 in parallel with the impedance of dipole 5 reflected by the last right line section. As each line section has length equal to λ_g or $\lambda_g/2$, the load at points 1, 2, 3, 4 and 5 are RL/5, RL/4, RL/3, RL/2 and RL, respectively, where RL is the input impedance of a single dipole (which is 300Ω in the project, as seen before). The impedance seen by the generator at input AB is RL/10, or 30Ω.

The losses in a FR-4 strip line, 3mm wide and with length equal to λ_g for 5.2GHz, were simulated by the Sonnet software. The line was connected between port 1 and port 2 of the simulator. The input signal was injected at port 1 and port 2 was terminated with a resistive load with a variable value.

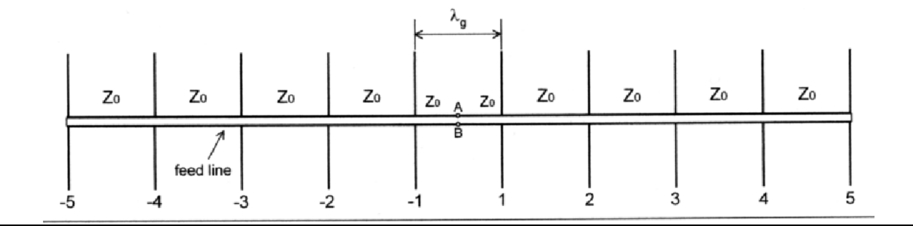

Fig 15: Feed line with uniform impedance.

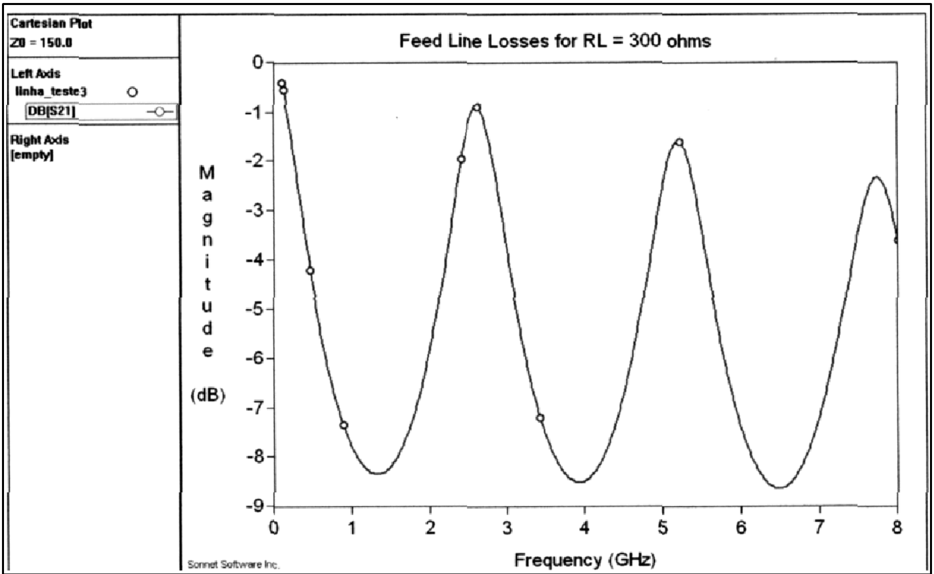

Fig 16: Transmission loss simulated for a 3mm wide, 30mm long, FR-4 strip line with RL equal to 300Ω.

Fig 16 shows the simulated transmission loss (S21) as a function of frequency for a 300Ω load. The minimum loss occurs at the frequencies where the line length is equal to multiples of $\lambda_g/2$, that is, for 2.6, 5.2, and 7.8GHz. For these particular

