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Wide-Frequency Range, Dynamic Matching Network and Power System for the “Shoelace” RF Antenna on the Alcator C-Mod Tokamak

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Abstract—A wide-frequency range (50-300 kHz) power system has been implemented for use with a new RF antenna – the “Shoelace” antenna – built to drive coherent plasma fluctuations in the edge of the Alcator C-Mod tokamak. A custom, dynamically-tunable matching network allows two commercial 1 kW, 50-Ω RF amplifiers to drive the low-impedance, inductive load presented by the antenna. This is accomplished by a discretely-variable L-match network, with 81 independently-selected steps available for each of the series and parallel legs of the matching configuration. A compact programmable logic device (CPLD) provides a control system that measures the frequency with better than 1 kHz accuracy and transitions to the correct tuning state in less than 1 ms. At least 85% of source power is dissipated in the antenna across the operational frequency range, with a minimum frequency slew rate of 1 MHz/s; the best performance is achieved in the narrower band from 80-150 kHz which is of interest in typical experiments. The RF frequency can be run with open-loop control, following a pre-programmed analog waveform, or phase-locked to track a plasma fluctuation diagnostic signal in real time with programmable phase delay; the amplitude control is always open-loop. The control waveforms and phase delay are programmed remotely. These tools have enabled first-of-a-kind measurements of the tokamak edge plasma system response in the frequency range and at the wave number at which coherent fluctuations regulate heat and particle transport through the plasma boundary.

Index Terms—Wide frequency range, matching network, RF, fusion, plasma, dynamic tuning, PLD.

I. INTRODUCTION

SHORT-wavelength, electromagnetic fluctuations in the tokamak edge region regulate heat and particle transport through the plasma boundary [11–5]. This transport, in turn, plays a critical role in determining the overall performance of a tokamak plasma, with optimal configurations providing a transport channel to exhaust impurities while still achieving good energy confinement (low heat transport). These observations motivated the creation of a novel antenna structure for the Alcator C-Mod tokamak. Named the “Shoelace” antenna for the criss-crossing pattern of its single winding, the antenna was built to reproduce magnetic perturbations at the same perpendicular wave number, \( k_\perp \), as measured for intrinsic edge plasma fluctuations, particularly the Quasi-Coherent Mode (QCM, [6, 7]) and the Weakly Coherent Mode (WCM, [5]). A picture of the antenna installed in the Alcator C-Mod tokamak is shown in Figure 1.

The development of the antenna and associated power system builds upon experience obtained under Alcator C-Mod’s Active MHD program. The original concept of Active MHD experiments in a tokamak was to excite and characterize stable plasma fluctuations as a means to help diagnose the plasma state, as well as to gain physical insight about the modes, themselves [8]. Pioneering work on the Joint European Torus (JET) used saddle coils to try to excite toroidal Alfvén eigenmodes (TAE’s, hundreds kHz) [9–11], as well as to drive these modes non-linearly by scanning the beat frequency between two ion cyclotron range of frequency (ICRF, tens MHz) auxiliary heating antennas [12]. On the Alcator C-Mod tokamak, a pair of antennas [13–15], together with a switched mode power system [16, 17], were built and installed to explore the same type of plasma modes. Non-linear excitation using ICRF antennas was also explored using amplitude modulation of a single antenna.

The present work departs from these previous studies in several ways. Firstly, the focus of the experiment is on high-wave-number edge plasma fluctuations, rather than low- or intermediate-wave-number core modes. Distinctly different physics governs the behavior of the two classes of modes. Moreover, to the authors’ knowledge, there have been no previous attempts to excite these particular edge fluctuations inductively on a tokamak. Experiments examining active stabilization and destabilization of similar edge modes instead employed Langmuir probes in prolonged, direct contact with the plasma [18–20], which is not feasible for the higher temperature plasmas achieved on Alcator C-Mod. Promising work on linear devices has achieved open-loop control, using both electrostatic and inductive means, of the same class of plasma fluctuation as targeted by the Shoelace antenna [21–23].