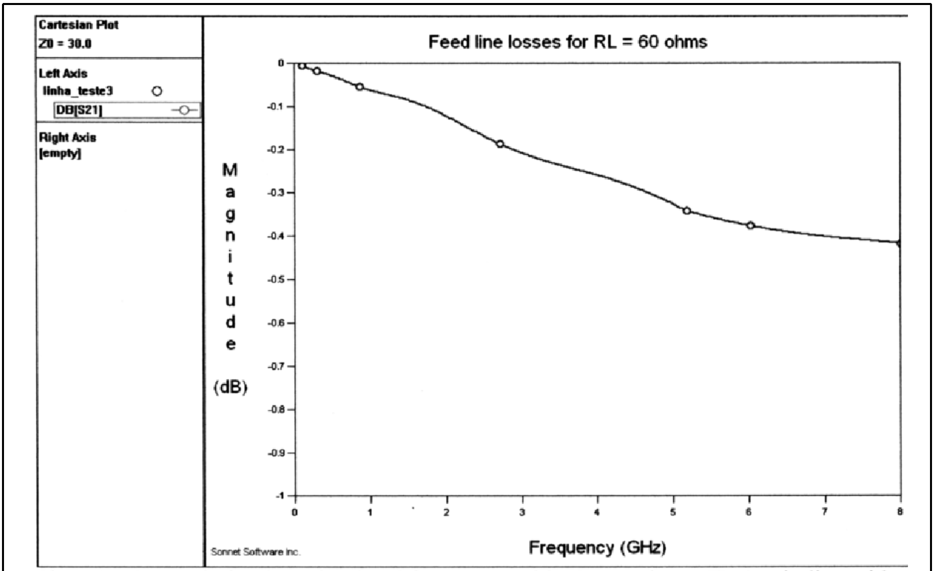

Fig 17: Transmission loss simulated for a 3mm wide, 30mm long, FR-4 strip line with RL equal to 60Ω.

Table 2: Data of relative power received.

R_{LOAD} (Ω)	Segment length	Loss at 5.2GHz (dB)	(%)
300	λ_g	1.57	30.3
150	λ_g	0.93	19.3
100	λ_g	0.74	15.7
75	λ_g	0.68	14.5
60	$\lambda_g/2$	0.34	7.5

frequencies, the input impedance is the same as the termination, as a half-wave line reflects the load impedance to the input. For other frequencies, the transmission loss increases because of the mismatch between the generator and the line input.

Fig 17 shows the transmission loss when RL is made equal to 60Ω. This load is nearly perfectly matched to the line. In this situation, the transmission loss increases gradually with the frequency and the ripple almost disappears.

Table 2 shows the transmission loss for the various section of the feed line, when terminated by 300, 150, 100, 75 and 60Ω. The line section with the greatest loss is the one feeding dipole 5 that is the most mismatched. The half section between the input and dipole 1 correspond to the lowest loss because of the good match and, principally, because it is the shortest one.

With the data from the Table 2 it is possible to calculate the relative power received by each dipole and so to estimate the total loss of the feed system. Fig 18 shows how the power is distributed in the antenna array. For example, the line section L1 between the input and the dipole D1 looses 7.5% of the power but delivers 92.5% to the terminals of D1. One fifth (20%) of the power that reaches the D1 terminals are radiated by dipole D1 and four fifth (80%) is delivered by L2 to the next dipoles, and so on.

The power radiated by each dipole can be calculated by:

- P_{D1} = 0.925 x 0.2 = 0.185 P_{in}
- P_{D2} = 0.925 x 0.8 x 0.855 x 0.25 = 0.158 P_{in}
- P_{D1} = 0.925 x 0.8 x 0.855 x 0.75 x 0.843 x 0.33 = 0.132 P_{in}
- P_{D1} = 0.925 x 0.8 x 0.855 x 0.75 x 0.843 x 0.67 x 0.807 x 0.5 = 0.108 P_{in}
- P_{D1} = 0.925 x 0.8 x 0.855 x 0.75 x 0.843 x 0.67 x 0.807 x 0.5 x 0.693 = 0.075 P_{in}
- P_{TOTAL} = 0.658 P_{in}

The total lost power in the feed line is 0.342 P_{in} that represents a total loss of 1.8dB. As a reference, a parabolic antenna with an efficiency of 50% wastes 3dB of the power, so a loss of 1.8dB in the feed line based on a FR-4 substrate appears to be reasonable.

4.3 Optimising the feed line

The loss in the feed line can be reduced if the characteristic impedance Z_{0n} of the each line section is matched to the resistance seen at each feed point. Referring now to Fig 19 and from the previous discussion, the characteristic impedance of each line section for a matched condition is:

- Z_{05} = 300Ω
- Z_{04} = 150Ω
- Z_{05} = 100Ω
- Z_{05} = 75Ω
- Z_{05} = 60Ω

Table 3 shows that the loss is reduced by a little more than ½ dB for the λ_g long sections. The $\lambda_g/2$ is the same as the non-optimised feed line. And note how narrow some lines are now, especially that for 300Ω.

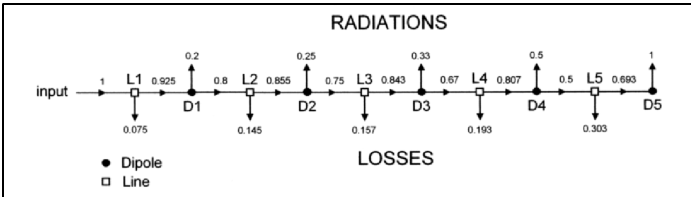

Fig 18: Flux diagram for power distribution.

Table 3: Revised data of relative power received.

R_{LOAD} ()	Line length	Loss at 5.2GHz (dB)	Loss at 5.2GHz (%)	Line width (mm)
300	λ_g	0.60	12.9	0.025
150	λ_g	0.64	13.7	0.75
100	λ_g	0.65	13.9	1.50
75	λ_g	0.66	14.1	2.25
60	$\lambda_g/2$	0.34	7.5	3.00

Calculating again the power radiated by each dipole:

- P_{D1} = 0.925 x 0.2 = 0.185 P_{in}
- P_{D2} = 0.925 x 0.8 x 0.859 x 0.25 = 0.159 P_{in}
- P_{D1} = 0.925 x 0.8 x 0.859 x 0.75 x 0.861 x 0.33 = 0.135 P_{in}
- P_{D1} = 0.925 x 0.8 x 0.859 x 0.75 x 0.861 x 0.67 x 0.863 x 0.5 = 0.119 P_{in}
- P_{D1} = 0.925 x 0.8 x 0.859 x 0.75 x 0.861 x 0.67 x 0.863 x 0.5 x 0.871 = 0.103 P_{in}
- P_{TOTAL} = 0.701 P_{in}

The total lost power in the feed line is 0.299 P_{in}, which represents a 1.5dB loss. This is an improvement of just 0.3dB over the uniform strip feed line. This modest improvement probably doesn't justify the use of a more complicated to construct "optimal" feed line.

4.4 Matching network and balun

The dipoles and the feed line are structures intrinsically balanced but the input connector is unbalanced. Moreover, the impedance between the feed points A-B is 30Ω whereas the impedance of the SMA connector is 50Ω. Therefore, for best performance, an impedance transformer and a balun should be used between the feed points A-B and the input connector. The impedance transformer employed was a simple quarter-wave line section with appropriate characteristic impedance. The balun was based on the coupling effect between a copper strip - with one end soldered to the ground (reflector) and the other left open – and the grounded conductor of the quarter-wave transformer (Fig 20).

The quarter-wave impedance transformer is made from an FR-4 substrate symmetrical microstrip {see footnote 2}. The quarter-wave section is 5.25mm wide and 6.25mm long, with characteristic impedance equal to $Z_0 = \sqrt{50 \times 30} = 38.7\Omega$. The upper end of the quarter-wave section is soldered to the feed line. The lower end is connected through a 3.75mm wide 50Ω section to the SMA connector. The conductor on one side of the 50Ω section is soldered to the pin of the SMA connector and other side is soldered to the ground. Note that the feed point B is connected to the ground through the matching network. However, the path is λ/4 long, so the impedance from B to ground would be infinite if there were no radiation. The quarter-wave stub placed beside the matching network works in a way similar to the "bazooka" balun. The current induced in the stub cancels the effect of the current that flows on the external face of the microstrip conductor from B to ground. The distance d between the stub and the microstrip can be used to tune the input impedance for best matching.

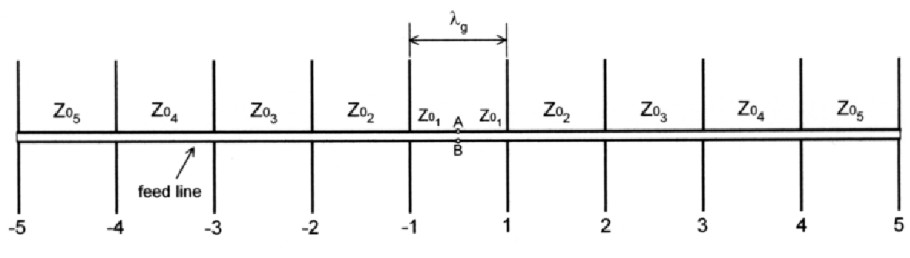

Fig 19: Feed line with impedance optimised at each section.

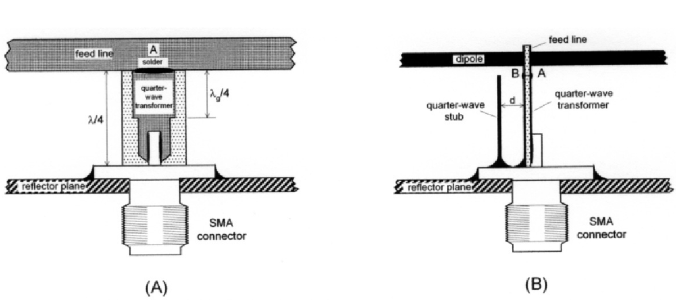

Fig 20: Quarter-wave transformer and balun.

5. Gain estimation

The gain of the antenna array can be estimated by several methods. The EZNEC antenna simulator gives 7.65dBi as the gain of an individual dipole over a conductor plane. As there are 10 such dipoles, the gain of array is 10 times greater, if the coupling between dipoles and the feed line losses are ignored. That would give an array gain equal to 17.65dBi.