Secondly, the power systems built for the Shoelace and Active MHD programs differ in approach. The Active MHD experiments employed a pair of custom switching RF sources able to drive a near-short in the frequency range from \( \sim 150 – 1000 \) kHz, as well as a matching network to cancel the antenna inductance, tunable to different frequencies through mechanical switching of a small number of discrete capacitors. The power system built for the present work uses commercial 50-Ω sources to drive the single antenna in the band from 50 to 300 kHz. This places greater demands on the matching network, which must limit reflected power to the sources. Finally, a phase lock system also allows tracking an intrinsic
plasma oscillation signal in real-time, a capability unavailable in the previous Active MHD experiments.

The RF power system built to satisfy these constraints is the primary focus of this paper. In Sections [III] and [IV] the design and construction of the matching network, source, and phase lock systems are described in turn, while Section [V] gives an overview of the matching network calibration procedure. Then, in Section [VI] the overall performance is discussed, and finally, in Section [VII] the work is summarized, with goals listed for future work.

II. MATCHING NETWORK

A. Design

The low-frequency matching problem with inductive and resistive load is solved by the L-match configuration [24 Chap. 10]. This network makes use of two reactive elements, chosen to map the two components of an arbitrary complex load impedance to a particular source resistance.

There are eight configurations of the L network, encompassing the combinations of the series leg being on the source- or load-sides of the parallel leg, and whether each leg is capacitive or inductive [24 Chap. 10]. Since the load presented by the antenna is mostly inductive, the L networks which provide a match over the largest frequency range are those with purely capacitive elements.

Figure 2 illustrates the L network implemented in the present work. This is combined with a two-input, single-output RF combiner built from two transformer cores, which completes the impedance match. The desired values of the capacitors, $C_s$ and $C_p$, are calculated from the constraint that the look-in impedance, $Z_{LI}$, should match the transmission line characteristic impedance, $Z_0$, divided by the square of the transformer ratio, $T$, so that $\mathcal{R}\{Z_{LI}\} = Z_0/T^2$ and $\mathcal{I}\{Z_{LI}\} = 0$. Representing the load by $Z_L = R + jX$, with $X = \omega L$, and denoting the parallel combination of $C_p$ and the load as $Z_1 = \left[ j\omega C_p + (R + jX)^{-1} \right]^{-1}$, the matching constraints are

$$\mathcal{R}\{Z_{LI}\} = \frac{Z_0}{T^2} = \mathcal{R}\left\{ \frac{1}{j\omega C_s} + Z_1 \right\} = \mathcal{R}\{Z_1\}$$

$$= \frac{R}{(1 - \omega X C_p)^2 + (\omega R C_p)^2}$$

$$\Rightarrow C_p = \frac{X \pm \sqrt{(R^2 + X^2) \frac{R^2 T^2}{Z_0^2} - R^2}}{\omega (R^2 + X^2)}$$

$$\mathcal{I}\{Z_{LI}\} = 0 = \mathcal{I}\left\{ \frac{1}{j\omega C_s} + Z_1 \right\}$$

$$\Rightarrow C_s = \frac{1}{\omega^2 \mathcal{I}\{Z_1\}} = \frac{(1 - \omega X C_p)^2 + (\omega R C_p)^2}{\omega [X (1 - \omega X C_p) - \omega^2 R^2 C_p]}$$

$$= \frac{Z_0}{\omega^2 T^2 (1 - \omega X C_p)} - 1$$

In this application, the negative root for $C_p$ is selected; otherwise, $C_s$ must be replaced with an inductor.