Another way to estimate the gain is through the –3dB beam width. A formula that considers that side lobes follow a Chebyshev distribution 15 - 20dB below the main lobe amplitude, gives the gain as:

$$G = \frac{32600}{\theta_1 \times \theta_2}$$ where θ_1 and θ_2 are in °

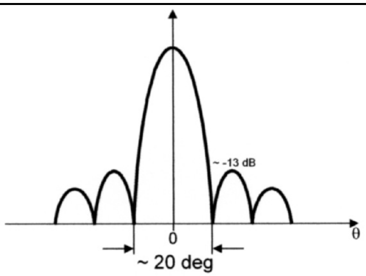

Fig 21: Measured radiation pattern.

According to the EZNEC software, the directivity in E-plane is 67° (given by the dipole radiation pattern), whereas the directivity in H-plane is 10.1 degrees for L/λ = 5 (see Table 1). The gain calculated from the formula above is 16.8dB, that is, about 0.8dB lower than the first estimation.

6. Measurements

The measurement of the radiation pattern requires an anechoic chamber or a reflection-free environment. Sadly, none facility was available, so the pattern measured wasn't very precise. Nonetheless, the first nulls and the amplitude of the first sidelobes were reasonably close to what was expected (Fig 21).

Fig 22 shows the return loss measured at the SMA connector. The bandwidth for RL equal to 20 dB (1.22 SWR) is nearly 280 MHz, but RL is better than 10 dB (1.9 SWR) for almost all frequencies between 4 and 6 GHz.

7. Bidimensional array

If larger gain or more directivity in the yz plane is required, the linear array can be

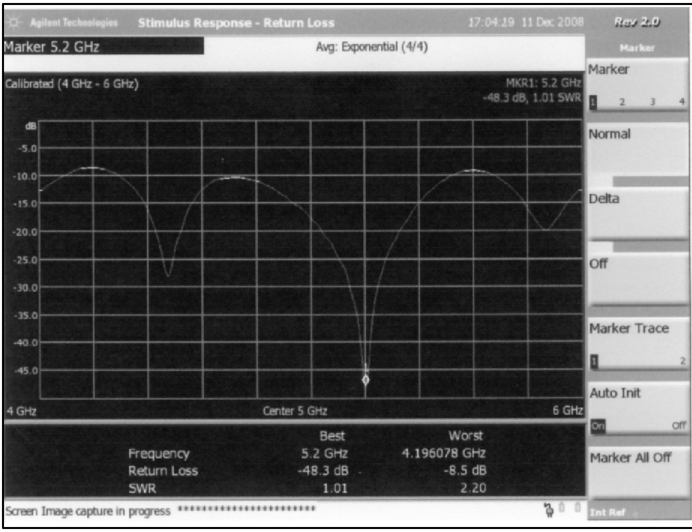

Fig 22: Return loss measured.

expanded into a two-dimension array as shown in Fig 24. For instance, if the linear array is replicated 8 times in the y direction, the gain will increase by 9dB and the estimated theoretical gain for the array will grow to 26dBi. A point important to remember is that the bidimensional antenna array will keep the same low profile of the linear array. The linear sub-arrays must be fed in phase by an appropriate feed network, which can be constructed, for example, by using the same idea of the strip line feed for the individual dipoles. Another idea is using coax to feed the sub-arrays. An even larger gain can be achieved if the array is also replicated in the x direction. For example, duplication in the x direction increases the gain by 3dB and the expected theoretical gain for the entire array would be about 29dB.

Of course, as the antenna array is expanded the gain increases at the expense of the simplicity. There is a point when the parabolic reflector becomes more interesting, especially if the goal is only very high gain, without much concern about antenna profile.

I wish to thank to Carl Lodström, KQ6AX & SM6MOM for his help with the production of this article

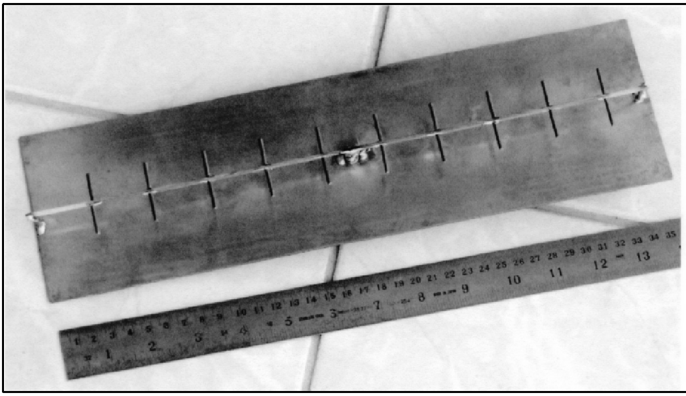

Fig 23: Picture of the prototype antenna array.