An RF combiner accommodates multiple RF sources, as shown in Figure 3 and also further adjusts the look-in impedance of the L network to the line-matched value. Assuming the combiner consists of $M = 2$ identical transformers, each with individual winding ratio, $N_p/N_s = T_i = 4$, the effective winding ratio for the combiner is $T = \sqrt{MT_i} \approx 5.7$, such that the effective transformer ratio in Eq. 1 is $T^2 = MT_i^2 = 32$. A more precise matching condition may be further rendered by characterizing the transformers using a T equivalent network. Doing so, and neglecting the magnetizing and core losses, requires replacing $Z_0$ with $Z_0 - Z_{\text{short}}$, where $Z_{\text{short}}$ is the short-circuit impedance presented by the transformer looking in from the high-voltage (source) side. The required action of the L-match network is then to map the

![Fig. 2. Schematic of the L-match network used in this study. L and R are the antenna load inductance and resistance; $C_s$ and $C_p$ are the series and parallel capacitors; $Z_{LI}$ is the impedance looking into the matching network. For simplicity, the transformer/combiner is not shown.](image-url)
antenna load impedance to $Z_{LI} = (Z_0 - Z_{short})/(MT_L^2)$.

1) Choosing Discrete Capacitor Levels: Because the power system must operate over a wide frequency range from 50-300 kHz, and the $Q$ of the system is fairly high ($\omega L/R \sim 5 - 15$), the matching network capacitors must be variable, and set dynamically according to the RF frequency. To achieve this, a large number of discrete capacitor levels are arranged in parallel and switched into the network as needed to match the impedance at a particular frequency. In particular, 81 discrete capacitor levels for each of the series and parallel legs of the network were chosen in the ranges, $3.3 \leq C_p \leq 694$ nF and $48.5 \leq C_s \leq 1385$ nF, according to a power law, with $C_{s,n+1} \approx 1.043C_{s,n}$ and $C_{p,n+1} = 1.069C_{p,n}$. This scheme provides current sharing between the discretized capacitors, so that no one capacitor channel carries more than $\sim 5\%$ of the total current running through the series or parallel pathway. This selection, in combination with the current limitation through the solid state switching, ultimately provides an upper bound on current that may be coupled to the antenna load; the design target for this bound was 200 A, based on heating considerations of the antenna winding.

An alternative distribution was considered which would have attempted to place the various resonant characteristics of each switching combination so as to minimize the reflection coefficient across the entire frequency band (see Figure 7). In practice, however, the antenna impedance is not static enough to merit such a high degree of optimization, and a more flexible capacitor distribution also allows more independent development of the matching network from other components in the system.

B. Implementation

The conceptual design described above is realized by the system outlined schematically in Figure 3. The completed system comprises 80 dynamically-switched, as well as one static “base load,” discrete capacitance levels for each of the series and parallel matching legs. Solid state switching is used to add or subtract the appropriate capacitance to achieve a match. The static series and parallel capacitors are always in place. The capacitors are laid out on custom circuit boards; a photograph of one such board is reproduced in Figure 5. The boards are 6U high (Eurocard dimension, 23.335 cm) and 22 cm deep. Each board provides four series and four parallel capacitors, as well as associated solid-state switching and control circuitry, so that a total of 20 boards provide all of the dynamically-switched capacitor channels. The static base load makes use of five additional boards, as described below.

Robust capacitors are needed to survive the high voltages (up to 1 kV) and currents (200 A maximum design target) produced within the power system. The components selected to meet these requirements were AVX Hi-Q, AVX High-Voltage (HV), and Kemet C-series multilayer ceramic (MLC) capacitors, all of which are ceramic, C0G (lowest thermal coefficient), low-equivalent-series-resistance capacitors. The inventory of capacitance values was 0.22, 0.39, 0.47, 0.68, 1.0, 4.7, and 8.2 nF; in all, 839 capacitors were used. Breakdown voltages ranged between 1 kV for the larger capacitance components and 3 kV for the smallest. All capacitors fit the surface mount 2225-case (0.22-in×0.25-in, 5763 metric) footprint.

Multiple capacitors are combined in parallel to produce the capacitance increment for a particular level. In the dynamic-switching boards, the capacitors were allotted in each discrete level according to the constraints that (a) no capacitor could account for more than 1/3 the total capacitance in the level while (b) no more than 11 capacitors were allowed in a level (leaving one spare solder pad in the channel), and (c) target capacitance values were achieved to within half the smallest allowed capacitor for each level.

Each capacitor switch utilizes two ST Microelectronics STW13NK100Z power MOSFETs, connected drain-to-source such that the MOSFET body diodes do not short out the

\[ \text{Due to inventory constraints and base load requirements, 0.22 and 0.39 nF capacitors were reserved for the smallest capacitance levels.} \]
switch. These transistors were selected in part because of their high voltage rating (1 kV drain-to-source). The antenna sinks current from both the series and parallel branches; this, in conjunction with the 5% increment in the capacitor distributions, means that the 13 A rating of these FET’s comfortably exceeds the 200 A antenna current design target. The 0.56-Ω drain-to-source on resistance is adequately low given the high-degree of current sharing, while intrinsic parasitic capacitance is low when the devices are fully in saturation \[25\], though not negligible (see Section IV).

Figure 4 shows the MOSFET driver circuit. A 500 kHz logic signal is generated on a compact programmable logic device (CPLD), divided down from a 4 MHz clock. This runs two drivers, exciting 1:5 transformers in a push-pull configuration. The stepped-up voltage undergoes full-wave rectification with a 100-µs RC filter and drives the MOSFET across the gate and source terminals. To turn the switch on and off, the 500 kHz control signal is amplitude-modulated by the CPLD, either at the full-amplitude (“on”) state or the zero-voltage (“off”) state. The driver circuit produces gate-source voltages of \(\sim 14 \text{ V}\) from the 5 V square wave input, fully turning on the MOSFET switches in sub-millisecond transition times.

Since the source terminals may float at RF voltages, the transformers in the driver circuit must also provide isolation between the logic circuitry and the RF power. The 1500 \(V_{rms}\) isolation afforded by the S5499-DL transformers selected for this role surpasses the design requirement \[26\].

Base level series and parallel channels replace dynamic switches with wire shorts to provide a static, minimum capacitance for each pathway. When all switched levels are disengaged, the base capacitors must be able to carry the entirety of current running through the matching network, so their construction needs to be very robust. To satisfy this constraint, the current load is divided across many capacitors and board channels. The ideal, design-value base capacitances were distributed roughly uniformly across five circuit boards, using a total of 15 components for the \(C_p\) branch and 49 for the \(C_s\). However, leakage capacitance through the boards and MOSFETs provides some base level capacitance, albeit amplitude-dependent. The base boards have no switching capability; in order to adjust for the leakage capacitance, only three of the original five boards were used in the final matching network, trimming the total static contribution to \(C_p\) and \(C_s\).

C. Diagnostics

High-voltage, high-current, low-loss probes are required both to monitor the power system performance and record antenna voltage and (crucially) current waveforms for referencing against plasma diagnostics. Moreover, the diagnostics must be compatible with the \(\sim 10 \text{ kΩ}\) input impedance presented by the D-tAcq ACQ216 digitizers available for data collection on Alcator C-Mod.

To address these needs, several current/voltage (I/V) probe units were constructed. These employ custom-built capacitive voltage dividers, nominally providing division of 200:1, as well as Pearson Model 101 current monitors with peak amplitude, 200 A, and an operational band from 0.25 Hz to 4 MHz \[27\]. The lower frequency bound of the voltage divider is determined by the resonance between the low-voltage leg capacitance and the magnetizing inductance of the isolation transformer, while the upper frequency bound results from the need to keep the impedance of the probe much larger than that of the antenna, which is in parallel. The circuit parameters were selected such that the phase shift in the voltage measurement resulting from the low-frequency resonance would be \(\leq 5°\) at 50 kHz, while the reflection coefficient, \(\Gamma\), would be increased by no more than 5% due to the modification of the load impedance by the voltage divider. Actual performance meets and exceeds these constraints.

The voltage and current probes are housed in Compac SRF RF-shielded boxes. These provide insulation against the noisy environment of the Alcator C-Mod experimental area, while also protecting against possible leakage of high-frequency signals which can be accidentally coupled from other antenna systems within the tokamak.