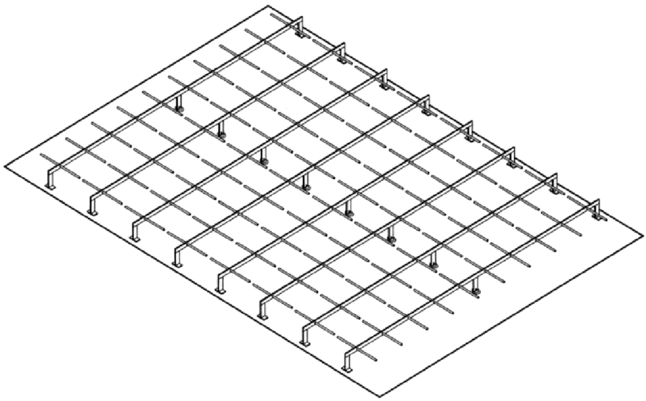

Fig 24: A two-dimension array with 10 x 6 dipoles.

8.

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[5] The ARRL Antenna Book, 20th edition

9.

Footnotes

{1} The double word strip line is used in this article to designate and differentiate the transmission line made of strip cut from a double-clad FR-4 board from the single-word stripline, which has the usual meaning of a transmission line made of flat strip of metal which is sandwiched between two parallel ground planes.

{2} Usually, a microstrip is a transmission line where one of the conductors is "live" and the other is a ground plane. In the quarter-wave impedance transformer used in the project, the conductors on both sides of the substrate are "live". The conductor printed patterns are the same on both sides

John Fielding, ZS5JF

John's Mechanical Gem No. 5 Fitting RF connectors onto panels

Traditional RF connectors such as the N type, BNC and UHF connectors are available in a variety of mounting styles. Some are known as "single-hole mounting" and simply require a hole of the correct diameter to be drilled in the panel and the connector fitted. Others are known as "4-hole mounting" and require the central hole and 4 smaller holes for the screws that attach the connector to the panel. Lining up the holes in the correct places can be difficult and if they are not correct the connector is mis-aligned, making the finished job look poor.

So how do we make sure the connector holes are all in the correct place?

One method is to mark out the panel with the correct pitch between the 4 mounting holes and the central hole. But this requires a great deal of skill and accurate measuring instruments. If we look at the details of the mounting holes for the N and UHF connector we see the problem. The UHF connector is more commonly known today as SO-239.

Examining Fig 1 shows the dimensions are very precise, to 0.25mm on the screw pitch centres. Also the outer dimension of the connector flange is another oddball dimension. But this is historical because all the common connectors were designed using imperial measurements, not metric measurements. 25.4mm as many

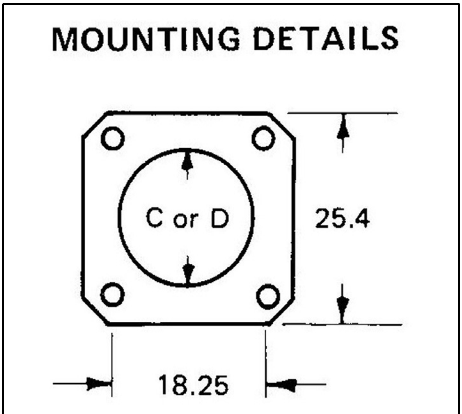

Fig 1: Dimensions for the mounting holes for the N and UHF connector.

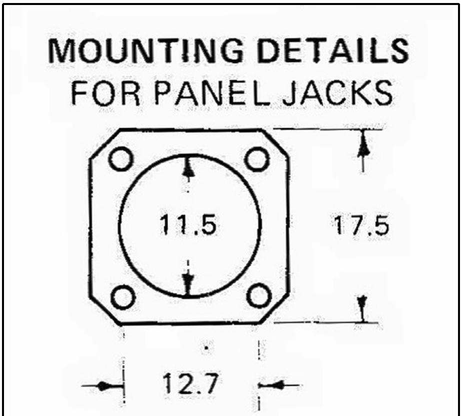

Fig 2: BNC 4-hole connector details.

Fig 3: 16mm flat wood bit.

older people will know is exactly 1 inch. Similarly 18.25mm is exactly 23/32 inch. The Americans who designed these connectors had an obsession with using 16ths, 32nds and 64ths of an inch. The centre hole is 16mm to provide clearance to the 5/8th inch diameter body. Now we look at the BNC 4-hole connector.

Again the dimensions seem to be unnecessarily difficult. Instead of 12.7mm why not make it 12.5mm? It is because the 12.7mm is derived from ½ inch, 12.7 being half of 25.4mm. The dimension 17.5mm is 0.6875 inch, which is 11/16 inch originally. Similarly 11.5mm was originally 0.452 inch which is 29/64th inch and provides clearance for the body which is nominally 7/16th inch.

The best method is to use a drilling template to ensure the holes are in the correct places. Whereas one could make such a drilling template we have an easier method. Taking a scrap connector we can modify it so it can be clamped

Fig 4: Chassis punches.