The capacitive voltage dividers are calibrated against 100:1 oscilloscope probes, using a high-input-impedance (\(\sim 1 \text{ MΩ}\)) digital oscilloscope to record the waveforms. The oscilloscope probes, themselves, were not suitable for typical operation because they require higher input impedance than that presented by the digitizers available during experiments (\(\sim 10 \text{ kΩ}\)). The Pearson current monitor’s 100 A:1 V factory calibration is generally acceptable for interpreting current data. In fact, long cable lengths (\(\sim 15 \text{ m}\)) result in a phase drift in the current measurement from 1° to 5°, increasing with frequency across the system’s operational band. Since we are typically interested in \(\sim 180°\) phase shifts in transfer functions between the antenna current and plasma diagnostic signals, and over a fraction of the whole band, these phase errors are negligible in interpreting physics results. However, because the antenna resistance is so much smaller than its reactance, a careful
accounting of this effect is required to extract the antenna impedance from current and voltage waveforms.

The antenna, itself, can also be used as a $k_{\perp}$-specific receiver to diagnose plasma fluctuations. In this operational mode, the power system is disengaged, and the voltage induced in the antenna by oscillations in plasma radial flux is coupled to a digitizer channel via an isolation transformer and a discretely-variable voltage divider. The antenna also picks up the 500 kHz beat frequency between two ICRF heating antennas, which operate at 78 and 78.5 MHz. The pickup is of sufficient amplitude that it can saturate the digitizer channel if left unmitigated. As such, it is suppressed with a simple LC notch filter. The digitized voltage is calibrated based on a characterization of the isolation transformer, divider, and filter to yield the voltage induced across the antenna.

D. Matching Network Control System

The matching network adjusts its tuning state according to the Sync signal, an output from the RF Generator Board described in Section III and Figures 5 and 6. The period of this signal is measured in the Master Control Board (MCB), and this quantity is then mapped to two independent tuning numbers – one each for the series and parallel capacitor branches – using a lookup table. These two states are communicated to all dynamically-switched capacitor boards, which adjust their states accordingly to produce the desired total series and parallel capacitances.

The logic used to control the dynamic boards was programmed onto Altera EPM7128SLC84-15 CPLD’s; a version of the Verilog code used for programming is available from GitHub [28]. A different CPLD, the Altera EPM2210F256C5, was used to accommodate the more complicated logic of the Master Control Board. The Verilog code used to program this device is also publicly available [29].

In the following, the logic flow is described briefly (see also Figure 5). The MCB’s CPLD implements a period counter to interrogate the Sync signal. The period is averaged over $M = 25$ Sync cycles at an $f_{\text{clk}} = 8$ MHz clock rate resulting in a worst-case quantization error in frequency resolution of $\approx 2f_{\text{sync,max}}^2/(Mf_{\text{clk}}) = 900$ Hz. The minimum response time is $M/f_{\text{sync, min}} = 500 \mu s$, which occurs when the drive frequency is at the 50 kHz lower bound of the operational band.

Separate series and parallel capacitor lookup tables are programmed onto the MCB CPLD. The two lookup tables specify the bounds of each particular tuning state in terms of the Sync period, measured in clock counts. The period, rather than the frequency, is used in the bounds to avoid a division operation on the CPLD.

The state of the system is encoded in two independent, seven-bit binary numbers, one for the series capacitors and one for the parallel. These are broadcast from the MCB on a custom backplane feeding all capacitor boards, together with two separate enable bits to indicate changes in either the series or parallel states. These two states, together with the measured signal period, are also encoded in three serial bit streams which update on every change; these diagnostic outputs are recorded on digitizers during typical operation in order to monitor proper functioning of the matching network.

The capacitor boards’ responses to the global tuning state are determined from each board’s six-bit address number, which is parsed from a set of reconfigurable dip switches. State changes on the capacitor boards are triggered by the enable bits, with the series and parallel states controlled independently and in parallel.

III. Source and Control System

RF power is provided by two T&C AG1010 50-Ω, Class B amplifiers which provide 600 W continuous power and 1 kW pulsed power in the band from 20 kHz to 1 MHz [31]. Typical operation on the Shoelace antenna system is limited to 1 s with several minutes between pulses, so it is the 1 kW power limit that is relevant. The output from both amplifiers is combined in the matching network.

Figure 6 illustrates schematically the basic construction of the function generator which feeds the RF amplifiers, as well as its control system. The amplitude of the function generator output follows an open-loop program with a single analog control signal, while the frequency and phase follow either open- or closed-loop (feedback) control. In the open-loop case, an analog input provides a control signal for the frequency of the output. In the closed-loop case, the function generator output is synchronized, with variable phase delay, to a real-time plasma density fluctuation signal via a phase-locked loop (PLL), as described in Section. The selection between the two frequency control pathways is carried out.
automatically by an enable bit which is set when the cross-power between the PLL output and plasma fluctuation signals is sufficiently high, with an auxiliary manual remote switch to disable phase locking. All analog control signals are generated on a remotely-programmable BiRa Systems, Inc., Model H910 function generator, which is part of Alcator C-Mod’s CAMAC-based data acquisition system.

Besides the two RF power outputs, the source also provides an additional TTL square wave – the “Sync” signal – that is generated with and synchronized to the sinusoidal input to the RF amplifiers. It is this signal that provides control of the matching network tuning state.

IV. CALIBRATION

The problem of assigning frequency ranges to discrete capacitance levels is manifested in Figure 7. Here, the fraction of power transmitted to the matching network is shown against frequency for many different combinations of discrete series and parallel capacitance levels. At a given frequency, a good calibration picks a capacitance configuration with a resonant curve whose power transmission is near 100%. However, it is infeasible to characterize all of the available resonance curves, not only due to the large number of such curves and their variability with plasma conditions and power level, but also because mapping out the full curve at high power is not possible, since the RF sources will trip as reflected power increases off-resonance. Instead, calibrating the lookup tables requires careful characterization of the antenna load, the effective series and parallel capacitance, and the transformer, resolved across the entire frequency range of interest. These quantities are then integrated into the idealized models of Section II-A to synthesize initial tables, which are then optimized manually with additional trials.

The most reliable technique for these operations is to excite the system under RF power, sweeping the frequency across the operational band and digitizing voltage and current waveforms across the components on an oscilloscope. The impedance over each component is then computed by processing the recorded signals; this frequency-dependent impedance data can be used directly, or parameterized in circuit models. Figure 8 compares the antenna impedance obtained with this procedure (at room temperature) against estimates provided by simple analytical models for resistance 32 and inductance. The transmission lines and vacuum feedthrough connecting the power system to the antenna contribute a non-negligible amount of impedance, and are included in load characterization.

In fact, the antenna impedance is not static; it varies with plasma conditions as heating from the plasma reduces the winding conductivity, and as antenna/plasma coupling changes. Fortunately, it is not necessary to re-tune the matching network between plasma discharges. Instead, characterizing the antenna impedance once under realistic conditions is sufficient, barring major faults developing in the antenna, and adequate results can also be obtained by assuming an unchanged inductance from the value with no plasma present, since antenna/plasma coupling has only a minimal effect, and a resistance increase commensurate with an estimated temperature rise of the antenna winding (in this case, $R$ increases by approximately 50% during a plasma discharge).

With regard to characterizing the matching network’s discrete capacitor levels, there are two non-ideal effects to consider. The first is the inclusion of stray capacitance between traces on the circuit boards. In each board, the static stray series and parallel capacitances are $\Delta C_s \approx 16$ and $\Delta C_p \approx 18$ pF. Totaling over the 20 dynamically-switched boards and five base level boards gives total static stray capacitances of $\Delta C_s \approx 400$ and $\Delta C_p \approx 450$ pF, both negligible values.

Leakage also occurs across the MOSFET switches. This is a nonlinear effect, as the drain-to-source parasitic capacitance drops off rapidly with increasing drain-to-source voltage. Since the range of parasitic capacitance in these transistors spans hundreds to thousands of pF per device 25, this effect cannot be neglected, particularly for the higher frequencies, which require smaller capacitance values. This causes a complication in calibrating lookup tables. Initially, low power must be used to characterize the effective capacitances and load; otherwise, an abundance of reflected power will result in a trip at the RF sources. However, at higher powers, the nonlinear capacitance changes, detuning the system and again leading to high reflected power and trips at the source. As such, an iterative procedure is required, stepping the power up gradually while retuning the system after every step. In practice, three to four steps are adequate for the full frequency range.