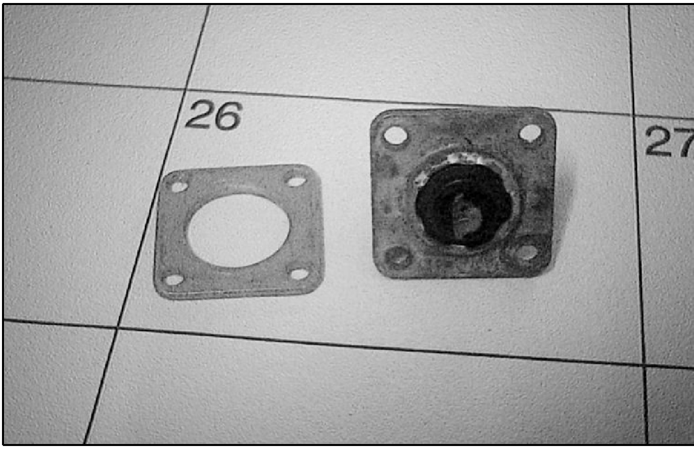

Fig 5: SO-239 connector and drilling. template

onto the panel in the required position and then use the holes to spot through with a drill bit to leave witness mark where the holes need to be drilled.

The best method is to firstly drill the centre hole for the connector body. For the N and UHF types this is a hole of 16mm. The body of the connector is 5/8th inch (15.875mm) and the hole provides the required clearance. Drilling large diameter holes in thin sheet metal often results in a triangular hole and not a round one. In the older days chassis punches could be bought which made the correct diameter hole after a pilot hole had been drilled. Chassis punches today are expensive and the writer prefers an alternative method using flat spade bits used for drilling holes in wood. These will cut a perfectly round hole requiring little deburring in glass-fibre board, aluminium sheet or diecast boxes. The drill needs a small pilot hole and should be run slowly with plenty of cutting lubricant.

When the large hole has been drilled the connector is pushed through the hole from the back of the panel and held in place with a clamp so that the top face or side face are correctly lined up with the edge of the panel. A 3mm drill is then used to mark the start of each hole. The connector is removed and the holes drilled and deburred.

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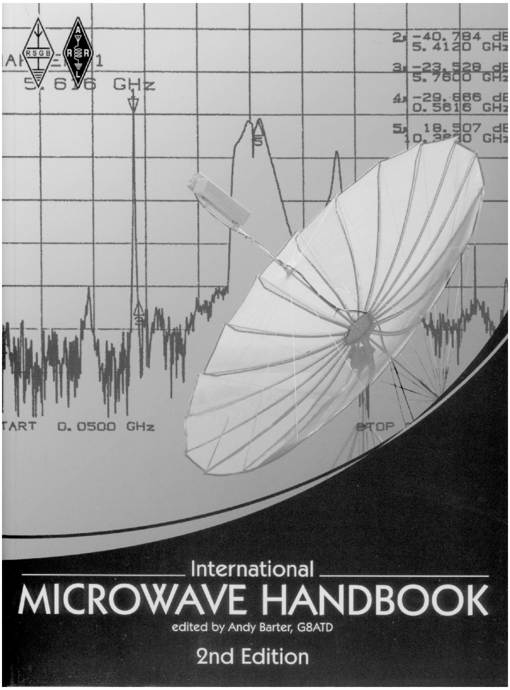
The International Microwave Handbook 2nd Edition



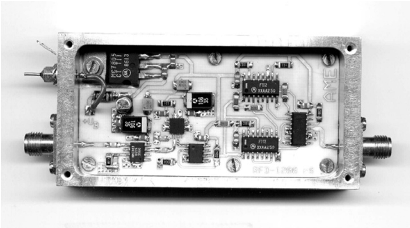
The microwave bands are an excellent area for radio amateurs who want to experiment and construct their own equipment. The RSGB in partnership with the ARRL has produced this invaluable source of reference information for those interested in this area, along with excellent designs from around the world to fire the imagination. Material has been drawn from many sources including the RSGB journal RadCom and the ARRL publications QST & QEX. Alongside this material a truly international range of sources have been used including items from Germany, Denmark, New Zealand, Slovenian and many more.

The earlier chapters of the book provide invaluable reference material required by all interested in this exciting area of experimentation. Techniques and devices are covered in depth, leading the reader to understand better the wide range of equipment and techniques now available to the microwave experimenter. This book contains a wide selection of designs using the latest technology that can reasonably be used by radio amateurs and ranges from ones that can be reproduced by most radio amateurs to those that require a high degree of skill to make.

With the explosion in consumer electronics using microwave frequencies the opportunity to experiment has never been greater and this book is simply the.