Figure 9 shows one of the finished pairs of lookup tables used in the first round of antenna experiments.
Fig. 8. Antenna load, together with predicted value from model. The measured load characteristic is shown with and without the contribution from 4.3 m of transmission line and one vacuum feedthrough.

Fig. 9. Lookup tables, mapping frequencies to series and parallel state numbers across the antenna operational band.

V. Phase Locking to Real Time Fluctuation Measurement

The desire to explore feedback stabilization or destabilization of plasma oscillations motivated the development of a phase lock system. Figure 6 provides a simplified schematic of this system’s operation. A phase-locked loop generates a square wave that follows a real-time analog output from the Phase Contrast Imaging (PCI) diagnostic [33], which resolves line-integrated plasma density fluctuations. When the cross-power between the diagnostic input and the locked square wave passes a threshold level, an enable bit is set to indicate a successful lock. If a second remote switch is also set, frequency control for the power system is changed from a pre-programmed evolution to the live lock.

The phase relationship between the locked square wave and the plasma signal can be adjusted in two ways: either via hardware switches to produce a 0, 90°, 180°, or 270° lag, or a separate phase delay circuit board. In the separate phase delay unit, the square wave is sampled at 16 MHz and stored in a cache. The input signal period (typically $\sim 6 - 20 \mu s$) is also measured by counting the number of clock cycles, $N_{clk}$, between rising edges - an $M = 1$ period counter. From this measurement, a delay is calculated. The output of the phase delay board is the state of the input square wave delayed by this amount, as retrieved from the cache. An Altera EPM2210F256C5 CPLD implements the required logic. The Verilog code used to program the CPLD has been made publicly available [34].

The challenges of locking to a plasma mode in this way are suggested by Figure 12, which shows actual performance of the locking system. The rapid evolution of the QCM – the fluctuation to which the power system must lock – is apparent from the PCI spectrogram in the top of the figure, where the time-evolving peak in the short-time spectra is due to the QCM. Building a phase-locked loop to track such a variable signal requires a careful balance between stability and response time. Moreover, rapid state changes in the matching network are required to provide a good impedance match over the duration of the antenna pulse. Indeed, the initial response time of the matching network state changes – 1 ms – was found to be too slow, such that the MCB clock rate was doubled to 8 MHz, and the period count halved to $M = 25$, in order to reduce the upper bound on the response time to a faster 500 $\mu s$.

VI. Performance

Figure 10 summarizes the capability demonstrated by the Shoelace power system. The top frame shows the power fraction transmitted to the matching network across the entire frequency band during an actual plasma discharge, the middle frame shows the current amplitude in the antenna for the same discharge, while the bottom frame shows the total power output from the RF sources. At least 85% of source power reaches the antenna across the entire frequency band from 50-300 kHz, with better efficiency in the lower band from 50-150 kHz, which was of primary interest in experiments. In a number of experiments, currents in excess of 80 A were achieved routinely in the lower frequency band, and the system operated reliably for hundreds of pulses. These performance characteristics exceed the design goals set forth at the project’s inception.

Figure 11 shows results from a typical antenna experiment [35]. The top pane is a spectrogram of a PCI signal – it shows the evolution of the spectral content in line-averaged plasma density fluctuations. The distinctive, somewhat broad feature setting in at around 0.98 s and spinning down in frequency indicates the presence of a quasi-coherent mode. The antenna-driven perturbation is visible as a triangle wave following the pre-programmed frequency waveform; it appears in the
Fig. 10. Matching network operational performance from an Alcator C-Mod discharge. The top curve shows the fraction of power entering the matching network. The remaining curves show the current and power entering the antenna. The power demand at higher frequencies is reduced in the system programming to avoid tripping the RF amplifiers.

spectrogram just prior to the onset of the QCM. The antenna frequency is compared with the frequency of maximum spectral content in the middle pane of Figure 11 and the bottom pane shows the current driven in the antenna. The current varies between 68 and 83 A in this discharge; it tracks the antenna frequency, falling slightly with increasing frequency as the skin effect raises the antenna resistance. A slow droop in current over the duration of the pulse follows from the increase in the antenna temperature, and hence resistance.