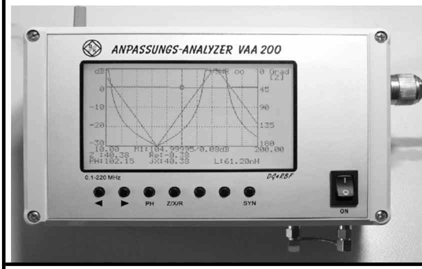
The handbook is available from The RSGB or ARRL

Ready built units, Kits and PCBs available for recent projects



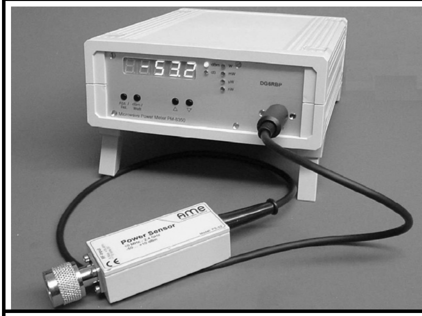
12GHz Divide by 1000 prescaler by Alexander Meier, DG6RBP from issue 4/2003

Ready built in diecast housing with SMA connectors	**£215.00**
Also availabe an 18GHz version of the divide by 1000 prescaler	**£330.00**



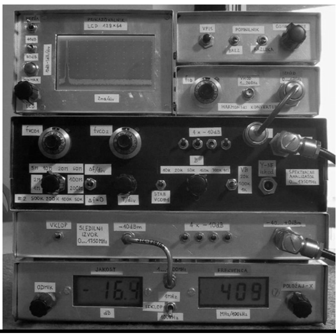
Vector analyser VAA 200 for the 0.1 to 220MHz range with graphical display by Bern Kaa, DG4RBF from issue 3/2007

PCBs	**£31.50**
Micro-controller with software	**£95.00**



Display unit for the power detector by DJ8ES, Alexander Meier by DG6RBP from issue 1/2007

PCB	**£25.00**
Micro-controller with software	**£65.00**



Spectrum Analyser by Matjaz Vidmar, S53MV

This project was first published in issue 4/1998 to 4/1999 (now included on CD1, see back cover) The LCD oscilloscope is in this issue.

Complete set of PCBs for all modules including the LCD oscilloscope **£70.00**

To order use either the web site www.vhfcomm.co.uk or contact
K M Publications: Tel +44 (0)1582 581051, email andy@vhfcomm.co.uk

Gunthard Kraus, DG 8 GB

Internet Treasure Trove

Vivaldi Antenna for Snow radar

A really interesting project and a good article with lots of basic information for this type of antenna its characteristics and applications. Not only for antenna specialists.

Address:

http://www.ittc.ku.edu/research/thesis/documents/ravi_prakash_rajaraman_thesis.pdf

Digitally signal processing Tutorial

This tutorial is for anyone who wants to read detailed information about modern technology. This is a pleasant read for an introduction to this subject. There is also a Java applet with information on topics such as FFT principles and designing digital filters or DTMF generators.

Address:

http://www.dsptutor.freeuk.com/

Tonne software

This is for lovers of modern software used in our hobby. Everything is here from antenna design to filter design, PLL calculation and some Application Notes.

Address:

http://tonnesoftware.com/

EE Times Europe

You should visit this homepage regularly. Again and again there are interesting articles, particularly on the forefront of technical developments. An example is the article: “Using high-spees A/D converters beyond baseband frequencies”.

Address:

http://www.eetimes.eu/201804147

TechOnline

An interesting technical site with clearly divided topic areas and many useful things. There should be no fear of the registration with your own personal data;

this gives access to a large amount of interesting documents.

Address:

http://www.techonline.com/

24GHz

For those for who become bored with conventional RF technology or become over courageous or who simply look for new electronic challenges, why not try the 24GHz microwave band. This homepage is the best way because there is concentrated information on the topic.

Address:

http://www.kielnet.net/home/j.schaefer/atv_24ghz.htm

Texas Instruments

For the big players in the electronics industry communications is important. The frequency ranges are getting higher with ever more interesting and low priced components to stimulate your fantasies.

This applies to the “RF/IF and ZigBee Solutions” from this company. The page addressed below will stimulate your interest. What could you use it for? A further search of the homepage shows possible uses in the amateur radio range.

Address:

http://www.ti.com/corp/docs/landing/cc2591/index-google.htm?CMP=KNCreprisegooglehpa&HBX_PK=HPA_2-4GHZ+2.4GHz&HBX_OU=50

Matjaz Vidmar, S53MV

Readers of VHF Communications Magazine should know this name very well because of the developments and projects that have been published.

There seems to be no end to the new projects that come from Matjaz, probably something to do with his job as a university professor. Take the time to look at his latest activities.

Address:

http://lea.hamradio.si/~s53mv/zifssb/kband.html

Aerocomm Antenna Tutorial

Another antenna topic, but this time a general tutorial. This is not only for the specialists but for revision and to test your own knowledge or for filling the gaps in your knowledge.

Address:

http://www.aerocomm.com/docs/Antenna_Tutorial.pdf

Allocation of the radio spectrum

It is interesting: to see who uses certain frequencies or what frequencies are approved for which services. There are some unscrupulous people who make a lot of trouble.

Address:

http://www.jneuhaus.com/fccindex/spectrum.html

144MHz Beacon list

Not of interest to everyone but if you needs it:

Address:

http://www.dk3xe.com/b_144.htm

This is only a small part of the interests that this amateur shows the homepage. Please take a further look:.

Address:

http://www.dk3xe.com/index3.htm

NASA Tech briefs

A mine of information, white papers etc.. Partially accessible after free registration. The titles whet the appitite.

http://www.techbriefs.com/content/view/3267/34/

Logbook, contest and award system

The Chris Townsend, G8PUT Logbook Project started in late summer 2007, the original idea was to create a simple and easy to use logbook system. Since then he has received many requests to include additional features. The most significant are perhaps Contests and Awards. Work is progressing well on including these and many other features.

The way the project is developing is that new functions and features which have been requested are being added in no particular order. A feature which takes little time to include might be added on the day it is requested. Other features which take a significant amount of work will be made available when they are completed.

Address:
http://www.g8put.com/

eBay shop: HamPlace

I came across this eBay shop while trawling for some parts for a new project. The shop has some interesting items at very good prices. Some eBay fans may not want to purchase from Hong Kong but I have had good experiences and in fact the items that I purchase from this shop arrived in 4 working days, much quicker that some major suppliers can manage. – Andy

Address:
www.ebay.co.uk - then select shops and enter the shop name

Note: Owing to the fact that Internet content changes very fast, it is not always possible to list the most recent developments. We therefore apologise for any inconvenience if Internet addresses are no longer accessible or have recently been altered by the operators in question.

We wish to point out that neither the compiler nor the publisher has any liability for the correctness of any details listed or for the contents of the sites referred to!

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Tel: (909) 338-6887, email: wa6svt@atvquarterly.com

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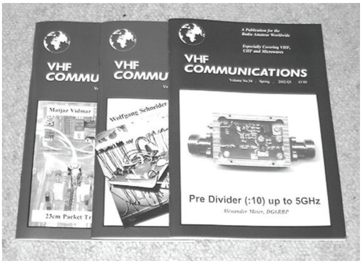
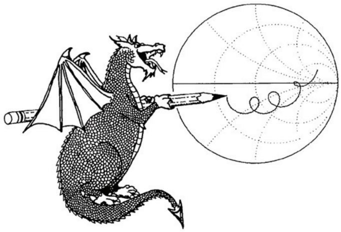
Back Issues



Available either as photocopies or actual magazines. Issues from 1/1969 to 4/2007 are £1.00 each + postage. Issues from 2008 are £4.70 each or £18.60 for all 4 issues + postage. See web site or page 34 of issue 1/2009 for back issue list. There are two back issue sets that contain the available "real" magazines at a reduced price, see web site for details. There are 3 DVDs containing a decade of back issues as pdf files of scanned magazine pages. 1970s (including 1969), 1980s and 1990s at £20.00 each which includes UK postage or surface mail overseas. Air mail postage is £0.60 for Europe and £1.00 outside Europe.

Blue Binders

These binders hold 12 issues (3 years) and keep your library of VHF Communications neat and tidy. You will be able to find the issue that you want easily. Binders are £6.50 each + postage. (UK £1.55, Surface mail £1.85, Air mail to Europe £2.40, Air mail outside Europe £3.60)



PUFF Version 2.1 Microwave CAD Software

This software is used by many authors of articles in VHF Communications. It is supplied on 3.5 inch floppy disc or CD with a full English handbook. PUFF is £20.00 + postage. (UK £0.70, Surface mail £1.45, Air mail to Europe £1.60, Air mail outside Europe £2.60)

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Two CDs containing compilations of VHF Communications magazine articles are available. CD-1 contains 21 articles on measuring techniques published over the last 8 years. CD-2 contains 32 articles on transmitters, receivers, amplifiers and ancillaries published over the last 5 years. The articles are in pdf format.

Each CD is £10.00 which includes 2nd class postage in The UK and surface mail overseas. Air mail postage is £0.60 for Europe and £1.00 outside Europe.

VHF Communications Web Site www.vhfcomm.co.uk

Visit the web site for more information on previous articles. There is a full index from 1969 to the present issue, it can be searched on line or downloaded to your own PC to search at your leisure. If you want to purchase back issues, kits or PUFF there is a secure order form or full details of how to contact us. The web site also contains a very useful list of site links, and downloads of some previous articles and supporting information.

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