Figure 12 shows the operation of the antenna with the phase-locked loop engaged. The format of the data is as in Figure 11. Initially, the source frequency remains at a stationary, pre-programmed value. At 1.128 s, the lock enable bit goes high, indicating that the cross power between the locked waveform and the real-time plasma signal from the PCI diagnostic has crossed a threshold value. Subsequently, control passes to the phase lock system, which successfully tracks the QCM frequency until the mode coherence drops around 1.4 s. However, more careful analysis shows that the phase lag between the antenna current and plasma signal is not fixed; this is lost when the function generator tries to lock to the output from the phase delay board. The square-wave envelope of the antenna current shows the amplitude modulation employed to help discern the antenna’s effect on the fluctuation signal. Several very short trips are visible, but the nominal current level stays constant despite the rapidly-varying frequency.

It should be noted that the matching network calibration is adaptable. At one point in the experimental campaign, a fault developed in the antenna: half of the windings were shorted out. This required a new calibration for the capacitor look-up table. Despite operating at roughly half the normal impedance, the system was still able to drive up to ~ 80 A in the antenna, albeit in a reduced band from 80 to 150 kHz.

Fig. 11. Typical operation of antenna system with open-loop frequency program. (Top) Spectrogram of plasma density fluctuation. (Middle) Antenna frequency and plasma mode nominal frequency. (Bottom) Antenna current. Reprinted from Golfinopoulos et al., Phys. Plasmas (2014-submitted).

Fig. 12. Demonstration of phase lock. (Top) Spectrogram of PCI diagnostic; thin dashed black line highlights peak frequency in PCI spectrum, thick black line indicates transition from open- to closed-loop control. (Middle) PCI peak frequency \( f_{\text{max,pci}} \) (blue) overplotted with antenna frequency \( f_{\text{ant}} \) (green). Cross-power between phase-locked-loop-generated signal and PCI diagnostic signal surmounts logic threshold at 1.128 s, so that control is automatically switched from a pre-programmed (open-loop) constant frequency to the locked signal, whereupon \( f_{\text{ant}} \) closely tracks \( f_{\text{peak}} \). (Bottom) Antenna current is intentionally amplitude-modulated at 9.5 Hz.
The antenna system produced significant results from a scientific perspective, providing a unique capability to drive short-wavelength plasma fluctuations in the edge of a tokamak plasma using an inductively-coupled antenna. A more detailed account of the physics elucidated from these experiments will be published elsewhere [85].

VII. Conclusion

A new antenna – the “Shoelace” antenna – and associated RF power system have been built at the Alcator C-Mod tokamak to explore the excitation of edge plasma fluctuations. These fluctuations play a critical role in determining transport through the plasma boundary; the experiments enabled by this system give a unique insight into their behavior, and also test the possibility of building structures to actively control boundary transport through this pathway.

The RF power system built for the Shoelace antenna provides a match over a broad band from 50 – 300 kHz, with a very low level of reflected power, and scalability of power using multiple commercial RF sources. A sophisticated control system allows fine adjustment of amplitude and frequency according to remotely-programmable waveforms, and also has the capability of locking to a plasma oscillation in real time. These features improve upon the capabilities of a similar power system used at Alcator C-Mod in prior Active MHD experiments.

The system may be refined by achieving a tighter phase lock on plasma modes, such that feedback stabilization or destabilization of intrinsic modes may be more readily studied. Moreover, an expansion of the available RF source power is desired to increase the amplitude of antenna-induced plasma perturbations. It is also of interest to broaden the frequency range of the system still further, adapt to changing loads in real-time, and generally make the system more flexible and simpler to adapt to multiple applications.

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